

Textbooks in Telecommunication Engineering

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Radio Receivers for Systems of Fixed and Mobile Communications

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Series Editor

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Dr. Tarek S. El-Bawab, who spearheaded the movement to gain accreditation for the telecommunications major is the series editor for Textbooks in Telecommunications. Please contact him at telbawab@ieee.org if you have interest in contributing to this series.

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Dedicated to the Moscow Technical University of Communications and Informatics (MTUCI) and the Moscow Power Engineering Institute (Technical University) within the walls of which the great scientists of their time in the field of radio worked, Kotelnikov V.A., Kharkevich A.A., Aizenberg G.Z., Kataev S.I., Siforov S.I., Evtyanov S.I., Goron I.E.; a huge number of graduates who continued the best traditions of Soviet/Russian scientists – Fomin N.N., Zubarev Yu B., Shahgildyn V.V., Shinakov Yu S., Varakin L.E., Kaganov V.I., Bogache V.M., Kapranov M.V., Kuleshov M.V., Utkin G.I.; and many others, for whom alma mater became a school of life and creativity.

Preface

This book, *Radio Receivers for Fixed and Mobile Communication Systems*, introduces the reader to the principles of building receivers of basic types of modulation and features of their architecture when using digital modulation methods, multiple access, and changing the structure of a radio engineering system. It is shown how the design of subscriber terminals changes when organizing simultaneous communication in opposite directions and mobility of at least one of the subscribers, when moving from the classical scheme of information transfer in fixed networks to mobile communication systems, handover of a mobile subscriber when switching between cells of one system or between different systems. The use of a cellular networking structure allows increasing the number of simultaneously served subscribers (increasing the system throughput) compared to systems served by a single BS (a single transmitter), while at the same time complicating the interaction algorithm between the BS and AT. The cellular structure of a mobile wireless communication system improves the efficiency of using the bandwidth allocated to the system by reusing operating frequencies.

This book is intended for students of classical and technical universities and colleges. It contains a comparative analysis of the principles of building radio receivers for analog fixed communication systems using basic types of radio signal modulation, as well as the features of their development and improvement in the transition to digital methods for forming and processing radio signals to use the frequency spectrum allocated for radio systems. This book has been prepared on the basis of many years of experience of the authors teaching at various Russian technical universities, and many years of experience of the authors in short courses for process engineers and researchers.

Radio receivers BS and AT, united by a single radio interface, can be developed using both identical and different types of block diagrams, with the active introduction of both circuit solutions and software methods for implementing individual nodes (mixers, demodulators, detectors, etc.). Despite the general trend to reduce the analog part in the structure of receivers, their role as part of the radio system is becoming increasingly important. This is due to the fact that in addition to signal transmission/receive between AT and BS within the network or inter-network roaming, the receiver is assigned the function of a device that changes the parameters of the terrestrial radio access network to provide a given data transfer rate. This is achieved by choosing the type of modulation and expanding the available bandwidth using static or dynamic network planning methods.

The success of interaction between Wireless Personal Network (WPAN), Wireless Local Area Network (WLAN), Wireless Metropolitan Network (WMAN), Wireless Wide Area Network (WWAN) with various interfaces is ensured by the inclusion of parallel structures in the analog and digital parts of the receiver architecture, adapted to signal processing of the corresponding standards and allowing operation in different frequency ranges.

The development of the principles laid down in SR is the creation of Cognitive Radio (CR), the working band of which is formed by dynamic monitoring of all possible areas of the available radio frequency spectrum, the allocation of currently free frequency areas in the systems, and the formation of an aggregated channel, including free frequency bands of various operators and systems. Thus, the receiver, having gone through all the stages of improving the architecture, becomes not just a terminal device of a radio channel, but an element of a cognitive network structure that unites several systems, different standards, and frequency ranges.

The Fourth Industrial Revolution and the Role of Infocommunication Systems

Radio receivers are necessary components of radio systems with different purposes: radio communication, TV and radio broadcasting, navigation, detection as well as the important element of functioning communicational Internet network (Internet of People), and the promising network of the future generation NGN (Next Generation Network), which realizes the Internet of Things principle. This direction of infocommunications' development is the composite part of large-scale

variations in human society, which is known as the “fourth industrial revolution” [1, 17, 18], which can fundamentally transform society. A combination of known and inceptive technologies, for instance, the artificial intelligence, robotization, three-dimensional printing, nanotechnologies, quantum calculations, and Internet of Things, will provide the mobile access to billions of people to the area of information processing and storing in different fields of human activities.

Under the Internet of Things we understand the totality of different devices, sensors, and instruments, which interact between themselves, including the human being integrated into the network with the help of available *communication channels* with various protocols. As a global network, Internet interconnects with the forming network through Internet protocol (IP). Recommendation ITU-T Rec. Y.2060 [2] defines IoT as the *global infrastructure of the informational society* providing promising services of interacting physical and virtual things, which can be identified and integrated into the network by compatible information and communication technology. The problem of things compatibility, including the data acquisition, processing, and storage, is entrusted with the equipment part (device) having the communication possibilities.

Organization of connections anticipates the utilization of the IoT network of different technologies: global networks, local networks, wireless sensor networks, ultra-wideband networks, and personal networks. In the structure of the Internet of Things network, the communication technologies are used on the lower level providing physical or virtual connection of the “smart objects,” which are integrated with sensors, and sensor connection with an aggregator (a gateway).

Sensors having small memory provide information acquisition, storage, and processing in real time. Owing to small physical sizes, they can be embedded directly into objects. Connection of sensors to control devices occurs through gateways (sensor aggregators) of the Local Area Network (LAN). It can be built either as a wired network or as Wireless Local Area Network (WLAN), including Wi-Fi, Ethernet, and Bluetooth. Sensors, which have low power consumption and small data transmission rate, form the wireless sensor networks (WSN). Utilization of the personal network WPAN (Wireless Personal Area Network) for communication with servers/applications can include Zigbee, BlueTooth, and DECT. For transmission of large data arrays to short distances with high rate, the ultra-wideband networks are used. Independently on the type of the local networks, for integration with local networks and for communication with the control server or the global network, they must include gateways using geographically distributed sensor networks. In the simplest case, for these purposes, computers are used, which have access point to the Internet network. This makes such a structure bulky depending on the power supply and the access point to Internet.

If the gateway contains the receiver-transmitter (the transceiver) connected with the sensor network, and in the presence of the micro-controller, it is compatible with the wired network, usually Ethernet, then this aggregator will have lower cost. The construction of the autonomous gateway, which is at the same time the access point of Internet, will require two transceivers, one of which is connected to the sensor network, while the second with the global wireless network, the operation area in which this gateway is situated. For the application of global wireless networks WWAN (Wireless World Area Network), such as GSM, GPRS, WiMAX, UMTS, LTE, LTE-A, and the 5G standard, preference is given to GSM and WiMAX networks, which have low power consumption.

Communication systems of various levels are important (or even the main) components of network formation in the IoT structure. In this book, we consider radio networks (network of wireless communication), in which the free space (the “ether”) is used as the transmission medium. Properties of the radio channel define the transmitted signal parameters, reception distance, types of modulation, and receiver sensitivity.

Signal Propagation Features in the Free Space and Interferences in the Radio Channel

In systems of fixed communication, which use the basic types of modulation, the communication distance is defined not only by the receiver sensitivity and the interference level. The distance of signal propagation, acting in the radio channel, is defined not by noise immunity of used modulation, but by features of the signal propagation. If in ranges up to 10 km (Low Frequency, Middle Frequency) electromagnetic waves propagate along the Earth surface, forming the surface currents and rounding of its irregularities, then it leads to essential power losses. Shorter waves (Middle Frequency, High Frequency, Very High Frequency) propagate on a large distance, which is caused by reflections of radio waves in an ionosphere. In this case, radio waves propagate toward the ionosphere and reflect in it from the layers on different heights, leading to signal appearance in the reception point, which are achieved by different ways (different phases) causing its partial compensation [3]. Such signal reception is called “multipath,” and distortions caused by this phenomenon are called the interference (hashing) distortion. In ranges of very high frequencies of 30/300 MHz, ultrahigh frequencies of 300/3000 MHz, and microwave frequencies of 3/30 GHz, which are actively used in the mobile communication systems and in systems of radio access [4], the

radio signal propagates in straight lines. In this case, signals arrived to the receiver input by direct visibility or due to reflections and scattering from the surrounding objects and constructions creating the interference distortions. At some angles of transmitter emission, the signal may penetrate outside the ionosphere limits providing the communication with satellites. This allows organization of several thousands of radio communication channel or hundreds of TV channels providing the information transmission on very long distances. The radio signal in the radio channel with the restricted frequency band is exposed by the interference impact of various natures. The outside radio signals, radio communication, TV signals, radar technology, radio navigation, etc., are the most significant, exerting the noticeable influence on the quality of the receiving signal. Similar impact is exposed by out-of-band emissions of radio transmitting devices. To reduce its influence, the filters are included in the transmitter output, which restricts the emitted frequency band in the transmitter and, at the same time, suppresses the influence of nonlinear modes of active elements in the output stages.

In the frequency region below 25 MHz, the atmosphere interferences and outer space emissions expose the receiving signal quality, which reduces the receiver sensitivity and leads to the necessity to use the retuned input circuits of receivers for narrowing the band of impacting frequencies.

In the urban environment, the industrial interferences (transport, industrial, medical installations) exert the essential impact on the quality of receiving signals and for stationary receivers – the conductive interferences. To reduce such types of interference influences, they should be eliminated from where they arise. Other types of interferences are those that arise inside the radio receiver, and their sources are active receiver components, resistors, and current-conducting circuits. Internal interferences, arising at the first stages of the front-end, define the minimal value of the receiver noise-factor, and hence, its maximal sensitivity. Independently on the receiver structural diagram, the lumped (concentrated) interferences having the narrowband spectrum exert the maximal influence on its quantitative indices. If the input circuit is tuned, the frequency selection and utilization of the quasi-optimal filters in it allow effective fight with such interferences. In tuned input circuits, which are typical for receivers of the mobile communication systems and systems of radio access, we can use the signal pre-distortion in the transmitter, for example, by the function of raised cosine besides mentioned methods, as well as utilization of active components in the preselector with the large dynamic range.

Radio Systems and Networks

The first systems of fixed radio communications operated on the principle “point-to-point,” providing the information transmission between two terminals. Revolutionary developments in vacuum electronics allowed building of the radio broadcasting system, which realizes the principle “point-to-many point,” when the power transmitter provides the message transmission simultaneously to the large number of subscribers. This led to low effectiveness of transmitter power utilization and allocated frequency range at low population density.

Systems of the First Generation (1G) Arrangement of a large number of low-power transmitters in towns at variation of the control algorithm for their power and frequency planning permitted to build the wireless radio communication networks with the type “point-to-point.” In addition, this allowed the subscriber exit into the telephone network of general usage and the relay subscriber transfer (handover) between the base station of the network. The base station checks the channel occupation and renders this channel to subscribers at its dismissal. To reduce the channel mutual influence, which is allocated by the base station (BS), the directional antennas serving the single sector in 120 of the separate cell are installed. The analog mobile systems, Nordic Mobile Telecommunication (NMT), in the frequency range of 450 and 900 MHz operated on the similar principle, which provided the transmission rate of 1.2 kbit/s using the frequency modulation, and the Advanced Mobile Phone System (AMPS) were developed with the transmission rate of 10 kbit/s in the range of 800 MHz.

2G Systems Exploitation of mutually incompatible analogs, which were used in Europe, relatively low transmission rates, and difficulties in equipment unification led to appearance of Global System Mobile (GSM) [5, 6]. Trunking systems [7, 8] was the investigation testing area for construction of digital wireless communication networks, which allowed testing of new effective methods of digital information coding, modulation methods, the access to a channel, the control of transmitter power of mobile stations, and so on. Requirements to interfaces of other interacting systems are formulated in the structure of the multilevel model of the Open System Interconnection (OSI). This allowed the standardization of the physical channel interface between the base station (BS) and the mobile station (MS) by choosing a group of logical channels for transmission

of system information (control, synchronization, access control) and a channel for transmission of subscriber useful information (traffic and control for allocated traffic channel). The group of allocated control channels provides the subscriber identification, its location, and information about the allocated channel [6]. Independently on physical channel characteristics in the transmission medium (BS $\leftrightarrow$ MS), it provided interaction with other systems (Internet, PSTN, ISDN, etc.) using protocols of higher level (transport, network, and so on).

The speech signal transmission, for which the GSM system was developed, is provided by its transformation from the analog signal having the standard bandwidth of $F_h = 4$ kHz to the digital signal with the help of a nonlinear pulse-code modulation (PCM) encoder. The discretization frequency $f_d = 8$ kHz is chosen from the Kotelnikov-Shannon condition [22], forming the 8-bit sequence on its output. The GSM system of wireless communication provides the data transmission in the group channel with maximal rate of 9600 kbit/s and at modems, first of all, in wireless access systems Wi-Fi, WiMAX, and Hiperlan, using the interaction with other systems. The GSM system is the *system with channel switching*, when the channel with width of 200 kHz is allocated for the subscriber on all levels of the OSI model during connection time at interconnection with other systems. The transmitted information in the time domain is formed as packets, whose duration is defined by their destination.

To obtain the operation mode with “missing,” when each of the conversation participators can simultaneously listen and speak (duplex mode), the operation at the physical level is applied. In the GSM system, frequency duplexing is used (FDD, Frequency Duplex Division), when division of reception and transmission channels is provided in frequency, when the mutual antenna of the transceiver is connected to the receiver input or to the transmitter output in time of data transmission (20 ms).

We must distinguish duplexing and the method of channel formation (access to the channel and its division), which can be with frequency division (FDMA, Frequency Division Multiplex Access) and time division (TDMA, Time Division Multiplex Access). In GSM systems, the frequency duplexing is used, when bands of reception and transmission are allocated at simultaneous frequency multiplexing (FDMA) and time multiplexing (TDD, Time Duplex Division). In the frequency band of 200 kHz allocated to the single channel, at the same time, the access to the channel is provided on the same carrier frequency to eight subscribers of TDMA (this is defined by the bit number allocated to the subscriber at the block length of 20 ms), which permits to additionally increase the subscriber number by eight times. Thus, in the GSM system, simultaneously FDMA and TDMA are used at frequency duplexing.

3G Systems Increased subscriber needs in the growth of data transmission rates, which allows obtaining of such services as games in the “on-line” mode, the mobile radio, and TV, leads to appearance of open systems of the third generation (3G) with code channel division. The great variety of communication 2G systems, systems of cellular communication (GSM, IS-95, IS-136, PDS), systems of cordless telephony (DECT, PACS, PHS), systems of data transmission (GPRS, EDGE), systems of personal satellite communication (Iridium, Globalstar) was another reason for the 3G system appearance [9]. The difference of these systems’ equipment, which solved the specific task, essentially complicated their interconnection, which led to the creation of universal system of the radio access Universal Mobile Telecommunication System (UMTS) with a universal standard [10, 11]. Similar tasks were solved at the creation of **CDMA2000** system of wideband access, which is the open standard of IS-95 system. Protocols of the UMTS system include three levels: planes of control, subscribers, and control of the transport network. Their sufficient mutual independence allows protocol changing of ground-based networks of radio access Universal Terrestrial Radio Access Network (UTRAN): physical, channel and network, individually with each other. The subscriber of the UMTS terminal using the higher frequency range (2 GHz) than in the GSM network, nevertheless, should support the operation in both networks.

The main problems solved at development of 3G systems are application of wideband signals, utilization of the spatial-temporal signal processing, application of methods of the reception, and transmission diversity, as well as noise immunity coding. The radio signal broadbandness, transmitted in the system, means the presence of signal spreading procedure in the transmitter, which increases its base B :

$$B = \Delta f_s / \Delta F_s, \quad (1)$$

where Δf_s is the signal spectrum width in the radio front-end and ΔF_s is the spectrum width of the initial (informational) sequence (Base Band). If the spreading on the base band video pulse of the sequence with the single chip duration T is replaced by the chip sequence with the single chip duration $\tau_{ch} = T/N$, where N is a number of elementary pulses (chips) on the video pulse length, then

$$B = \Delta f_s / \Delta F_s = T / \tau_{ch} = N, \quad (2)$$

i.e., the signal base increases by N times.

The signal base growth increases the stability to the interference effect, which is caused by low signal spectral density. Application of spread sequence (the Walsh functions) of high order provides the high resolution in time of radio signal reception using the common frequency band for simultaneous communication of many subscribers.

At the same time, the signal base expansion leads to SNR increase by the same value compared to the narrowband network. Reduction of the radio signal spectral density by B times simplifies the electromagnetic compatibility providing radio access inside the network and between networks. Spreading sequence forms the wideband signal, which has the noise-like spectrum with the spectral density at the level of fluctuation noise of the receiver. This sequence permits the building of the communication systems, which provide a high secrecy of transmission. Application of Walsh sequences, which have the mutual orthogonality on the length of the single symbol of the video sequence T , forms the independent (orthogonal) chip sequences for each information source of the terminal.

In a 3G system that uses a common radio channel to organize communication between BS and AT in the network of one operator, it is necessary that each subscriber has a unique “address” that ensures that only he receives the intended information. For this, a scrambling procedure is used that simultaneously solves two problems: the formation on the BS of a unique code combination known only to one AT of the network, and the change in the structure of the group information signal, which ensures on the length of the radio frame the approximate equality of the number of “zeros” and “ones.”

When using randomization, which is a special case of scrambling, this is solved by adding the mod 2 baseband digital stream with a pseudo-random sequence (PRS). As a result, a new digital sequence is formed, orthogonal to all other sequences used in the terrestrial radio access network.

In the 4G system, the number of zeros and ones are also aligned, and the AT address is indicated in the cell ID, which includes the 168 ID of groups, each of which contains three AT IDs.

Application in each terminal (BS and the subscriber terminal) of two or more antennas is the development of this technology, which increases the noise immunity of the transmitted signal owing to utilization of orthogonal coding of each informational stream in the input/output of the radio channel. The multi-antenna technology (MIMO, Multi Input Multi Output) with signal formation in the radio channel, which uses the single carrier frequency for information transmission, will allow the processing of independent signals, which are formed in each receiving antenna.

4G Systems In the technology of *orthogonal frequency multiplexing and channel division* Orthogonal Frequency Division Multiplex (OFDMA), which has wide distribution in systems of various destinations and standards, not high-speed modulation of the single carrier (as in UMTS) is used, but the modulation of the large number of sub-carriers by the signal, which is formed according to some algorithm from the group informational stream. The whole frequency range allocated to the system is divided into radio blocks, in which each sub-carrier is modulated by slow mutually orthogonal sequences. The group of sub-carrier included in the radio block is integrated in symbols. The achievement of mutual orthogonality of modulated sequences is provided by insertion of protective time intervals, and the orthogonality of the OFDM symbol group by cyclic prefixes.

For effectiveness increase of such a technology, as the means to fight with the intersymbol interference, a number of used sub-carrier should be large enough. So, in radio access Wi-Fi systems of the IEEE 802.11a, g standards, 52 sub-carriers are used, and in **WiMAX** systems of the IEEE 802.16 standard from 200 to 2048 sub-carriers are used depending on the used frequency bandwidth; in ground-based digital TV broadcasting systems **DVB-T** (6 K), 6817 sub-carriers are used.

In 4G network of mobile communication (LTE, Long-Term Evolution), a number of sub-carriers are different, depend on the frequency bandwidth allocated to the system, and are from 73 (1.4 MHz) to 1201 (20 MHz). Each sub-carrier is modulated depending on the channel destination of the quadratic phase modulation (QPSK) or the multilevel amplitude-phase modulation (nQAM). The signal recovering from the received multifrequency signal can be performed both by technical (quadrature demodulators) and by software (inverse Fourier transform). In the last case, the number of Fourier transform points essentially exceeds the number of sub-carriers in the band of effectively used frequencies. Unused sub-carriers are usually situated near boundaries of the allocated band and form the protection interval separating the various systems.

Implementation of new technology led to parameter variation of the *ground-based network of radio access*, Evolved Universal Terrestrial Radio Access Network (E-UTRAN), which is included in the LTE network [12, 13]. First of all, this is concerned to the spectral effectiveness, which achieves 5 bit/s/Hz in downlink (maximal transmission rate of 100 Mbit/s) at the channel bandwidth of 20 MHz, and application of MIMO technology. For such parameters of an interface and the sub-

subscriber terminal mobility up to 500 km/h, the cell radius is 5 km, and the number of active terminals in the cell is not less than 200 at the signal spectrum width of 5 MHz. User data transmission rate in the downlink at two transmitting and two receiving antennas of the subscriber terminal is approximately three to four times more than at application of the high-speed packet mode, High-Speed Downlink Packet Access (HSDPA), in the UMTS system with the identical structure of terminals. Compatibility of the E-UTRAN network with other 3GPP networks of second and third generation provides the inter-network handover and simultaneously utilization of various network channels. The ground-based E-UTRAN network with radio access to different frequency bandwidths allocated to the system (1.4; 3; 5; 10; 15; 20 MHz) renders all services to users, which are defined by its technical indices, and at the same time can vary the information transmission rate depending the channel characteristics and the traffic intensity, using the multilevel modulation types.

At present, among other systems, the following systems with small operation radius, which have low data transmission rate (up to 20 kbit/s), have got the widest distribution: the system Zigbee, which is described by the IEEE 802.15.4 standard, the system Bluetooth (IEEE 802.15.1) with the transmission rate of 1 Mbit/s, and micro-cell networks DECT (ETS 300175, 176) with the rate of 320 kbit/s. If at present such networks are mainly used in networks of Internet of People, the intensive developments in the field of creation of the global network Internet of Things (IoT) will require the inclusion in its structure of networks, which have a capability to transmit on small distances (several 10 m) the informational streams with the rate of several GBit/s.

These are *ultra-wideband networks* of the IEEE 805.15.3a standard operating in the frequency band of 3.1/10.6 GHz [71, 16] with the power spectral density restricted by -41.3 dBm/MHz. The special place at construction of the IoT network is allocated to the system of *radiofrequency identification* **RFID** [52, 86] providing the wireless identification of objects, correction, and registration of data. The system includes tags and readers. Tags receive via radio channel the data about the object, store these data, and transmit them via the radio channel according to request. Such systems are built on various frequencies: from 10 MHz to 900 MHz.

At present, the existence of telecommunication systems and systems of different destinations is already unthinkable without interconnection with navigational systems, which allow determination of object location and its displacement in a space. The role of such systems merely increases with the implementation of communication systems of the fifth generation, which by natural course is embedded in the IoT ideology.

Radio Receivers in the Radio Electronic Systems

Interaction of separate devices and units of above-considered wireless systems and networks is inconceivable without communication means, where the main role belongs to radio receivers and radio transmitters. Their appearance during more than centenary history of radio from A. Popov and G. Marconi becomes a base of modern radio receivers. Application of digital methods of modulation and coding allowed obtaining of the data transmission rate R (the capability of the communication channel) close to limited values [16], according to Shannon:

$$C = W \log_2 \left(1 + \frac{P_{av}}{WN_0} \right) \quad (3)$$

where P_{av} is the average power of the input signal, W is the width of the channel (upper modulation frequency) with strictly limited band, N_0 is the spectral density of the additive Gaussian noise, and WN_0 is the noise power in the communication channel. From the moment of the radio receiver invention, their structural diagrams were improved relying on inventions of electronic components (electronic tubes, semiconductor diodes and transistors), as well as on needs of rate of increase of transmitted information, which in turn led to assimilation of new frequency ranges. The main task solving at receiver development was the allocation of the radio frequency band, where the supposed information source is located, and suppression of different types of interferences, which degrade the quality of the received signal.

Radio receivers, which effectively operated in the low-frequency region, were constructed using the structure of *receivers with direct conversion*. However, the high level of external interferences and the necessity to operate in the frequency range complicated the construction by insertion of simultaneously and uniformly tuning filters (single oscillating circuits). The increase in the signal level of the radio frequency front-end output by means of inclusion of amplifiers, which are loaded on the single oscillating circuits, amplified the interference leaking in the stages inputs, due to the broadbandness of circuits. This restricted the sensitivity of such receivers and complicated their operation at transfer in the higher frequency range, creating the self-excitation danger of the chain of cascade-connected amplifiers. The construction of such a receiver allowed

formation of the output of the simple nonlinear (noncoherent) detector of the voltage, which was extracted with the help of the low-pass filter and was close in shape to the law of the envelope variation of the input signal.

Appearance in 1918 (100 years ago) of the new *super-heterodyne structure* of the RF front-end executed the real revolution in the design of radio receivers. In this structure, the small retuned part extracting the operating frequency band and suppressing the collateral conversion channels is saved, and then the non-tuned part is included, which has the receiver part with non-changed parameters up to the device of signal reproduction. This permitted to design receivers of amplitude-modulated oscillations for any frequency region by changing only the input circuit and parameters of the local oscillator. Such circumstances of designing significantly simplified the receiver manufacture and decreased the cost, leading to active assimilation of the high frequency region saving the stability of receiver operation due to operation frequency transfer in lower radio frequency region – intermediate frequency region $f_s = f_{IF}$. The increase in the stage number in the intermediate frequency section allowed the growth of the receiver real sensitivity compared to the receiver of direct conversion of the same frequency range at the same SNR in the detector input.

Such receiver architecture with all its advantages has serious shortcoming – formation of collateral conversion channels among which the most dangerous is represented by: the channel of direct amplification and the concentrated interference $f_{int} = f_{IF}$ on the intermediate frequency and on the image channel frequency $f_{im.ch} = f_{int} = f_s \pm f_{IF}$. Therefore, the main function of the preselector (stages inserted between an antenna and a mixer) is the suppression of conversion channels, and the main task of the receiver is the suppression of the adjacent channel and it is solved in the intermediate frequency section. This simplifies the preselector construction making it more wideband and, at the same time, simplifies the implementation of the intermediate frequency section by using the concentrated filtering of the adjacent channel with the help of standard discrete piezo-ceramic and piezo-electric filters. Detection of the received signal can be performed depending on the system type and the modulation type with application of the coherent (synchronous) and noncoherent detectors.

The structural diagram of the direct conversion receiver known approximately from the middle of the past century but not found as wide distribution in the analog systems is actively used at implementation of modern devices with digital signal processing. The main difference of receivers realized on this architecture is the transfer of received useful signal spectrum not to the intermediate frequency but in the low-frequency region (modulation frequency in the transmitter, of the base band region). The carrier frequency (the average frequency of the receiving signal) f_s is converted to the zero frequency (DC component) of the converted signal. Extraction of the frequency band of the useful signal (suppression of the adjacent channel) is performed with the help of simple low-pass filters (LPS) of 4th–5th order. The task of the circuit preselector, which is inserted in the mixer input, is the extraction of the operating frequency band, which reduces the interference effect created by the adjacent systems and increase of the receiver sensitivity.

The main difficulty at realization of *direct conversion receivers* is the construction of the mixer (synchronous detector), for effective operation of which it is required to provide the synchronism and in-phase of signals acting in its inputs. This restricted application of such receivers in analog systems especially at operation in the frequency range. In communication systems with the phase modulation (BPSK, QPSK) and the multi-position amplitude-phase modulation nQAM, this task is successfully solved with the application of frequency synthesizers in receivers, which are covered by the phase-locked loop (PLL) system and voltage-controlled oscillators (VCO).

The common trend in the development of telecommunication and infocommunication systems to creation of new global system IoT renews an interest to direct conversion receivers. The necessity of the global roaming between different systems at IoT realization requires the terminals of the signal processing with various interfaces. Solution of this task is possible either with utilization receivers, which operate only with the single radio access system, and hence, with its number growth, we have multiple increase of the radio block number, or application of the *software radio* (SR) conception. In the last case, the input part of the receiver with the software control is implemented as wideband and in it, on the radio frequency, the analog-to-digital conversion (ADC) is performed, which allows elimination of the part of the radio front-end, and transfers these functions in the software-controlled digital part. At realization of such a receiver, the main difficulty is necessity of high speed (high frequency). ADC provides the required accuracy of signal conversion. At present, such an architecture of the SR-receiver coincides with the architecture of the super-heterodyne receiver, and the analog-to-digital conversion is performed on the intermediate frequency, that is, we have the receiver with the *digital intermediate frequency*.

Receivers of Modern Systems with the Digital Modulation

The receiver structural diagrams described in Sect. 4 are used also in the modern digital systems of fixed and mobile communication and TV and radio broadcasting, satellite personal communication, etc., but they have some features connected to methods of group signal formation, types of used modulation/demodulation, features of detection (decoding) of the digital signal, and so on.

At present, among the telecommunication systems, the most claimed are systems of mobile communication and the radio access. Nevertheless, methods of construction of base stations and subscriber terminals and used technologies of radio signal formation and methods for fight with interferences and network construction happen to be close enough for networks of different destinations.

The subscriber terminal and the base station are implemented as transceivers, which simplifies the construction of the frequency synthesizer and functioning of digital processors in both devices. This makes the whole device more compact and using the common antenna. For speech, data, and control signals transmission between the subscriber terminal and the base station, the complicate methods of digital signal processing and modern methods for coding and modulation are used, which provide the transmission rates comparable with rates of the wired networks.

Methods of Group Signal Formation and Methods of Multiplexing

The system of the GSM standard. In digital systems, transmitted signals from different sources form the group signal, representing a sequence, which is divided into radio frames with duration of 20 ms in the GSM standard and 10 ms in UMTS and LTE systems. To provide frequency duplexing mode (FDD), the paired frequency bands are allocated to systems: in the first band, the radio signal transmission is provided, while in the second band, the reception to the common antenna operates on the reception or on the transmission.

In the first case, the antenna is connected to the receiver input on the length of one radio frame, and in the next frame, it connects to the output of the transmitter power amplifier with the help of the input radio frequency module (e.g., the simplest switch). At utilization of the single frequency band on reception and transmission (TDD), each radio frame is divided 50-50, and in one half, the radio signal is transmitted, and in the other half, the reception occurs. To exclude the possible overlap of reception-transmission intervals, the protection intervals, in which information is not transmitted, are introduced in the radio frame. Subscriber terminals belonging to the separate cell, which are included in the network structure of usually cellular structure, communicate through the physical channel through radio interface with BS belonging to the ground-based network with the radio access. The subscriber terminal should provide the operation in any frequency range of the system corresponding to the chosen standard.

The most common second-generation digital systems, GSM (Global System for Mobile Communications) and Damps (Digital Advanced Mobile Phone System), are designed to transmit speech signals using frequency manipulation with minimal phase shift keying (MSK) or GMSK, which uses MSK with a preliminary limitation of the frequency band of the modulating sequence by a Gauss filter. To organize the simultaneous transmission of signals in opposite directions, frequency duplexing FDD (Frequency Duplexes Division) is used, when transmission and reception in each terminal device of the radio channel (BS and MS) is carried out in a separate frequency range. Multi-station access to the radio channel is provided by the method of frequency multiplexing FDMA (Frequency Division Multiple Access), when each connection in the radio channel is allocated a frequency band with a width of 200 kHz (for the GSM system). On the receiving side, the useful signal is allocated by a narrow-band filter that eliminates the influence of neighboring channels.

The frequency band allocated to one subscriber in the GSM system is significantly larger than the bandwidth of the standard channel of the tone frequency of analog systems, which is caused by the rectangular shape of the signal modulating sequence. As is known [20], the spectral power density of such a video signal has a maximum value near zero, and its envelope is described by the ratio $(\sin x / x)^2$ with a theoretically infinite width of the occupied frequency band.

It is obvious that such a signal cannot be transmitted over a radio channel with a limited frequency band. Limiting the bandwidth allocated to the subscriber is possible by turning on a band-pass filter (BPF) at the input of the signal converter “up” into the operating frequency band or at the output of the power amplifier in the transmission path, which simultaneously reduces the level of the side lobes of the antenna pattern. But the method of limiting the frequency band occupied by one subscriber in the GSM standard, which uses the inclusion of a Gaussian filter at the input of the transmitter modulator with

a bandwidth $\Delta f_{GF} = 81.3$ kHz, has become the most widespread. Then, in the radio channel, the modulated signal will occupy a frequency band more than twice the bandwidth of the Gauss filter.

Along with frequency division access (FDMA), the GSM standard simultaneously uses time multiplexing access (TDMA), when not one signal is transmitted on each carrier in a user frequency band with a width of 200 kHz, but signals of eight subscribers are transmitted sequentially at a length of one bit. This increases the number of subscribers with access to the radio channel by eight times and increases the efficiency of the system as a whole.

In the *system of third generation* (3G), the frequency and code channel division **FDMA/CDMA** is jointly used, when the frequency band for separate network, in which all subscribers operate simultaneously, is allocated in the system frequency band. Each from subscriber terminals has a distinguished feature – the unique (etalon) code sequence, which is orthogonal to any other sequence used in the network, is conferred to each terminal.

Orthogonality of spread sequences permitted to transfer to utilization of the whole frequency band of the radio channel by all subscribers of the cell, which provided the signal transmission rate in the radio channel up to 2000 kbit/s depending on the type of transmitted information and unattainable in the GSM standard. This provided the independence of source signals (traffic, synchronization, transmitter power control, etc.), and application of phase modulation provided the utilization of correlation processing of the receiving signal. Spreading of signal source spectrum is simultaneously the method of fight with the interference effect in the receiver input. The sequence of Walsh functions containing narrow pulses (chips) decreases the signal spectral density on the output of the spreading device and leads to larger spectrum spreading making it noise-like. At the same time, application of short spreading pulses makes the sequences, which arrive to the receiver input by different ways, independent if the delay is more than the single chip (0.26 μ s).

The next procedure is scrambling, which is usually realized by summation modulo 2 of transmitted sequence and the pseudo-random sequence (PRS), providing approximately constancy of average values of the quadratic component power in the output of spreading devices. Pseudo-random sequences are usually the Gold sequence truncated to 10 ms, have the length of 38,462 chips with the repetition period of 215–1, and intended for cell distinguish.

For radio signal transmission via the channel with the restricted pass band, we need to connect to the up-converter output of the band-pass filter, which limits the frequency band before the spectrum transfer of the modulated signal into the operating range saving mutual orthogonality of its responses. This is achieved by insertion in the transmitter section of the Nyquist filter having the frequency response of the raised cosine type *RC* with the rounding coefficient $\alpha = 0.22$ (for the LTE system). Application of the common radio channel with the fixed bandwidth (1.25 MHz for the CDMA2000 system and 5 MHz for the UMTS system) allows utilization for pulse coding of the information sequence by Walsh functions of different lengths. Depending on the coding sequence length, this leads to variation of the service zone (the cell) radius: cells “breathe” at constant speed of elementary pulse of 1.23 Mchip/s in the cdma2000 network and 3.84 Mchip/s in the UMTS network.

4G systems. In systems of wideband access, mobile communication, personal satellite communication, digital broadcasting, and the technology of orthogonal multiplexing with frequency channel division, Orthogonal Frequency Division Multiplex (OFDM), is widely used. This technology popularity relates to achievement of high data transmission rate in the radio channel, its protectability against the inter-symbol interference caused by the multipath property, by the flexibility in rendering of frequency band to the subscriber (the modulation speed), etc. Standards for the 4G systems of mobile communication, Long-Term Evolution (LTE), anticipate that the maximal data transmission rate in the radio channel must be 1 Gbit/s in downlink and 300 Mbit/s in uplink for the width of the allocated frequency band of 20 MHz.

The feature of the OFDM technology is characterized by the fact that evenly existing multipath property, which is a reason for the inter-symbol interference, is taking into consideration the step of group signal formation. Information sequences from separate sources form the group signal with the sequential structure, which after interleaving and sequential/parallel transformation according to some algorithm, is transformed into the group of mutually orthogonal sequences. The symbol duration in each formed sequences is significantly more than the duration of symbols in the group signal and is $T_0 = 66.6 \mu$ s for the LTE system, which provides the orthogonalities of modulated sequences in time. The choice of such an interval of orthogonality provides also the orthogonality of modulated carriers by them, which is spaced in frequency by $\Delta f = 1/T_0 = 15$ kHz. In other systems using the OFDM technology (radio access, digital broadcasting, TV, etc.), these quantities differ from that mentioned for the LTE system and are defined by properties of modulating sequences. To save the mutual orthogonality of slow video sequences during their conversion and passage through the radio sections of the receiver and the transmitter and in the free space, which is caused by inter-symbol interference, the radio signal in the transmitter input is exposed to predistortions with the help of the filter with AFC of the raised cosine *RC*.

Besides the frequency band reduction, which is occupied in the modulators output, such a filter forms the sequence of orthogonal responses, which is transferred into the radio channel with the help of the up-converter. In the radio frequency front-end output, with the help of the matched filter, the correction eliminates, which was introduced in the transmitter, and

therefore, the orthogonality of received symbols is recovered. The distortion of the video signal shape as the appearance of the time of the rise-up portion of the digital signal in the output of the radio receiver front-end leads to disturbance of mutual orthogonality. Recovering of orthogonality during symbol duration is performed by introduction of the T_g protected interval in the initial segment of the symbol duration T_0 , replacing it by the finite part of the proper symbol of the same length. By this, the mutual orthogonality of slow modulating symbols is achieved on the full length T_0 of the symbol.

The group of orthogonal symbols from different sources that decreases the danger of error grouping, which has the parallel structure, is integrated according to some algorithm into OFDM symbols with duration T_0 . Sequences included in five (or six) OFDM symbols are multiplied by 12 orthogonal sub-carriers forming the radio block, which is the main unit of the spectral domain of the frequency-time structure allocated to the system. The integral number of such orthogonal radio blocks NRB defines the system effective frequency band. Each OFDM symbols in the time domain is separated from other by the cyclic prefix (CP) providing the mutual orthogonality of radio blocks and symbols. The modulation procedure of each from mutually orthogonal sub-carriers can be performed with the help of the quadrature I/Q modulator (by circuit theory) or by the software approach: fulfillment of the inverse Fourier transform (IFFT). To increase the effectiveness of the Fourier transform, a number of Fourier transform points should exceed the number of really used sub-carriers: $NRB \times \Delta f$. Evidently, since a number of used sub-carrier depend on the frequency bandwidth allocated to the LTE system and can be more than 2000 at the bandwidth of 20 MHz and for application of the I/Q modulator, this makes the IFFT utilization preferable. Transmission through the radio channel, which has the restricted band pass and dynamically changed parameters, of the signal with the theoretically infinite bandwidth requires its limitation with keeping the mutual orthogonality of sub-carriers. This is provided by connection in the transmitter input of the up-converter of the Nyquist filter having the frequency response of the raised cosine type RC with the rounding coefficient $\alpha = 0.22$ [21, 24].

Obviously in the receiver, the functions of signal conversion are realized, which are conjugated to be used in the transmitter; therefore, in the receiver input, the matched filtering of the received signal is used providing minimal error probability. After signal conversion from the operating frequency domain into the intermediate frequency domain and reformatting in the parallel form, the equalizing is performed. Signals on the separate sub-carriers, which have the same power, are exposed to the Fourier transform than, after conversion from the parallel structure into sequential, they are demultiplexed forming informational sequences for the separate subscribers (traffic, synchronization, etc.).

As it follows from features' consideration of CDMA and OFDM technologies, the base for increasing the probability of correct reception is the requirement of signal orthogonality in the time domain (CDMA) or in time and frequency domains at the same time (OFDM). To achieve this, the Nyquist filters are used on the transmitting side, and for recovering mutual orthogonality in the receiving side, the matched filtering is applied. This is the necessary condition for signal coherent detection with the phase (BPSK, QPSK) and multilevel amplitudes-phase (nQAM) modulations.

Another necessary condition for achievement of the maximal signal transmission rate in the channel closed to the limited-possible according to Cl. Shannon (Sect. 3) is the transfer from the character-oriented detection (decoding) to processing of the symbol chain (informational sequences). Decision making about obtained «logical 1» or «logical 0» can be achieved by the utilization of the Witterby algorithm, obtaining the shortest way from one from the states in the $(n-1)$ moment and the way from this state into the initial state in the n -moment on the lattice diagram. Application of the iteration mode for practical realization of the method of the shortest distance search simplifies the obtaining of the minimal function of the D cost, which is the specific case of the maximal likelihood criterion.

Elements of the Radio Front-End: Features of Realization

Described features of signal processing relate to application of digital technology, and they almost do not touch the receiver radio front-end. Therefore, considered structural diagrams of the receiver construction for analog modulation methods have saved their urgency at development of modern radio receiving devices including the utilization of integral technologies. A choice of specific method of the radio front-end implementation is mainly defined by the level of technological achievements in the development of ICs for the UHF range (band-pass filters, analog-to-digital converters, low-noise amplifiers, frequency converters).

Studying of the operation features of each mentioned devices allows optimization of the operation modes of active components and the choice of the coupling circuit parameters, including the type and depth of the negative feedback. The matching condition of the active component with the standard wave resistance of feeding lines is obligatory for ICs, which realize

both the separate function (amplification, conversion) and stages included in the system structure. This condition permits to investigate separate stages independently on parameters of connected circuits.

Modern large ICs including stage of different destination are often used the known operation principles but realized with utilization of new technologies. So, the field-effect MOS transistors, often together with the traditional application as amplifying components, are actively used as active loads of amplifying stages. On their base, the active frequency converters known as the Gilbert cell are developed, and these converters are close in properties to the ideal multipliers, do not require the high level of the input signal, and are widely used in receivers. The need to apply the unified technology for manufacture of active and passive components has led to wide utilization of MOS transistors as equivalents of diodes (transistors in diode connection), parametric components with controlled conductivity. Functional blocks can be created on their base, for instance, oscillators of stable current, electronic switchers, and using the method of component imitation, to realize passive components: equivalents of resistors, capacitors.

During the manufacture of passive components in receivers of the microwave range, both traditional methods of its implementation, which are based on utilization of strip line segments, and the planar inductances are used. The last elements turn out to be technologically simpler, especially at transfer to operation in the higher frequency range. This relates to reduction of physical sizes of the strip line segments, and hence, the large dispersion of inductance values at their implementation. Well-known in radio electronics, piezo-electric components have got new application as the band-pass filters of the microwave range based on propagation of surface acoustic waves in composite materials. Such filters have the small losses on surface wave propagation, and simple technology of conducting structure laying on the substrate surface can implement the amplitude-frequency characteristics (AFC) to provide filtering properties, which have practically any one thinkable characteristic. The process of such filters' manufacture provides high repetition of forming indices both in the band pass and in the stop band, providing the high degree of collateral channel suppression.

Operation amplifiers realized on the MOS transistors become a base for the whole class of band-pass filters known as complex or polyphase [25], which have the changing band pass and low energy consumption. New class of filters on the base of programmable integrated circuits, which are realized on MOS transistors, are the low-pass filters and band-pass filters on switched capacitors [105]. Such devices are characterized by high parameter stability, low energy consumption, and the possibility to control of the band-pass width by electronic methods.

Antennas and Input Circuits

The important feature of the radio receiving device of any destination in microwave range is the application of tuned antennas, whose parameters remain unchanged at operation in the whole system frequency range. This is peculiar to all systems besides the ultra-wideband system operating in all allocated frequency range from 3.1 to 10.3 GHz. To operate simultaneously in several frequency ranges, for instance, GSM900 and GSM1800, we should apply two antennas, which are considered as tuned in each sub-range. It is admissible at smallness of the ratio of the system frequency band to the average operating frequency, which is the value less than 10/15%, when the transfer function of the input circuit can be considered as constant at given irregularity. Utilization of the tuned antenna and non-retuned input circuits leads to appearance of intermodulation distortions as a consequence of powerful out-of-band interference influence, which are typical for cellular networks. Estimation of their effect on characteristics of the receiving section significantly influences on its structure and must be considered at network planning. A tendency to increase the signal transmission rate in the radio front-end by expanding the list of rendering services, which was mainly solved by complication of the processing algorithm of the digital signal, was supplemented by technologies using the formation of spatially separated channels.

Application of multi-antenna systems forming the narrow-directed beams (by beamforming) and spatial multiplexing (the MIMO technology) is one of the methods providing the growth of the data transmission rate. The diversity reception and transmission, which was applied earlier in the HF and VHF ranges, when several antennas are used, finds the limited application in mobile communication systems. This is caused by small physical sizes of the subscriber terminal, while for elimination of mutual correlation between channels, the distance between receiving antennas should be not less than a half of wavelength. For this reason, it is impossible to realize the traditional antenna diversion in the subscriber terminal for the frequency range below 700 MHz.

In the wideband systems (UMTS, cdma2000), the multi-beam diversity is used; when signals arrived to the receiver input by different ways are exposed to correlation processing, then they are brought into phased, multiplied by the weight coeffi-

cients, and added (the RAKE receiver). An absence of correlation between arrived echoes is achieved only in the case, when the minimal difference between the time delays of echoes exceeds the chip duration.

As an alternative of the spatial diversity in UMTS systems of mobile communication, the spatial multiplexing has got a distribution in LTE, which uses the really existing multipath phenomenon in the radio channel, when each antenna transmits at the same time several independent data streams [22, 112, 113]. Utilization of multi-channel antennas (MIMO) is effective, when the subscriber terminal is capable to distinguish the spatial channels. This is effective only at enough high SNR, which is typical for cells of small sizes. Reduction of antenna physical dimensions and the decrease in distance between the subscriber terminal and the base station at operation in the frequency range above 1000 MHz allow easily realization of the MIMO technology in any subscriber terminal. Application of two antennas in the subscriber terminal provides the energetic improvement of the received speech signal by 2/5 dB (the frequency range 700 MHz) and by 6/9 dB (the frequency range 2100 MHz), and it increases approximately by 1/2 dB at data transmission.

The systems of 3G mobile communication and 4G radio access of CDMA and OFDM technologies, considered as examples of specific types, find the wide application in the other types of systems. These can be the base technologies adapted for traffic features, as well as combination of technologies, for example, UWB/OFDM, as in the ultra-wideband communication system. Realized on different element base and using the proper technologies, they supplement the advantages of each of them.

Noise Characteristics

Considering that modern radio engineering systems are mainly used in the microwave range, the sources of external noise at the receiver input are mainly galactic objects, thermal noise of the Earth's surface, and antennas. In this case, it is necessary to carefully select the components of the input circuit (preselector) of the receiver, which have a major influence on the sensitivity of the receiver. These include, first of all, AE in the first stages of the radio path (amplifiers, mixers), which create the main contribution to the intrinsic noise coefficient of the radio path (and the entire receiver). These noises, amplified by the following cascades, make the greatest contribution to the resulting noise power at the input of the detector (demodulator) of the receiver and reduce the reception quality due to a decrease in the signal-to-noise ratio.

Sources of internal noise of the receiver are both passive components of the circuit (resistors), which have losses and create thermal noise, and active components (transistors). In addition to thermal noises, transistors also produce the shot noises, the flicker noises of the channel (for MOS transistors), and the noise of current distribution between a base and a collector for bipolar transistors. Therefore, at input circuit implementation, in order to achieve the highest receiver sensitivity, in some cases, we must use matching according to the minimum noise instead of the antenna matching mode to the power.

Modeling of Units and Devices

Analytic investigation of stage properties in the radio front-end is restricted by the adequacy of used models of active components. If in the frequency ranges below VHF in an analysis of the stage properties of the radio front-end, we can use the linear physical model or the equivalent circuit for description of the small-signal parameters of transistors, then in the microwave range, we must use other approaches of their descriptions. To describe transistors, we may use the scattering matrices in the system of S-parameters [64] restricting by the power characteristics of the examined devices.

Another method, which saves the usual process description in the stage or in the device through time and frequency characteristics, is the utilization of computer modeling. This becomes especially relevant in connection with complication of active component models, which is described through parameters of semiconductor materials during manufacture, for instance, the MOS transistors. They contain information not only on the properties of transistor active regions (channels, junctions) but about mobility of charge carriers of the substrate material, the influence of the volume charge forming in the semiconductor structure, and so on.

The task of property investigation of separate stages and units is essentially simplified by using SPICE models [28, 96] of transistors and micro-assemblies. The modern professional software packages for an analysis of separate stages or units, such as AWR, MATLAB/SIMULINK, and Agilent, allow obtaining usual characteristics in frequency and time regions. However, application of the professional software for educational purposes is a sufficiently expensive and labor-intensive process, and therefore it is not always a defensible method for properties investigation of separate stages. As the experience

of the MicroCAP software [31] application shows that even in truncated (demo) version, which has a friendly interface, this version allows performing the analysis of separate radio front-end units even by students, who have limited experience with the computer. Application of the student and particularly professional version of MicroCAP in educational process allows usage of both simplified and complete SPICE models of active components by saving the “physical sense” of examined processes without the expense of time with the interface simplicity.

The total organization of our book can be clearly seen from the table of contents. Of course, in our opinion, this book may present an interest not only for students but also for engineers-developers of companies and research organizations, who are involved in the development and design of modern radio receivers. The distinguished features of this book are less number of complicated formulas and many complicated issues are explained to students by words instead of formulas.

With this remark, we wanted to point out to the reader that the text may contain terms that have not yet become generally accepted in the field of communications and radio engineering systems, despite their frequent use in regulatory guidance documents, professional periodicals, and manuals for students of colleges and universities. If you think that the material presented in the book is presented clearly for future specialists, then we agree to remove this sentence from the text of the book.

Their sources are both passive components of the circuit (resistors), which have losses and create thermal interference, and active components (transistors). In addition to thermal noise, transistors also produce shot noise, channel flicker noise (for MOSFETs), and current distribution noise between the base and collector for bipolar transistors. Therefore, in order to achieve the greatest sensitivity of the receiver in some cases, for example, when developing professional devices, it is required to use the antenna matching mode to minimize the noise coefficient instead of the power matching mode when implementing the input circuit or the preslector.

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Introduction

Wireless communication systems that use free space as a medium of transmission (ether) actually appeared immediately after registering the phenomenon of propagation of electromagnetic waves with the help of a lightning detector invented by A. S. Popov in 1895. Depending on their purpose, radio engineering systems can be distinguished that are designed for: signal transmission, signal detection, and combating organized message transmission. The simplest networks for wireless transmission of analog signals include a single transmitter and one or more receivers. Their ability to move in space without losing communication between subscribers under certain conditions led to the appearance of mobile communication systems.

The scarcity of the free frequency spectrum has led to the widespread use in wireless communication systems of multiple access technology to a common radio channel, which provides simultaneous operation of several users in a common channel. This significantly complicated not only the structure of the terminal devices in the transmission systems (transmitter and receiver) but also the structure of the network, which includes a terrestrial network, usually a cellular structure and a core network. The terrestrial network provides a radio interface at the physical level between the base station (BS) and the subscriber terminal (MS). Through organized logical channels, BSs exchange commands with the network controller (RAN), containing not only traffic but also service information about the network structure, channel status, types of connected devices and networks, and information about the recipient of information.

The controller's functions also include evaluating the traffic intensity and assigning the type of modulation in the radio channel, as well as relay transmission from one base station to another when it moves within a cell, network, or between networks of different operators. Thus, the pure physical radio channel in mobile systems, without which it is impossible to build a system, exists only between the BS in the terrestrial radio access network and the subscriber connected to it. The BS receives information from the controller on service channels about the network status and types of connected external networks, changes the structure of logical channels to transport channels and transmits received commands from the top level of the hierarchy to the AT about the level of radiated power by the AT transmitter, type of modulation, transmission speed data, recipient address, etc.

Replacing the continuous signals of message sources with their digital counterparts significantly improved the efficiency of using the frequency spectrum, for example, up to 2.6 bit/s/Hz for the LTE system, which was the result of the introduction of channel coding, noise-tolerant multi-access technologies, sequential detection, that is, the improvement of the transceiver. The desire to increase the transmission speed and reduce the delay time in the network at all stages of processing the transmitted signal, including reducing the busy time of the allocated frequency range, required to switch from channel switching, as was used in the 2G system, to packet switching. In the first case, the frequency range with a width of 200 kHz was fixed for the duration of a communication session between two subscribers. During the same period, transport and network level channels were formed, which led to irrational use of the frequency and time resource allocated to the system. In packet-switched mode, the entire array of data (traffic and service information) transmitted at all levels is divided into packets of various lengths determined by the transmitted information. Short packets with an indication of its size, the address of the previous and subsequent packets in the structure of the transmitted message, and the recipient's radio access network (BS) are transmitted over a common channel of a high-speed terrestrial or other type of network. With a large length of the formed packet, it is automatically divided into two or three short ones.

Attachment of an equipment of various destination (and from various manufacturers) with the help of interfaces provides the effective interconnection not only by connecting equipment, but different systems, for example, the telephone public network and the system of satellite communication. For interaction of the subscriber terminal and the base station on the physical level, it is enough to be limited by information about the type and number of levels of the carrier oscillation switching, the data transmission rate, the occupied frequency band, the duration of the radio frame, and the method of compression and duplexing.

The multilevel model of digital systems allows taking into account not only the type of used modulation, the data transmission rate, but, the method of its switching. In systems of the second generation, the fragment containing 456 bits of information transmits during 20 ms (a frame) on the allocated channel in the operation frequency band with the width of 200 kHz. The channel is fixed for subscribers during all time for connection in all levels of interconnection of the model, which reduces the effectiveness of the frequency-time resource utilization, which is allocated to the system.

The need in the interaction of radio electronic systems of various destinations led to the necessity of its structure unification and to application of unified standards to their interfaces approaching them to the Internet standards. A possibility of different system interaction was provided by implementation of the unified standard of interconnection of the open systems OSI defined levels of interconnection (physical, transport, network) and types of formed channels providing the data exchange.

One carrier, traditionally used in the radio channel of analog and digital fixed and mobile radio systems, has been replaced in the transceivers of systems using OFDM technology with a group of orthogonal carriers. Each subcarrier in the transmitter path is modulated by an orthogonal sequence formed according to some algorithm from the information sequence. The pulse length of the modulating sequence T exceeds the duration of the information sequence pulse T_c , a number of times equal to the number of subcarriers in one resource block.

Described methods of signal processing had found the further development at construction of the terminal equipment of the 5G systems. At the same time, they must take into consideration the constructive peculiarities and properties of the radio channel in the allocated frequency range, as well as methods of construction of the multilayer networks on the base of femto-cells, pico-cells, and mega-cells.

The terminal devices of the mobile communication systems (subscriber terminal, base stations) in the radio channel structure usually include both transmitting and receiving devices. On the one hand, this allows the organization of communication in opposite directions in the real-time mode, and from the other hand, it complicates its configuration and makes the high requirements to the receiver sensitivity. At the same time, this forms the increased requirements to the power supplies of the subscriber terminals and to conditions of high effectiveness achievement of operation modes of active elements of transmitters.

The RF front-end of modern receivers of digital radio engineering systems is implemented using the basic structures of analog signal receivers. However, the transition to operation in the microwave range made it possible to simplify the design of the preselector, making it non-tunable. Mastering the methods of digital modulation has changed the methods of processing received signals, supplementing the incoherence detector of analog receivers with a demodulator that performs coherent processing of received signals, forming a group of symbols at its output. The decision about the received signal is made in the detector based on the analysis of a group of symbols using the maximum likelihood method.

Conversion of a symbolic sequence into a digital one in the ADC made it possible to switch to sequential detection methods, when the decision on the adopted “logical zero” or “one” is made based on a group of elementary symbols, which significantly reduces the probability of erroneous reception. The use of coherent demodulation methods in the receiver has led to the replacement of local oscillators with voltage-controlled oscillators (VCOs) from the digital part of the receivers.

The contradictory requirement of simplicity of frequency control of the VCO in combination with the high frequency stability of the harmonic voltage created is realized with the help of a reference generator, the frequency of which is stabilized by quartz, and a frequency synthesizer that generates a voltage close to the voltage of the specified frequency at the output of the VCO. Taking into consideration the modern tendency of implementation of the whole device on the single substrate (on one-chip) and the transfer to programmable methods of signal processing, the structural diagrams of radio receivers become more and more often the functional diagrams describing processes, which occur in the part of on-chip structure keeping the purpose of separate units, although realizing the other principle of construction.

The complex approach to solution of the task of data transmission rate growth at given SNR requires (parallel to the application of coherent signal processing methods) to solve, for instance, the problem of the unauthorized access to the channel, which is the part of the struggle for reduction of the transmitting spectrum power. This concerns the aspiration for decreasing an influence of the effect “near-far.” Transmission on the opposite channel, not only the confirmation signal ACK/NACK of the correctness of received signal but the requests on the variation of the transmitting message power, the variation of the modulation type, the message transmission rate variations, the channel parameters, etc., led to increase in the true reception probability.

The main goal of opposite channel organization is the arrangement of the traffic (duplex mode) together with the overhead information. At present, the frequency duplex (FDD) is widely used, which requires the allocation of the twice band of occupied frequency (the one band for signal transmission, the second for reception) compared to the time duplex (TDD), which has a more complicate processing algorithm for generated and received signals in the common frequency band.

Up to now, the main ways of the high-frequency part perfection of the terminal equipment are directed to the increase in data transmission rate. They relate to the extension of the transmitting/receiving frequency band due to the spectrum aggregation and almost do not depend on the modulation type. The application of spatial-temporal coding together with MIMO technology can solve this task at the expense of the more effective utilization of allocated frequency band and the complication of the signal processing algorithm in the analog part of the receiver.

Implementation of various technologies of channel division (CDMA, OFDMA) in systems, which are exploited in different frequency ranges and provide the maximal possible data transmission rate under condition of multipath property and the cellular structure of the ground-based network of radio access, caused the necessity of convergence of existing and promising technologies based on the unified type of equipment. Topicality of such a reform confirms a deficit of available frequency spectrum, which required the effectiveness increase of its utilization. One from ways for this goal achievement is the reuse of the allocated spectrum by introduction of the dynamic control of the free spectrum in order to solve the specific task in the current time interval. Another approach for effectiveness increase is the variation of the structure of network radio access by means of the creation of multispan retranslations. This creates a need in formation of new network elements and the subscriber equipment adapted under the specific conditions.

This puts forward new requirements for the modernization of the network architecture, including the use of virtual nodes and cloud storage, as well as changing the structure of user devices. The first step in the development of multi-standard, multi-band, and multifunctional mobile devices adapted to the environment was the creation of transceiver devices, the necessary indicators of which are provided by software defined radio (SDR). The development of this structure was the creation of cognitive (learning) radio (CR), the feature of which is the dynamic distribution of the spectrum between systems based on knowledge of the statistics of traffic changes in them. This creates a need for the formation of new network elements and new user equipment adapted to specific conditions in the user plane of the Enhanced UMTS Terrestrial Radio Access Network (E-UTRAN). The development of these areas of wireless network modernization creates new requirements for receivers that can meet emerging needs, which has led to the emergence of software-defined radio (SDR) and cognitive (training) radio (CR).

On the basis of SDR structure, there is the utilization of the equipment set and software technologies permitting the reconfiguration of the radio interface of wireless networks and subscriber terminals. Parameters of the receiving-transmitting equipment define not by the construction of the applied devices but by the software support. The basis of constructive realization of the radio receiver device is synthesis of two structural diagrams: the receiver with direct conversion and the receiver with digital intermediate frequency. Such circuit implementations allow the receiver to operate in a wide frequency range (up to 100 MHz), which is achieved by aggregating potentially free frequency bands of each of the interacting systems at the network planning stage. The structure of such receivers provides an effective solution to the problem of building multi-standard, multiband, and multifunctional mobile devices that can adapt to the external geophysical and electromagnetic environment.

The direct conversion of the receiving signal into the modulation frequency band (Base Band) permits simplification of the front-end of the receiver at some construction complication of the software program's adjusting antenna and the input radio-frequency module. The construction of the receiver with digital intermediate frequency must be simplified, and discretization (analog-to-digital conversion) should be provided on the frequency of the received signal, when the transfer into the region of reproduced frequencies is fulfilled by discretization of the intermediate frequency signal. The main difficulty in performing the microwave range signal discretization, which converts the radio signal directly in the input of digital signal processor, consists of constructive realization of such an ADC, although in this direction there is noticeable progress.

The signal in the programmable central signal processor (CSP) of the receiver with the SDR structure is exposed (after detection) by demodulation, correction, and channel filtering, as well as it specifies parameters of the synthesized frequencies. The structure of such receivers provides the effective solution of the task of construction of multi-standard, multirange, and multifunctioning mobile devices capable to adapt to the external geophysical medium and to electromagnetic situation.

Development and extension of possibilities of high-frequency part of the SDR receiver for monitoring of allocated energy spectrum in order to make a decision about the reuse of its part by another user. The restrictions of spectrum possibility utilization for the main owner should not arise. Such a combination of functions of the SDR system with dynamic analysis of the Incumbent Profile Detection (IPD) led to appearance of the cognitive radio (CR).

The earlier described statistical approach to aggregation, which allows the increase of transmission rate by broadening the occupied spectrum band, loses its urgency with utilization of the cognitive radio structure, when the spectrum extension occurs in the dynamic mode. The availability of such an approach confirms the active implementation of the cellular communication 5G system. In the standard system of new 5G generation (New Radio) having a huge frequency range (3 ... 77 GHz), the struggle for extension of the frequency band allocated to the operator ceases to be relevant. Evidently, that

development of the receiver operating in the whole frequency band is also inexpedient due to the strong dependence of parameters of the propagation medium upon the operating frequency and its growth with the frequency increase. The entire frequency range is divided into two sub-bands up to 6 GHz (FR 1: 0.415 ... 6 GHz) and above 6 GHz (F2: 24.25 ... 52.6 GHz). In the high-frequency part, femto- and pico-cells will be organized, which have a small coverage radius of several tens of meters but provide a data transfer rate of several Gbit/s. In the low-frequency range-networks with a large coverage radius in the urban and suburban service area with a transmission speed of about Gbit/s.

The variety of existing wired and, above all, wireless radio networks are determined mainly by their purpose, which also influenced the structure of the radio receivers included in them. It retained its functionality but changed its architecture in connection with the transition to digital form of transmitted signals and the development of multifunctional VLSI. This most strongly affected the organization of channels at the physical level of the global 5G network, which should provide interaction with sources of various types, modes of operation, and different frequency ranges, being mobile or stationary, and changed the architecture of receivers.

The first chapter of the book describes the features of the propagation of a radio signal in free space and the interference acting on it caused by various sources. Then, types of radio systems for transmitting information and options for their interaction are discussed using the example of a reference model.

The second chapter is devoted to fixed and mobile wireless networks that implement the function of transmitting messages in various systems and standards.

The third chapter describes the structural diagrams of radio receivers of the basic types of modulation with and without reconstruction of the preselector filters for fixed and mobile communication systems. Then, receiver structural diagrams for specific types of systems are considered, taking into account the applied modulation/demodulation, detection, and multi-access technology.

The fourth chapter describes the types of interference operating in the radio channel and their influence on the characteristics of the received signal in fixed and mobile communication systems. The main methods of reducing the error rate in the received digital sequence are considered, using preventive methods based on knowledge of the parameters of the radio channel in real time and circuitry, taking into account its static indicators. Particular attention is paid to the use of MIMO space-time coding technology.

The fifth chapter describes the numerical parameters of the RF front-end that characterize individual cascades, nodes, and the receiver as a whole. Such estimates allow us to compare linear and nonlinear distortions that occur in the front-end, to find the causes of the distortions that occur, and to formulate methods for their elimination.

In the sixth chapter, we consider the main indicators and methods of using the common Micro CAP program to model the properties of individual cascades of the RF front-end, described by the schematic diagram. To describe individual components, their PSPICE models are used, the conventional graphic notation (CGN) of which have the usual graphics symbols and are widely used in modern professional MATLAB/Simulink, Agilent, and AWR programs.

The appendices to each chapter of the book contain verification questions, examples of calculating the technical parameters of the RF front-end, and modeling units in the time and frequency domain.

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Chapter 1

Radio Systems and Radio Signals

We begin our acquaintance with considered problems in this chapter from the general initial description of radio telecommunication systems and radio signals. At first (Sect. 1.1), we formulate the most important definitions of radio systems used for information transmission and reception through the radio channel and then briefly describe its functions for system of different deployment (ground-based, mobile, aircraft-based, satellite-based). Then (Sect. 1.2) we briefly describe the features of electromagnetic oscillation (radio waves) propagation in various frequency ranges and the influence of the different layers' properties of the Earth atmosphere on this propagation. Surface and spatial electromagnetic waves and also the reflecting waves are discussed. Section 1.3 is devoted to brief description of the most important interferences and to issues of noise immunity provision of telecommunication systems. Classification on interference types includes external spurious radio signals, signals emitted by external radio transmitters, the atmosphere interferences, industrial interference, deep-space noises, internal noise of radio receivers, and other electromagnetic interferences. Concentrated (lumped), pulse, and fluctuation radio interferences both of additional and multiplicative types and methods of struggle against interferences and noises are briefly described. In Sect. 1.4 we consider the systems of fixed and mobile communications at first from the historical positions and then transferring to technical characteristics. We describe the most important models of communication systems and their functional structures taking into account the interaction of open systems. Special attention is paid to the so-called etalon systems and to its levels: physical, channel, network, transport, session, representation, and applied. The modes of channel and packet switching and the necessary protocols for them are discussed. Methods of duplexing for the radio communication systems are described, as well as the important issues of radio access providing to the communication systems and to communication networks for different modes of channel division in time, frequency, and code. Radio broadcasting systems, relay systems, mobile communication systems, and satellite systems in various digital variants of realization are described, as well as the most important navigational radio systems.

1.1 Types of Radio Systems

According to accepted terminology, under the *radio communication* we understand the communication realized with the help of radio waves. Radio waves are the electromagnetic oscillations propagating in a free space without the artificial waveguide (with frequency less than 3000 GHz).

Systems differing on application of radio communication principles for information transmission between the users can be divided in *ground-based* and *space-based* systems. Any radio communication system, with one or several space stations or one or several artificial Earth satellites, may be attributed to the space system. All radio communication systems besides space systems and radio astronomic systems can be attributed to the ground-based systems.

Depending on the *service customer type*, we may distinguish the following systems: fixed, mobile, and broadcasting systems, each of which can be ground-based or space-based. If at least one of the customers is in the moving process during the communication session, the communication system is mobile.

The technical system, in which operations are performed by radio electronic devices, is referred as the *radio system*. A transmitter, a receiver, the antenna-feeder devices, devices for message conversion into the electrical signal, and devices performing the inverse conversion as well as indicating devices are usually included in the radio system structure.

Due to expansion of the interpretation of concept of “communication”, radar and radio navigation systems and systems that prevent the interception of information during its transmission over communication channels are referred to as communication systems.

According to the information purpose, radio systems are intended for:

1. Information transmission
2. Information extraction
3. Contraction of information extraction

The first class of radio systems above **transmits information** of various types (e.g., a message) from one space point to others. In this case, the functional diagram of the transmission radio system is the radio or TV broadcasting system and shown in Fig. 1.1.

In Fig. 1.1, the information source creates the effect $\lambda(t)$ to EPIC (electric-physical information converter) in the form of varying audio pressure, the object illumination or color, etc., which converts in EPIC into the electric signal (U1) corresponding to the varying parameter law. The electric signal passes to the encoder (in the simplest case, a modulator, which varies some parameter of the carrier RF oscillation according to the input signal law, for instance, the amplitude U1) and then passes to the transmitter (Tx) (U2). In the transmitter, the signal is amplified, and its compact spectrum is formed and then passes to the antenna (U3), where the electric signal is converted into the electromagnetic field (E4), which is radiated into the free space (into the radio channel). During the radio signal formation in the transmitter and its conversion to the electromagnetic field, and in the radio channel, the signal is exposed to the influence of natural or industrial interferences. The *electromagnetic field* received together with interferences (E4') converts in the *receiving antenna* into the *electric signal* (U3'), where it is also exposed by the interference influence (U2'). In the receiver, attenuated and distorted electrical signal is amplified, the interference influence is attenuated, and the radio signal is converted into the bandwidth of modulating frequencies (U1') (the so-called baseband). The inverse conversion in EPIC from the low-frequency electric signal forms the impact in the form of varying audio pressure (the audio signal) and the brightness or color (the TV signal), which are perceived as the some message $\lambda'(t)$ by the information customer. Evidently, under interference impact, the original message suffers a distortion, and the received message will reproduce it with definite accuracy. The accuracy of received analog message reproduction is estimated by the truth and for digital messages by the probability of correct reception or the error.

In systems of *fixed radio communication*, the counter-direction is created for information transmission, if the separate devices are used as the terminals, which receive the information or being the information sources. At that, for the counter channel arrangement in wireless communication systems, a different frequency band is used (with frequency duplexing) and the same frequency band (with time or code duplexing). In systems of *mobile communications*, the structure of information transmission system most often changes slightly since the terminal (the end device, e.g., the cellular phone) is implemented as a *transceiver* (a transmitter-receiver), when both the transmitter and the receiver locate constructively in the common case. Nevertheless, this system differs principally from radar, which also includes both a transmitter and a receiver implemented in one system, since here we solve the problem, the aim of which is not to extract the transmitted information, but instead the information formation about geographical and kinematic parameters of objects under investigation, based on received information.

Radio systems intended for **information extraction** (the *second type*), say, radars or measuring systems, have the structure shown in Fig. 1.2.

In the radio systems of the second type, at electromagnetic oscillation emission, no information is transmitted by an antenna, but it forms as the result on the investigated object irradiation. The useful information source is the object under

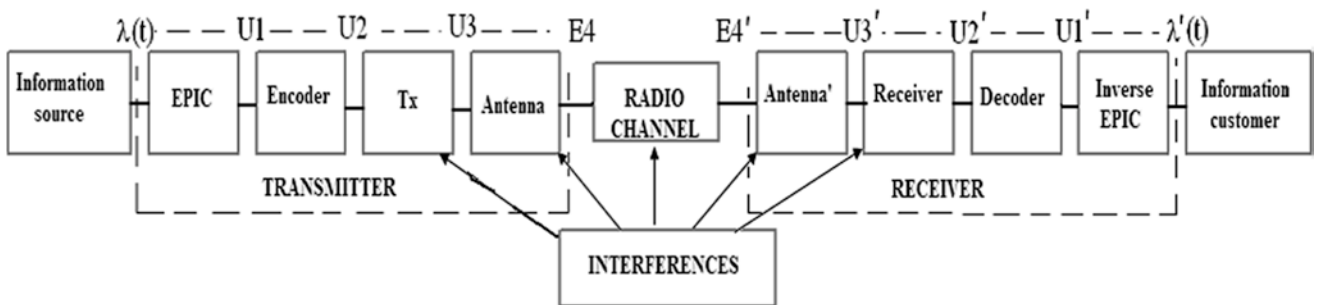
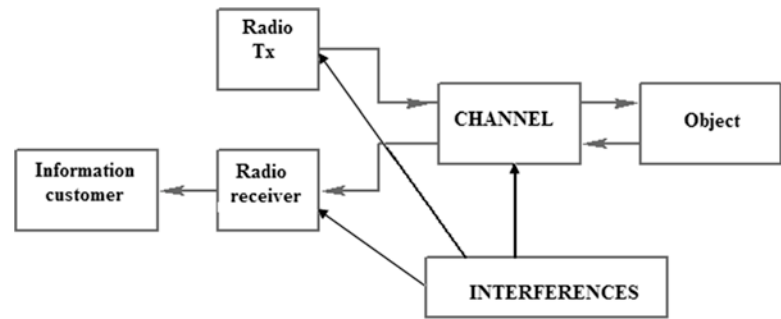


Fig. 1.1 The functional diagram of the radio communication system

Fig. 1.2 The functional diagram of the radar or measuring radio system



examination, which forms the reflected signal, containing information about the range, the velocity, and coordinates of the object. The feature of such radio systems is usual utilization of spatial combining of the radio transmitter and the receiver, which operate on the single frequency. The active radar stations usually have the pulse mode and the single antenna for transmission of probing pulse and reception of pulses reflected from the object under investigation. For protection of the radar receiver against the impact of powerful transmitted pulses, we can use pulses with short duration and high-speed antenna switchers, which switch off the receiver during transmission of probing pulses. The reflected signal acts on the receiver input through the dischargers' system, which switches-off the transmitter from an antenna. Reflection from objects located near to radar station or near objects under examination (so-called clutter) may also prevent the normal operation of the receiver. The intended interference (so-called jamming) also may prevent the reception and processing of signals.

Radio systems of the second type may be implemented to provide the passive radar techniques, for example, in radio astronomy, and in this case, the transmitter is not used at all.

Radio systems of the **third type** are used for creation of intended interferences for radio systems of information reception and transmission. These systems are often called the electronic warfare systems. In this case, the receiver is used to reproduce signals of suppressed radio systems with further intended distortions and signal reradiation toward the suppressed station.

Standardization of radio systems according to these purposes is general enough, and taking into consideration the basing construction principles, they can be divided into separate classes, for example, systems of digital radio communications.

The single-channel communication line containing transmitting and receiving equipment and the communication channel is the basing structure at the development of the communication system. Achievement of maximal effectiveness of the communication system is performed by matching the information source, which includes modulation and coding, with the channel and its output's matching with the customer, which provides demodulation and decoding.

1.2 Radio Signal Propagation

The electrical representation of a message, which carries information, is called the **signal**.

According to the recommendations of the International Telecommunication Union (ITU), in the development of radio communication systems, we must use the radio-frequency spectrum divided into ranges from 1 to 12. The modern radio receivers operate in frequency ranges from 5 to 11 (LF-EHF) inclusive (see Table 1.1).

At present [2], the information transmission between points located on the Earth surface is fulfilled most often with the help of *radio signals*.

Radio waves propagation along the Earth surface is defined by its relief and electric-physical properties of the propagation medium, which includes an atmosphere. An atmosphere is rather a heterogeneous medium, which depends on the geographic latitude, the Earth surface character, the Sun activity, etc., conventionally contains three areas that have a different effect on the transmission of radio waves from terrestrial and satellite radio stations:

- *The troposphere* is the lower part of the atmosphere with a thickness 10...18 km, which has the practically constant gas content – oxygen and nitrogen.
- *The stratosphere* extends from the top of the troposphere for about 80 km. Its lower boundary is characterized by constant temperature (of the order of minus 50...60°C), with a further increase to +80°C and a gradual decrease with increasing altitude. The temperature growth is caused by absorption of the ultraviolet part of the Sun emission, by the ozone contained in the stratosphere.
- *The ionosphere* is the vast layer of diluted gas containing ions and free electrons. This layer passing the electromagnetic waves is simultaneously the space mirror, which reflects radio waves falling to it. The ionosphere upper boundary is unstable and can be 60...1500 km, where the gas density is very low and depends on the day time and the year season.

Table 1.1 Ranges of wavelengths and frequencies

Range no	Wavelength ranges		Frequency ranges	
	Name of waves	Wavelengths, m	Name of frequency	Frequencies
4	Myriameter	$10^8 \dots 10^4$	Very low (VLF)	3...30 kHz
5	Kilometer	$10^4 \dots 10^3$	Low (LF)	30...300 kHz
6	Hectometer	$10^3 \dots 10^2$	Medium (MF)	300...3000 kHz
7	Decameter	$10^2 \dots 10$	High (HF)	3...30 MHz
8	Meter	$10 \dots 1$	Very high (VHF)	30...300 MHz
9	Decimeter	$1 \dots 10^{-1}$	Ultra-high (UHF)	300...3000 MHz
10	Centimeter	$10^{-1} \dots 10^{-2}$	Super high (SHF)	3...30 GHz
11	Millimeter	$10^{-2} \dots 10^{-3}$	Extra high (EHF)	30...300 GHz

Although the properties of the ionosphere are defined by many parameters, their relative regularity allows this layer utilization in permanently acting systems of radio communications. Irregular imperfections existing in the ionosphere can be also used or, at least, taken into account in the designing of radio communication system.

Assuming that the transmitter is located in the A point, and the receiving station (the customer) is located in the C point (see Fig. 1.3), we may note that there are some ways for signal transmission to the receiving point.

Surface waves propagating along the Earth surface induce the EMF in soil causing surface currents, which are converted into heat leading to electromagnetic oscillation losses caused by absorption. With growth of radio signal frequency, absorption in soil increases.

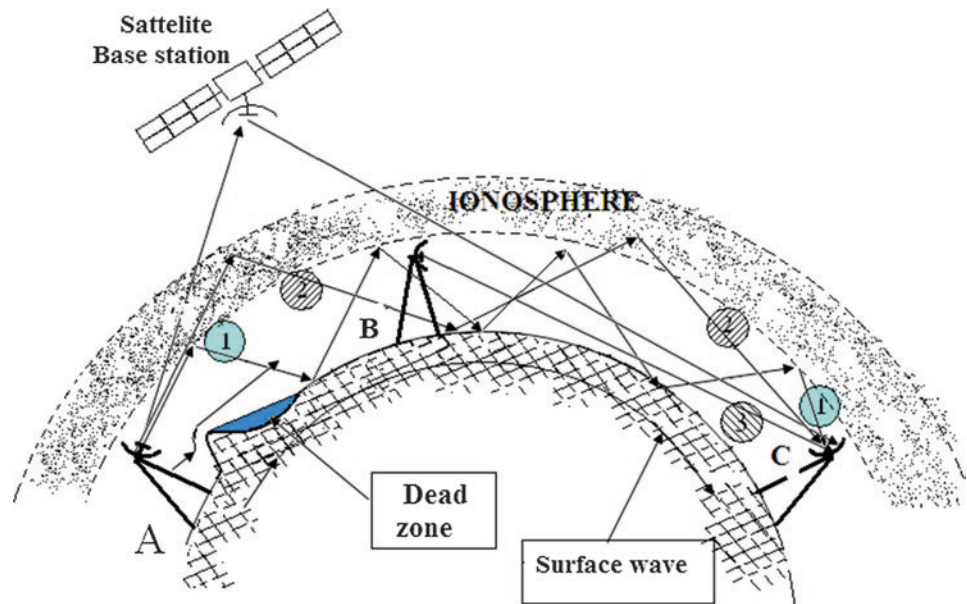
Spatial radio waves of myriameter and kilometer wavelengths (*the range of very low frequencies*, VLF) exceed the roughness of terrain and, due to diffraction, bend around them and, therefore, propagate over long distances. Having rounded a high obstacle, they then propagate in a straight line, creating a “dead zone” with a low signal level behind it (see Fig. 1.3).

If spatial waves propagate toward the ionosphere and reflect from it, the distance of communication essentially increases. However, at reflection from different ionosphere layers (located at various altitudes), the simultaneous reception of signals occurs, which arrived to the reception point by different paths (see ① and ②, Fig. 1.3). This leads to signal summation, which have different phases that may cause their partial compensation (reduction of the power of the resulting input signal) and decrease of the reception truth. The repeated radio signal reflection from the Earth surface may lead to significant growth of communication distance with simultaneous reduction of its stability caused by radio waves’ summation (the interference). The same is observed at summation of spatial and surface radio waves leading particularly to practical mutual suppression of signals for mutual phase shift of two signals closed to 180° . Such an attenuation of signals caused by difference of propagation ways is called fading. Signal reception arrived from the transmitter via various paths is called *multipath*, and distortions (fading) caused by this phenomenon are called *interference*.

Since losses in the soil increase with operation frequency growth in the *medium frequency* range (from 300 to 3 MHz), the radio communication distance in this range does not exceed 1500 km. The spatial waves of this range are intensively absorbed in the daytime in the ionosphere, which decreases at night, and for enough power of the transmitter and sensitivity of the receiver, it is possible to ensure stable reception of signals at a distance of 2000...3000 km. At distance to the receiver, when surface (“near zone”) and spatial (“far zone”) waves have approximately equal intensity, the strong interference fading is possible, which decreases the radio communication quality.

In the *high frequency* range (from 3 to 30 MHz), surface waves do not provide long communication distance more than 100 km due to their intensive absorption. Losses on propagation of surface waves decrease proportionally to the frequency growth, and existence of free electrons in ionized upper layers of the atmosphere leads to its high electric conductivity and hence to high reflection capability. However, instability and irregularity of this ionosphere layer due to “solar wind” influence lead to the appearance of a large number of radio waves reflected in different directions. At that, waves of various wavelengths may reflect from various ionosphere layers and even do not reflect at all at the frequency referred as the “maximal applicable” and higher passing through them. The part of transmitter signal energy may remain in the ionized layer propagating in it, as in the conductor. Some zone forms on the ground surface, where one may receive the reflected signal with quite different intensity, as well as the “dead zone” adjoined to the transmitter, when the surface wave and reflected spatial waves are absent.

Fig. 1.3 The atmosphere structure and radio waves propagation



Features of signal ionosphere propagation are the following: formation of the “beam bench” on the transmitter side, which leads to the appearance of several signals on the receiving side passed the various distances, distorted by the influence of atmospheric inhomogeneities and, accordingly, the formation of EMF from many signals having different phases. Their superposition (interference) at the receiver input creates the signal, intensity of which may change from the external noise level (mutual compensation of obtained signals) to values, which repeatedly exceed the power of the “single-path” signal. In real channels, the signal propagation time and intensity of each beam change slowly during the duration of the single message; therefore, such fading is called slowly.

In *very high* ranges (30 MHz...300 MHz), *ultra-high* (300 MHz...3 GHz) and *super-high frequencies* (3 GHz...30 MHz), the radio signal propagated in straight lines (stations B and C in Fig. 1.3), which allows to realize the radio communication in the ③ region of the direct visibility. At some reflection angles, the penetration of the radio beam through the ionosphere occurs (Fig. 1.3), and then placement of the base station on the orbit (usually near an equator) allows arrangement of several hundreds and even thousands of phone channels on inter-continent communication, which makes this communication very efficient. Application of satellites located on the non-geostationary orbit increases the cost of such line of exploitation caused by equipment complication of the ground-based stations (the Russian communication systems “Horizon,” “Shield”). Contribution into the cost of such lines from the onboard equipment is rather small because base stations may be implemented not only as active (receiver-transmitter) but also as passive detection.

The wide application of mobile communications, when several simultaneously operated transmitters cover the some zone, leads to strong interference in the receiving point, which is caused by useful signal fading as well as effects of adjacent transmitters.

At arising of signal reflections from the ground surface or from the troposphere irregularities at altitudes about 2...3 km, we may provide communication for much larger distances for the sufficient transmitter power. Growth of the communication distance may be provided using local irregularities in the ionosphere caused by meteorite passing through the atmosphere.

1.3 Interferences of Radio Communications

1.3.1 Interfering Impacts

At signal transmission through the radio channel (Fig. 1.1), in which bandwidth is restricted by some administrative and technical conditions, this transmission is exposed by impacts of various nature interferences. An *interference* is any external impact to the receiver, which is not related to the useful signal and prevents its correct reception.

Interferences are generated by external sources and by the sources created in the receiver itself (mainly, thermal noises of the first stages). On the receiving side, this becomes apparent as the noise or clicks, which degrades the speech legibility, the image quality, etc. A random process acts at the input of the receiver, created by a useful signal with digital or analog modulation,

which is subject to interference in the radio channel and other discrete and distributed electromagnetic influences. The total **estimation of noise immunity** at reception in the presence of spurious impacts at analog signals' reception is the *reception truth* and for digital signals *the error probability*, the error frequency per bit (BER – bit error rate). If the interference intensity is comparable to the received signal, this may lead to interruption of the signal reception.

The most typical types of interfering impacts are:

1. **External radio signals.** Since the radio-frequency spectrum is used by different services, such as TV, radio broadcasting, radio communications, radar technology, radio navigation technology, etc., there is their significant mutual influence. To reduce the mutual interference, we can apply various means of organizational (allocation of the strictly definite frequency bandwidth) and technical character: frequency reutilization is allowed only on the remote radio links, application of filters in the output stages of the transmitter and at receiver inputs. On the transmitter side, they restrict the frequency bandwidth of the transmitting signal, while on the receiver side the frequency bandwidth is allocated containing the useful signal. Application of directional antennas on the transmitter and receiver sides reduces the influence of signals arriving from the minor direction.

2. **Spurious emissions of transmitters.** The optimal shape of the signal *energy spectrum*, which is emitted by the transmitter, is the *spectrum of the rectangular shape* occupying the allocated band. This automatically assumes that the modulating signal spectrum is rectangular as well, i.e., it must be infinite in time and having the dependence of the $(\sin x)/x$ type (it leads from the Fourier transform). Evidently, it is impossible at transmission of real signals, and, accordingly, the transmitter output signal spectrum turns out to be more than the allocated bandwidth.

Besides, active elements of the transmitter output stages operate in the nonlinear mode (AB, B, etc. classes) for efficiency growth, which leads to the appearance of the components of modulated oscillations on harmonics multiple to the fundamental frequency (the average frequency of the emitted signal). Harmonics of operation frequency usually attenuate in filtering circuits of the receiver.

In addition to these parasitic frequencies ($f_{nin} = nf_0$, $n \in \mathbb{Z}$), other combination frequencies are emitted, caused by the nonlinearity of the process of converting the oscillations of reference frequency in range synthesizer $f_{comb} = mf_1 \pm nf_2$, $m, n \in \mathbb{Z}$. In cellular communication systems, the mutual influence of several transmitters of the base station operating on closely located antennas leads to the appearance of intermodulation distortions $f_{int} = mf_{01} \pm nf_{02}, \pm \dots \pm pf_{0k}$. Spurious emissions of the transmitter may turn out to be near the bandwidth of the operating frequency range, and therefore, they may be not suppressed and significantly influencing on the adjacent (in frequency) radio links.

3. **Atmosphere interferences** are formed by electric processes in an atmosphere (a lightning) having the spectrum with the highest intensity in LF and MF and less noticeable in the HF range and higher frequencies. Atmosphere interferences can be narrowband and wideband. The single interferences created by the close storm discharges have the pronounced discrete component, while created by the remote discharges have the fluctuation component. Before frequencies approximately 25 MHz, atmospheric interferences exceed the space noise by the level.

4. **Industrial interferences.** Industrial, transport, medical, scientific, domestic, and others electromagnetic emissions usually arise at abrupt variations of the currents in electric circuits. Industrial radio interferences may have the discrete and continuous spectrum and may be attributed to conductive and radiative ones. Interferences caused by medical equipment, by scanning of the electron-beam tubes, by local oscillators in receivers, by TV line scanning generators, by synchronization pulse amplifiers in TVs, and radio communication receivers may be related to the *radiative* interferences, affecting due to the created electromagnetic field.

Interferences created by the electric transport, power transmission lines, and systems for the motor-car ignition can be characterized by the continuous spectrum, since the single pulses and packets of pulses are their nature. Industrial radio interferences can be observed in the bandwidth from units of hertz to units of gigahertz; at that, with the frequency growth their intensity decreases.

The power consumers created the current (voltage) jumps in the power net, forming the *conductive* interferences, which propagate on large distances. Industrial interferences may be removed in places of its origination because if we do not take the measures for suppression, they would make the communication practically impossible.

5. **Space noises.** The reason of the space noise is processes on the Sun as the most close object to us of electromagnetic emission. Thermonuclear reactions happening inside the Sun create the intensive matter emission, which is the charged particle flow, which arises the solar storms in the upper layers of the Earth ionosphere. The most intensive radio emission related to solar spots forms in the meter wave range. The noticeable emissions create some nebulas and constellations, which must be taken into consideration in the development of systems of satellite personal radio communications, satellite TV systems, and systems of the space communications.

6. Internal receiver noises. The active elements of the receiver providing the signal amplification, resistors, and current-conductive circuit-power sources are the *sources* of such noises. In some cases, interferences created by the separate elements may be reduced down to the level when we can neglect their effect. Pulsations of the net voltage and the power supply unit are attributed to these interferences. Voltage pulsations of the power supply unit implemented with utilization of rectifying circuits can be reduced by means of additional filters at its output. The small pulsation level of output voltage in modern power supply units is provided by the preliminary voltage conversion of the primary source (the AC net, the rechargeable battery) into the pulsatory one. The frequency increase of converted voltage up to 2 kHz...4 MHz with the help of a vibrator allows utilization of small values of capacitors in filters at further rectifying, which simultaneously reduces the weight and the volume of the power source.

Fluctuation noises created in the active elements (transistors) and resistors have the essentially lesser level. The reason of fluctuation noises, which are very difficult to reduce, is the thermal motion of charge carriers in resistors and components of the active element structure, which has losses, as well as the generation and recombination processes of electron-hole pairs in semiconductors, the noise of current-distribution, etc. The main contribution in the front-end (RF section) noise factor is given by first stages of the receiver (the input circuit, the low-noise amplifier) since noises created by them are amplified in further stages. The condition of the minimal noise factor obtaining, which provides the maximal sensitivity, is achieved by the AE mode choice, by fulfillment of the antenna and input circuits matching on the minimum of the noise factor or by application of cryogenic cooling (for the requirement of very high sensitivity) of units closely located to the antenna.

The variety of noise impact sources, their frequency-temporal properties, and the character of interaction with the useful signal allow the specification of obtained classification.

7. Electromagnetic interferences (EMI) represent the random processes, which can be divided based on the *type of sources* into natural and artificial. Natural interferences are caused by various physical phenomena in the environment. Artificial EMI are caused by electromagnetic processes in technical devices and can be divided into station, industrial, and contact.

According to the *place of occurrence* in the real communication system, interference can be divided into own, external, inter-, and intra-system.

Own interferences are created by elements of the receiver itself; external interferences lie outside of the receiver; inter-system ones form by sources included in this *system*; intra-system ones form outside the system.

According to *frequency-temporal* properties, EMI can be represented in the form of models:

- Lumped (the interference spectrum is in the narrow bandwidth, usually commensurable with the useful signal bandwidth)
- Pulse (quasi-pulse) representing the non-periodic sequence of the single pulses with non-overlapping responses on this impact
- Fluctuation representing the thermal noises, noises of electronic devices, the total action of interferences from various sources with non-overlapping responses, and the simultaneous impact of many lumped interferences from operating radio stations, as well as space and atmosphere emissions

According to *interaction character* with the useful signal, one can distinguish additive (summing with the signal) and multiplicative (multiplied with useful signal) interferences.

The first category of interferences is the noises of *natural* origin, and they concern to additive interferences.

These are space noises caused by thermonuclear processes in stars, atmospheric interferences, and own noises of the equipment. Atmospheric interferences are the consequence of stormy and electrostatic discharges, and they occupy the frequency band up to 25 MHz and exceed the space noise level. The storm discharges have the spectrum close to the single-pulse spectrum, and superposition of many discharges forms the fluctuation component. Atmosphere interferences may be narrowband and wideband. Such process can be represented as the sum of random sequences of pulses with a certain pulse duration T_p , the repetition frequency F_p , and the random process variance σ_{pr}^2 .

Station EMIs related to *artificial interferences* are created by transmitter emission, the local oscillators (LO) of receivers, etc. Interferences caused by transmitter emission represent the narrowband random processes and have the largest influence. To estimate the station EMI influence, the models based on the theory of random pulse flows are used.

Industrial radio interferences (IRI) are created by technical means using the electrical current energy and can be conductive (propagating through physical power supply circuits and ground connections) and radiative. Industrial radio interferences generated by the medical equipment, by computers, etc. have the discrete spectrum. Interferences from electrical transport, industrial and domestic equipment, and power transmission lines have the continuous spectrum.

Radiative IRIs are the most dangerous in the near emission zone, and their intensity damps according to the $\approx 1/R^{2.2}$ law. Conductive IRIs attenuate slowly and propagate over the significant distances.

Contact interferences arise usually on the moving objects (trains, aircrafts) at impact of transmitter emission on the current-conducting objects located in closest emission zones and having metallic contacts with variable resistance. The secondary emission created in them may seriously affect on the receiver.

Classification of interferences on the base of their frequency-time properties is the most claimed, since the description (at least, on the level of simplified model) of properties of interference sources allows formulation approached and selection of technical methods to fight with them.

1.3.2 Lumped Interferences and Methods to Fight with Them

Methods for interference fighting are based on differences of interferences and signal characteristics.

Difference in *frequency spectra* allows separation of signals and interferences using the frequency-selective networks. At the relatively wide spectrum of interferences with regard to the signal spectrum, frequency selection allows reduction of the interference power passing to the receiver input with extraction of the useful signal.

Difference in *signal and interference phases* is used in devices reacting to oscillation phase (for instance, in synchronous detectors).

Difference in *amplitudes* of signals and interferences lies in the base of device application with amplitude selectivity.

Suppression of Out-of-Band Interferences

The receiver front-end operated practically in the linear mode at small levels of the signal and interferences, and one can get rid of the out-of-band lumped interference with the help of frequency-selective network (Fig. 1.4).

The out-of-band interference is fully suppressed by the ideal pass-band filter with the Π -shaped amplitude-frequency characteristic (AFC), and for interference suppression with partial or full spectrum overlapping, it is necessary to use devices with amplitude or phase selectivity.

At input action of significant (in levels) out-of-band interferences, active elements, for example, including in the RF amplifiers (RFA), become the frequency converters, which create the complex spectrum of the output current, containing (besides harmonics of the signal frequency ω_s and interference frequency $m\omega_{int}$) the combination components with $l\omega_s \pm m\omega_{int}$, which may be in the operating frequency band.

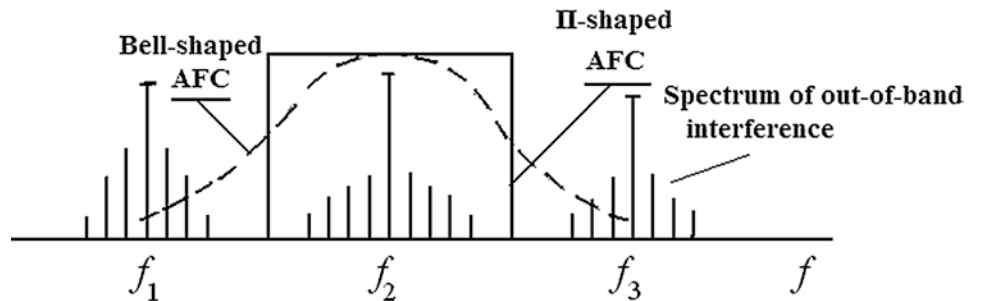
We can achieve the reduction of lumped interference impact on the reception quality applying the high-selective filters at the receiver input or increasing of volt-ampere characteristics (VAC) of amplifying elements (the dynamic range of amplifiers).

The increase of VAC linearity can be achieved by the mode choice of RFA, by application of powerful field-effect and bipolar transistors, connection by optimum load impedances (from the nonlinear effects point of view), and utilization of the compensation methods for nonlinear elements. Introduction of negative feedback in the amplifier is often used for nonlinear effect reduction.

If intensive *out-of-band* interferences on frequencies ω_{int1} and ω_{int2} act in the receiver input, the combination components with $l\omega_{int1} \pm k\omega_{int2} \approx \omega_s$ may arise in the AE output current, and some of them may fall in the receiver bandwidth creating interferences at useful signal reception. Distortions of such a type are called *intermodulation distortions*.

Attenuation of the lumped interference impact in receivers can be achieved by application of protection devices in the receiver input circuits at near located transmitters. Simultaneously, we can apply the signal frequency selection, optimal filtering, the signal predistortion, the interference compensation, application of integral reception, threshold-reducing approaches of frequency-modulated (FM) signal reception, and robust algorithms of signal processing.

Fig. 1.4 Frequency-selective network



When designing receivers, we seek to get a device possessing high sensitivity, for example, professional receivers fixed communications, which improves energy performance of systems. However, when one operates in an unfavorable electromagnetic environment (closely spaced transmitters, industrial installations, etc.), a large EMF of more than 100 V be created in receiving antenna, which leads to the failure of the RF amplifier. To avoid this, the threshold relays, the rejection filters tuned on the frequency of spurious emission, and the variable attenuators are included in the receiver input. Connection of attenuators with electrically controlled attenuation by 10...40 dB allows not only prevention the preselector active element failure, but extension of the receiver dynamic range.

Suppression of Lumped Interferences

Optimal filtering. The effective method of the fight with lumped interferences is optimal filtering. AFC of the *optimal filter* should be complex-conjugated to the useful signal spectrum at digital signal reception on the background of white Gaussian noises. Such a filter provides *maximal exceeding* of the output signal

$$h_{s \max}^2 = \frac{P_s T_s}{E_n^2}. \quad (1.1)$$

where P_s is the useful signal power, T_s is the signal duration, and E_n^2 is the rms noise EMF.

Difficulties at technical implementation of these filters at complex signal shape often force the use of *quasi-matched* filters. At such filter bandwidth B_{qmf} , which satisfies the condition $B_{\text{qmf}} T_s = 0.4 \dots 1.37$, we have somewhat lesser signal exceeding h_{qmf}^2 at output of the quasi-matched filter compared to optimal filtering $h_{s \max}^2$

$$h_{\text{qmf}}^2 \geq 0.82 h_{s \max}^2 = 0.82 \frac{P_s T_s}{E_n^2}. \quad (1.2)$$

The integral reception. In receivers of mobile communication systems, in which active and passive components are implemented by integral technology, one may apply the *integral* approach to fight with lumped interference, which is based on application of switching LC-circuits as integration devices. Such circuits can be implemented as parallel (integration over the voltage), being connected in the intermediate frequency section, and as sequential (integration over the current), being connected in the postdetector section. Periodical switching of the integration contour forms the dynamic bandwidth B_{eff} , which differs from the static one $B_{0.7}$ at ratio choice

$$B_{\text{eff}} = 0.2 T_s, \quad (1.3)$$

the largest exceeding of the integrator output signal is achieved, and the integrator is equivalent to the matched filter. Equation (1.3) $B_{\text{eff}} = 1.57 B_{0.7}$ means that the effective bandwidth is determined by static bandwidth at 0.7 level.

Compared to the quasi-matched filter, the integral reception provides the signal exceeding:

$$h_{s,i}^2 = 1.22 h_{\text{qmf}}^2 \quad (1.4)$$

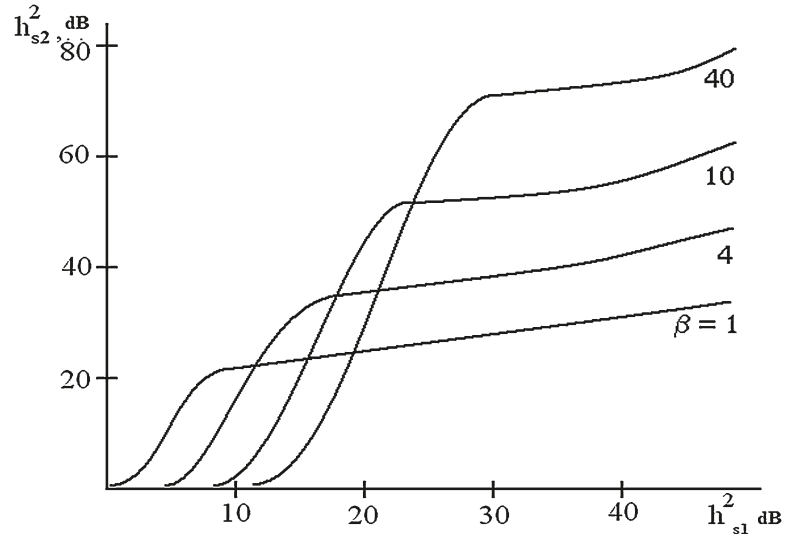
at impact of lengthy interferences $T_{\text{int}} \gg T_s$ where T_{int} is the interference duration.

At action of short-term ($T_{\text{int}} < T_s$) lumped interferences, the integral reception turns out to be more effective in channels with fading compared to the quasi-matched filter. In general, effectiveness of the integral reception is defined by the synchronism of blanking pulse transmission and by the moment of the pulse end.

The Modulation Index In the analogous receivers of VHF and UHF ranges and in the system of mobile communications (in particular, of the GSM standard and trunking communication), the frequency shift keying (FSK), which has the higher noise immunity compared to the amplitude modulation (AM), is actively used. At that, the interference level essentially affects the reception quality: interferences with low intensity cause insignificant fluctuations of carrier frequency; at low signal-noise ratio (SNR) at the receiver input, the instantaneous frequency deviation may be significant, which leads to voltage jump appearance at the demodulator output. The pulse character of the output voltage, which has the uniform spectrum, increases the noise voltage level in the LF region, decreasing the reception quality (the $\text{SNR} = P_s/P_n$ decreases, where P_n is the noise power). Typical characteristics of signal exceeding variations at the demodulator output (the ratio of the peak values of the signal and interference at the receiver output) h_{s2}^2 versus the input value (the ratio of the peak values of the signal and interference at the receiver input) h_{s1}^2 for different values of the β modulation index are shown in Fig. 1.5.

Application of the wideband modulation (with the large modulation index β) at large SHR at input (more than 10 dB for the fluctuation interference impact and more than 16 dB for the pulse interference) gives the great benefit compared to the amplitude one. This allows to get the high exceeding over the interference at their small level in the input. At small input SNR

Fig. 1.5 Typical characteristics of output versus input exceeding variations



(less than 10 dB for fluctuation interference and 16 dB for pulse interference), the wideband modulation has no benefit compared to AM. Nevertheless, for the small modulation index and the same values of SNR, the narrowband modulation ($\beta = 1$) provides the benefit compared to AM. In other words, at high interference level, the narrowband FM provides the slight growth of SNR.

The noise immunity of signals with angular modulation can be improved by the growth of high-frequency spectrum components in the input of the transmitter modulator at their attenuation in the receiver. Predistortions can be introduced at AM application, but FM signals have the triangle distribution of the interference spectrum in the detector output, which permits the SNR increase by 6...12 dB compared to the growth by 3...7 dB for AM.

The characteristics' curvature (Fig. 1.5) occurs smoothly, and the threshold region for large β is approximately $\exp(-h_{s1}^2)$, and then we have the function of $h_{s2}^2 \approx (h_{s1}^2)^\beta$, which gives the signal suppression in the before-threshold region. These curves show that at small h_{s1}^2 for any values of the modulation index β , the SNR decrease occurs in the demodulator output (signal suppression by interference takes place). At $\beta = 10$, the threshold value of h_{s1}^2 is approximately 25 dB, and the value of $h_{s2}^2 \approx 60$ dB, which indicates the SNR increase in the demodulator output (growth of reception truth). For the bandwidth of the linear path B_{lp} and for upper frequency F_{up} of the modulating signal, the approximate value of the threshold exceeding $h_{s,thr}^2$ can be determined as

$$h_{s,thr}^2 = 4 + 4.76 \lg(B_{lp} / 2F_{up}). \quad (1.5)$$

The threshold effect is the most dangerous in radio links with low energy potential at signal fading. At that, increase of the modulation index β and the signal power shifts the threshold region to the right but worsens the electromagnetic compatibility (EMC) of receiving-transmitting devices operating in this range.

The Predistortion Method Reduction of the threshold value $h_{s,thr}^2$ can be achieved by application of predistortion filters in the transmitter and tracking filters in the IF path of receivers. Introduction of predistortion filters is widely used in the satellite broadcasting links operating in the bandwidth, which is common with the ground-based radio relay lines. Filter inclusion on the transmitter side forms the spectrum of the modulating signal, equalizing it in the operating frequency band. For example, the amplitude spectrum of the speech signal in the 400 Hz frequency exceeds the level on the frequency 8...10 kHz by 14...15 dB. At the receiving side, the inverse operation is fulfilled with the help of the correcting filter, which recovers the real amplitude ratio in the spectrum of the modulating signal. In cellular communication systems, the correction of the digital modulation sequence without the return to zero (WRZ) is used on the transmitting side to form the compact spectrum. At that, this correction is performed by the function of the raised cosine (see Chap. 4, [76]).

Inclusion of the function called the dispersion signal and having the spectrum, which adds the initial signal spectrum to the uniform one, are a variety of the predistortion method. Introduction of the dispersion signal into the modulation impact in the transmitter is performed by scramblers, and at reception, it is removed by descramblers.

The Tracking Reception The reception with the *tracking filter* (the Yuzvinsky filter) is usually realized with the help of PLL of the local oscillator, which allows tracking of the instantaneous frequency of the receiving FM signal. The operation prin-

ciple of such a filter is based on the fact that for known instantaneous frequency $\omega(t_0)$ in the time moment t_0 , the current frequency value in the time moment $t = t_0 + \Delta t$ at modulation by the harmonic signal with the frequency $\Omega_m = 2\pi F_m$ with deviation $\Delta\omega_{\text{dev}}$ is

$$\omega(t) = \omega(t_0) + \Delta t d\omega / dt \leq \omega(t_0) + \Delta t \Delta\omega_{\text{dev}} \Omega_m \quad (1.6)$$

It follows from this that the frequency deviation at small $\Delta t \rightarrow 0$ allows application of the narrowband filter, with frequency tuned on the instantaneous frequency $\omega(t)$ with the bandwidth $B_{\text{tr}} \geq 2F_{\text{mod}}$. At non-coherent reception, the bandwidth of the front-end pass-band filter is $B \geq 2(1 + \beta)F_m$ and therefore, the noise level at the demodulator input may decrease by $B/B_{\text{tr}} \approx 1 + \beta$ times. The similar result is obtained at the fixed tuning of the front-end pass-band filter at introduction of the negative feedback loop on the frequency (PLL), providing the LO frequency control with the help of the control voltage, following to frequency variations of the input FM signal.

At transmission of TV signals of even constant brightness, the discrete components with significant level difference (back pulses, line synchronizing signals, the level of white) are in the modulating signal spectrum. At interaction of the useful signal with lumped interferences having the carious spectral structure, the combination components with regions of continuous and discrete spectra arise in the receiver input circuits on the demodulator output. The discrete part of combination components forms a group near frequencies multiple to $|f_s - f_{\text{int}}|$ with essentially different amplitude values depending on modulation indices of the signal and interference, the signal exceeding h_s^2 and the order of combination components.

1.3.3 Pulse Interferences and Methods of Their Influence Reduction

We assume for simplicity (which is close to the real situation) that the pulse interference (PI) is additive, and the front-end is linear (we assume that it is contained from the series-connected tuned circuits, neglecting the AE nonlinearity). The impact of the pulse interference on the first tuned circuit (Fig. 3.1) can be considered as the electric shock leading to capacitor charge with its further discharge and arising of oscillating process on the resonant frequency ω_{res} . The harmonic damping oscillation then affects on the RFA tuned circuit. The more the tuned circuit Q-factor on RF, the narrower its bandwidth is and the slower oscillation damps.

Methods of frequency and time selection as well as the above-considered methods (taking into account its specificity) lie in the basis of methods of PI fighting. The *time selection* is based on the blocking of receiving realization, if the interference appears in the channel. The amplitude selection supposes the IF path construction on the base of wideband-limiter-decision (WLD) circuit (Fig. 1.6).

The limiting level U_{lim} is chosen upper than the total of instantaneous voltage signal and the noise; therefore, if interferences are absent, the receiver operates in the linear amplification mode. The automatic fixing of the level $U_{\text{lim}} = U_{m0}$ where U_{m0} is the maximal overshoot of the total voltage, i.e., the tracking device should have time to react on variations of U_{m0} and not to react at interference action.

Another method of the PI amplitude selection with the deep limitation in the IF path is application of the wideband-limiter-narrowband (WLN) circuit (Fig. 1.7).

In the WLN system, the PI frequency selection is performed by a filter and the amplitude selection – by the limiter. The wider the wideband amplifier bandwidth, the lesser the PI response duration and the higher its level, and so the lesser power of PI at the amplitude limiter (there is limitation from above). After passing through the narrowband amplifier, the PI response growth occurs to values lesser than the limitation threshold from below (at the narrowband amplifier output). This permits to get rid of PI or to reduce essentially its influence.

Fig. 1.6 The IF path construction on the base of WLD circuit



Fig. 1.7 The IF path construction on the base of WLN circuit



The *quasi-optimal filtering*, when the front-end bandwidth (implemented using the simplest filters, for instance, single tuned circuits) is chosen as optimal from the point of view of maximal SNR, is among other methods of PI fighting in professional receivers.

The method of *PI compensation suppression*, when on the base of known PI spectral properties, in the moment of PI appearance in the receiver input, the pulse is formed in the additional channel, which is opposed in phase to the input PI, can be used as a drastic method of PI fighting. After summation of the main channel signal containing the interference and formed PI in the decision device, PI can be completely compensated.

1.3.4 The Fluctuation Interference and Methods of Its Weakening

Fluctuation interferences (FIs) usually relate with the receiving antenna and are caused by out-of-band transmitter emission and noise emission of the receiver local oscillator (which usually is essentially weaker).

In most of cases, FI may be considered as stationary random process, which is always present in the communication channel according to its own nature and cannot be fully removed. FIs cause the signals' masking, suppression, and distortion. Masking due to interference overshoots makes difficulties for distinguishing of acting signals. Interference of signals and noises at definite phase conditions may lead to useful signal suppression.

Weakening of FIs is conducted taking into account its frequency properties: the spectrum is practically infinite, and phases of spectrum components are random.

The *frequency selection* allows reduction of the FI level in the front-end. The front-end has the bandwidth necessary for signal amplification with permitted distortion level and, accordingly, restricts the FI power, which falls in the front-end (Fig. 1.8a).

At action of the discrete signal with the spectrum shown in Fig. 1.8b in the receiver input, the maximal FI attenuation is provided by application of the *matched filter* (MF), which has the comp shape of AFC, and suppresses the FI spectral components in interval between signal spectral components (Fig. 1.8b).

In the general case, AFC of the optimal (with respect to the white noise) matched filter is complex-conjugated with regard to the signal spectrum, and its pulse characteristic is the mirror signal representation. At that, the maximal SNR at output is realized, which exceeds SNR at input by $F_s T_s$ times for the pulse signals. Here F_s is the spectrum width, and T_s is the pulse duration. The signal shape at MF output is defined by the auto-correlation function (ACF) (at present of FI only).

The filter fulfilling filtering, which is matched only in bandwidth and provides the slightly lesser SNR at its output, has the significantly simpler implementation.

The reception of discrete signals with known initial phase on the FI background can be effectively fulfilled using correlation or auto-correlation reception, based on utilization of various statistical properties of the signal and FI. However, at correlation reception, the more strict requirements to synchronization are made.

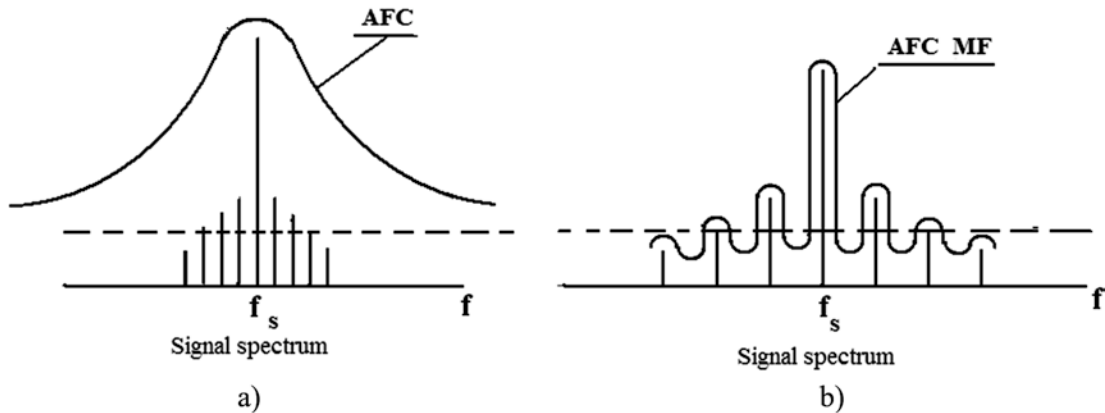


Fig. 1.8 Signal spectrum with the usual filter (a) and the matched comb filter (b)

1.3.5 Multiplicative Interferences

Until now, at analysis of the influence of different types of interferences upon the discrete signal, it was considered that there are not any aftereffects (the response superposition in channels is absent) and scattering of transmitting signals in time, in frequency, or in the space. However, the real channels have a *memory* (processes in the channel are inertial), which can be characterized by a presence of some equivalent reactance (L, C), which narrows the bandwidth and extends transient time.

If the communication system uses the frequency channel division, then scattering in frequency may cause overlapping of frequency spectra and, hence, the mutual interferences between channels.

Irregularities in the signal propagation medium, which give birth to reflected radio waves, and a presence of reactance elements in front-end channels cause the transfer function deviation from the ideal (permanent in transfer factor value in the infinite frequency bandwidth). In the time domain, as it follows from the Fourier transform, this leads to time-extending of the channel response compared to the transmitted signal duration.

The complete description of the *signal time-scattering* in the linear channel gives the pulse characteristic $g(t, \xi)$, which is the channel response to the unit pulse $\delta(t_1 - t + \xi)$, acting at the time moment $t_1 = t - \xi$, where ξ is the scattering parameter (this may be the time delay, the channel characteristic, the quality index). The measure of time-scattering is defined by difference in the channel response shape from the delta-function $\delta(\xi)$ acting in its channel input. The sequence of unit pulses in the channel input leads to mutual superposition of pulse responses, i.e., to the intra-channel interference at its output. Input impact of the real digital sequence causes inter-symbol interference. With growth of the channel bandwidth, we can neglect by the influence of channel superposition pulse responses (the pulse response duration reduces). At condition fulfillment that the signal duration T_s is significantly more than the pulse characteristic duration T_{pch} ($T_s \gg T_{pch}$), then the received signal will have the edge distortions, which can be removed during reception.

Time-scattering of signals under external factors' impact varies in the random manner; therefore, the pulse response $g(t, \xi)$ depends on time and current channel parameters, which leads to conversion similar to signal amplitude modulation.

Growth of the transmitter power does not permit to get rid of multiplicative interferences, since this leads to overload of the modulation section and output stage of the transmitter and gives birth to nonlinear effects in the amplifying path of the receiver, which results in additional interferences for signal reception.

Dielectric irregularities of the propagation medium insert distortion in the plane front of radio waves. At that, a complex picture of interference fading occurs, which depends on time, frequency, and the place of reception. All types of scattering form the multiplicative interference (MI). Fading represents the non-stationary random processes. If distribution laws the radio waves safe in short time intervals, then such processes can be considered as quasi-stationary but with varying values of numerical parameters. Fading in such short intervals is called the fast, and in intervals of mean values of signal levels, they are called the slow. Fast fading is mainly caused by random amplitude and phase relationships of interfering radio waves at multipath propagation. Slow fading is defined by processes of radio wave absorption and scattering in the atmosphere.

Processes determining the fading character form the methods for struggle with them. So, phase-frequency characteristics can be removed by phase correction in the channel, and slow fading can be removed by utilization of several antennas in the receiving or transmitting sides (or simultaneously in the received and transmitter).

The influence of amplitude-frequency factors disturbs the channel response orthogonality, which may lead to loss of transmitted information for deep fading of definite frequencies. Selective fading occurs at arising of multipath operation, since the propagation difference of separate beams creates the phase shift between them, which depends on frequency. Their interference leads to the varying truth of receiving information in random manner. At that, increase of reliability of receiving information, at in-phase summation of different beams transmitting the same information and decrease reliability, at anti-phase summation. This had caused the development of different methods for the architecture of receiving-transmitting equipment, which implements two opposite approaches: elimination of multipath operation and utilization of this phenomenon for signal noise immunity improvement.

Application of pencil-beam receiving antennas allows application of spatial beam separation, which increases the noise immunity. Construction of such antennas in the microwave range with electronic retuning of the pattern in the vertical plane turns out to be relatively simple.

The transmission of signal elements following each other on various adjacent frequencies is used at frequency beam division. This allows separation of spurious responses of the channel with the help of tuned filters. Such a principle requires the frequency reserve increase and inserts complexities of cycling synchronization and EMC problems.

The correlation reception assumes application of multipath phenomenon for in-phase summation of received signals using the pseudo-random processes (PRP) as the message carrier of wideband signals with the following base

$$B_s T_s \gg 1, \quad (1.7)$$

where B_s is the frequency bandwidth occupied by PRP, T_s is the chip (elementary symbol) duration, and $B_s T_s$ is the signal base. Fulfillment of (1.7) during the unit element (chip) duration provides the orthogonality of received signal samples. Extraction of each of samples on the receiving side, coincidence in time, and subsequent weighted coherent combining and non-coherent detection create the noticeable benefit in noise immunity. Each of the components of receiving signal has a stable enough power value during T_s with the lesser fading deepness on the fixed frequency compared to resulting fading created by all signal beams. However, the benefit in noise immunity in the wideband channels leads to reduction of the specific transmission speed. It follows from this that at implementation of such systems, it is necessary to obey the priority of transmission conditions: the channel state, a choice of the signal modulation-demodulation method, and complexities of output (terminal) devices. There are various solutions of the problem of these conditions' optimization through:

- A choice of the signal and modulation type for channels providing the required quality indices (for channels keeping the parameter constancy during duration of the elementary sample)
- The adaptive correction of channel (or signal) parameters on the base of regular estimation of the channel transfer function (the pulse characteristic) and the quality parameter of the received signal
- The estimation of the channel current state and the optimal demodulation method with further decision about the received signal on the base of complex estimation of its indices.

The choice of the specific method of channel and signal parameters' optimization is defined, in general, by complexity of its technical implementation.

1.4 Systems of Fixed and Mobile Communications

Historically, analog radio communications systems assimilated the LF-MF-HF ranges for organization of duplex communications on the "point-point" principle and radio broadcasting systems on the "point-many point" principle which can be attributed to the systems of fixed communications. Further development of radio communication and radio broadcasting systems went in the direction of assimilation of new frequency ranges and increase of the allocated frequency band effectiveness. The necessity of communication distance increase had led to transmitter power growth with keeping of the radial principle of radio communication networks' structure. Development of radio relay, satellite, and decameter communication systems, attributed in the modern classification to systems of the radio excess, preserves their status of the fixed communication systems, since the speed of radial displacement of the one object with respect to another (the receiver – the transmitter) turns out to be insignificant. However, already in 1930s, it became evident that growth of the transmitter power for increase of the communication distance is not the optimal method of communication organization, which gave birth to the alternative way called cellular. Simultaneously with application of transmitters with the essentially lesser consumed power, this increases effectiveness of the allocated frequency band utilization owing to its reuse in another cellular without decrease of communication quality.

A history of the mobile communication development began almost 100 years ago, when, in the beginning of 1920s, there was an attempt to transfer the one-way broadcasting in the 2MHz range into the two-way radio communication. In the beginning of 1930s, there were developed mobile transmitters having high energy consumption and being rather bulky. The first experimental utilizations of the mobile communication systems using the amplitude modulation had shown the high effectiveness of the super-heterodyne type of receivers and typical features of signal passing in such radio channels. Application of the carrier frequency modulation, which has the higher noise immunity under specific conditions of signal propagation, had allowed transferring to wide implementation of the mobile communication systems.

The growth of demands on mobile communication services had led to spectrum overloading and to the methods' search to reduce the frequency bandwidth left to the subscriber and to introduce automatic trunking systems and systems with multiplexing. The possibility for the mobile communication subscribers to enter in the phone public network and to call from fixed communication subscribers to the mobile communication subscribers as well as implementation of their automatic switching had led to the abrupt growth of the channel workload and, hence, to probability increase of the channel blocking (the loss of a possibility to have the connection or to find the free channel).

Since FM implementation led to allocation of 120 kHz bandwidth to transmit speech signal with bandwidth of 3 kHz, the decision was accepted about assimilation of the UHF range near 450 MHz. Application of more advanced FM methods allowed reduction of the bandwidth allocated for one channel to 30 kHz, which essentially increases the analog system effectiveness.

Another important technological advantage was the development of automatic trunking systems or systems with automatic multiplexing, which allowed allocation to the subscriber of not fixed but free channel from the channel group allocated for the direction.

The cellular radio communication (analog and digital), the conception of which was born in the 1940s, is rather the idea on application of known technologies in those time. Instead of utilization of the “radio broadcasting” model, when the one powerful transmitter with the highly lifted antenna supplies the large area, one can apply a large number of low-power transmitters. Each of them supports a small zone called a cell. Each cell can be simplistically represented by a hexagon, and the group of “channel” radio frequencies, which repeats in the zone of network cover, is called the frequency group. The same frequencies can be used in different cells, spaced one from another on the significant distance defined by the cell radius. The cell radius depends on the transmitter power, the antenna height, its pattern, and the terrain topography. The cellular structure of cover zones at achievement of the limited values of the service level allows providing of cell splitting onto smaller ones inscribed onto the previous cell area.

The hand-off control is the principally new property in the cellular communication system. The necessity of such relay-rate transfer of the mobile subscriber relates to the fact that not always the mobile station call will finish during subscriber location inside one cell. Owing to continuous monitoring by the mobile station of the power and other signal parameters, obtained from the base stations of adjacent cells, the base station controller (BSC) of the terrestrial radio access network (RAN) decides to transfer control of the mobile subscriber from the BS of one cell to the BS of another cell in the event of a fall quality of service (QoS) of the subscriber is lower than required. The relay-rate transfer service of the mobile subscriber (handover) is used both in analog and in digital mobile communication systems.

Digital technologies implemented in cellular systems of mobile communication systems of second generation have the indisputable advantages compared to analog systems. The key advantage of digital systems is their capability to support the higher capacity under conditions of strong interferences. Digital methods of signal processing allow formation of the radio signal having the minimal spectrum width and the modulation by orthogonal digital sequences, which provides the high noise immunity of the signal.

Formation of digital sequence from the analog signal at the input of the final stage and further processing methods for digital signals, which increase the noise immunity against spurious influences, is the reason of the wide application of digital methods in mobile communication systems. We may attribute the following to advantages of digital methods:

- Modulation of the current oscillation by the digital signal having the minimal spectrum width and (usually) being orthogonal to the other modulating digital sequences.
- Utilization of methods of nonlinear quantization and digital speech coding provides the low speed of the speech sequence and allows transmission on the one carrier frequency of several signals simultaneously.
- Reduction of capacity losses for transmission of overhead information, which is up to 6% of the total channel number in analog networks.
- Higher noise immunity, which allows provision of higher value of reception truth (lesser value of the error probability) under conditions of strong intra-channel (in the main channel) interference (so-called co-channel interference CCI) and the adjacent channel interference (ACI). Operation in such networks with the required error probability allows reduction of cell size (distance between cells) and simplification of the structure of frequency reuse.
- Flexibly changing of the channel frequency bandwidth.
- Introduction of new services (an authentication, data transmission, the Internet-service, multimedia).
- Increased effectiveness of access control and the call transfer. Decrease of cell sizes leads to the slight load growth on the synchronization channels compared to analog systems, which simplifies the calls processing on access and registration.

1.4.1 The Model of Digital Wireless Communication System

The model of digital communication is presented in Fig. 1.9, where the message source forms the electric time-continuous signal or the pulse sequence.

To transmit the analog signal through the digital communication system, it should be preliminarily converted into the digital form: at first, the time *discretization* of the signal is performed and then the further quantization (discretization on the level) of each sample. This is performed in the *analog-to-discrete converter* (ADC), where the replacement of the quantized sample by the binary sequence of specified length is then fulfilled, in which the interchange of units and zeros is defined by sample amplitude.

At transmission of the telephone message, this operation is fulfilled by the speech encoder (the *source encoder*), which uses the pulse-code modulation (PCM) or the differential pulse-code modulation (DPCM). In accordance with the Kotelnikov theorem (the samples' theorem), the discretization of the standard telephonic channel with maximal frequency $f_{\max} = 3.4$ kHz

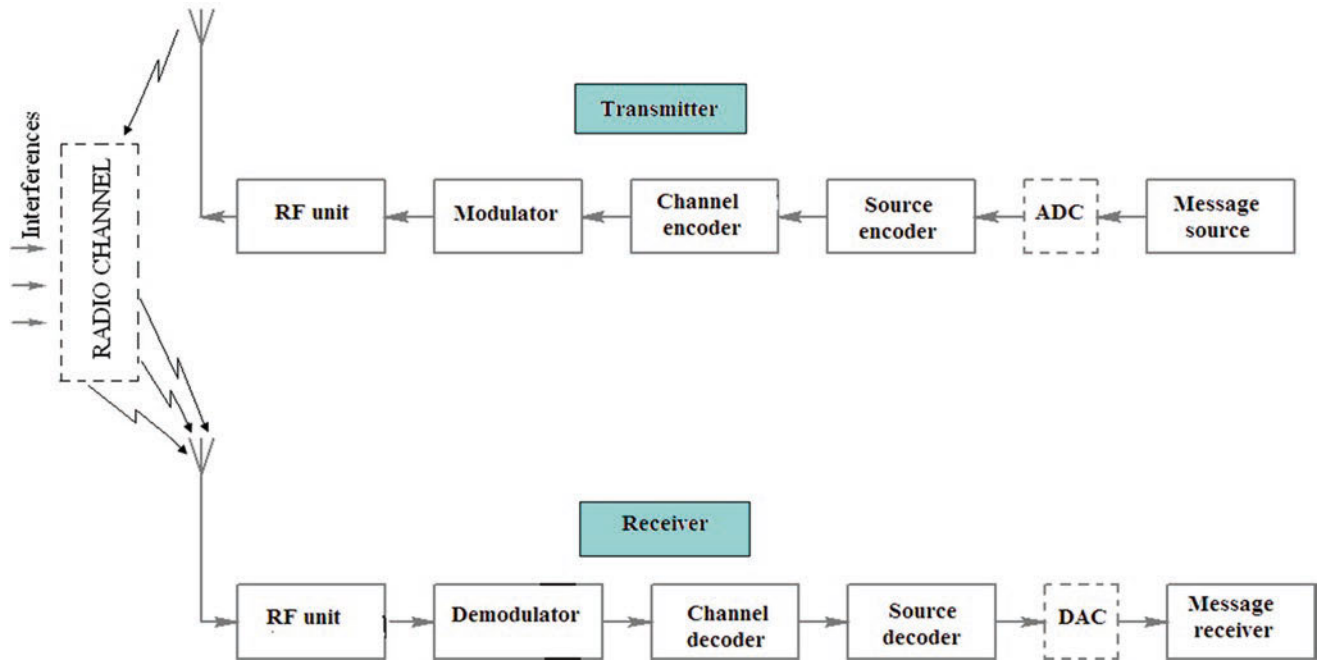


Fig. 1.9 The model of the digital communication system

is performed with the discretization frequency of $f_{\text{dis}} = 8 \text{ kHz}$. Applying the encoder with the nonlinear quantization characteristic (A- or μ -type), the 8-bit sequence is associated with each quantized value of the amplitude. The nonlinear quantizer characteristic takes into account the dynamic properties of the human ear, which allows utilization of the binary sequence with truncated length (8 bits instead 12–13 bits at the linear conversion) for description of the input quantized value of amplitude. The replacement of the real value of the discretized signal in the sample moment by the quantized value leads to the error appearance (the quantization error, the quantization noise) caused by the loss of information portion of the initial signal.

When, for example, the computer terminal is the signal source, then its alphanumeric symbols can be considered as the simplest messages, which are usually represented as 8-bit blocks in the ACSII alphabet.

Interference influence in the communications channel, the multipath reception, signal distortions in the front-end, etc. lead to the appearance of errors at reception (difference between transmitted and recovered binary information on the reception side). For correction or detection of errors, the additional bits (selected by definite manner) are added in the transmission side to binary information blocks in the *channel encoder* (see Fig. 1.9). The bit values are selected by means of congruence addition of two information symbols, thus creating with them the algebraic interconnection, which allows error correction or detection in received sequence. The sequence received with error indication is the directive on recurring transmission of this sequence.

In the *modulator*, which usually has the built-in device for frequency synthesis (or the separate generator of harmonic oscillations), the carrier oscillation parameter (an amplitude, a frequency, a phase) is modulated on the law of the digital sequence acting in its input. The requirement of compact spectrum obtaining in the RF range leads to the necessity of formation of the modulation signal from the digital sequence. That modulation signal must have the narrowband spectrum and simultaneously the high noise immunity. The transfer of information, which is carried by the digital sequence, into the appropriate region of the RF range is the result of modulation.

The *RF unit* of the transmitter is the additional structure providing amplification of the modulated radio signal and solving problems of effectiveness increase of allocated spectrum utilization, and adjacent channels influence reduction, struggle against multipath reception, etc. Operation transfer into the higher frequency part of the RF range and also implementation of new technologies of multi-station access and broadening of the offered service volume (application on IP-protocol for the mobile Internet and multimedia rendering) lead to complicating the signal processing algorithms.

One of the most important requirements presented to the RF unit both of the transmitter and the receiver is reduction of power consumption. This leads to the necessity of active elements of front-end amplifying stages to operate in the mode close to the saturation mode, which increases the nonlinear distortions although decreases the power consumption. Power-

consuming reduction by one of the most power-consuming units allows reduction of requirements to the power supply units and decrease of a number of recharging cycles of batteries of mobile terminals. Simultaneously, the necessity to broaden the amplifier dynamic range is increased. This unambiguously shows the necessity of digital methods of information transmission utilization, while the distortion reduction in the radio channel (including from external factors) can be achieved by modulation method usage with the constant or slightly changing envelope.

Practically, all communication systems with mobile objects are built on the cellular principle. In this case, an *antenna* of the mobile and especially base stations (or several antennas at MIMO technology application) plays a significant role in separating transmission and reception directions, in struggling with transmitter emission from the adjacent cells, in providing the required output power (sensitivity) of the terminal, etc.

Conversions in the *receiver* (see Fig. 1.9) have the inverse correspondence to processes that happened in the transmitter. After amplification and filtering of the received signal in the *RF unit*, the received signal after detection (conversion to the zero frequency) passes to the *demodulator*. The signal formed at its output depends on the type of used modulation and on further processing algorithms. The channel decoder, on the base of the received sequence and addition reserve bits inserted by the transmitter encoder, makes a decision about validity of received sequence or about the necessity of recurring transmission. Reconstruction of transmitted sequence on the base of received on the background of interferences occurs most often using the maximal likelihood method (the Viterbi algorithm) according to the lattice diagram, when the solution is obtained as the result of analysis of elementary signal chains.

1.4.2 Interactions of Open Systems

The Model of the Open Network (OSI) Existence of a large number of equipment manufacturers, which develop units and devices for specific purposes and systems implementing the different methods of network organization and methods for digital and information flow formation, which often use different interfaces, had required the development of the etalon (reference) model of the digital communication model. This model should take into consideration the types of organized channels (physical, logical), the methods of switching in the network (channel switching, packet switching), and the interface levels and its interaction at exchange of overhand information in the digital network structure, as well as foresee the development of offered services.

The description of any network for information transmission was developed by the International Organization for Standardization (IOS) represented in the form of the etalon *model of the open systems interconnection* (ISO), which consists of a set of levels. This model simplifies the procurement of information exchange at the development of various standard networks. At present, the most authoritative organizations, which approve standards, are ISO и IEEE (Institute of Electrical and Electronics Engineers). The leading national and international organizations including the European Telecommunications Standards Institute (ETSI) and the European Conference of Postal and Telecommunications Administrations (CEPT) deal with technical and engineering development of future standards, in which the main idea is adaptation of acting standards of mobile networks to the Internet standard – “all through the Internet protocol.”

Levels of the etalon interconnection model of open systems interact between themselves by means of clearly defined interfaces. At that, at each level the definite number of basing functions is realized, which are defined in the etalon model OSI of the underlying level. Each level presents the set of services to the *above level* without detailed elaboration.

The structure of the *etalon model OSI*, describing interconnection of two systems at transmission of the speech signal, is presented in Fig. 1.10.

Its interaction happens only on the identical level with the help of rules described by the multi-level protocol. The lower hierarchy level is the physical level. It defines electrical and technical characteristics of connections to physical channels as well as procedures of transmission of information symbol flows from one node to another.

The physical level renders the service to the channel level or the data link level, which is responsible for data transmission through the communication channel between two points (communication nodes). The physical level is equipped by jointing modules located at each end of the physical channel and usually acting as modems.

Information packing into frames of the definite length, formation of the check sum and checking of frame content after their transmission, formation of approvals on the frame reception, the automatic request to recurrence, etc. can be attributed to functions of the *channel level*.

The network level provides interaction between the node and the networks. On this level, the packet network addresses are formed, and the control of the flows, addressing, routing, organization, and maintenance of transport connections are performed. The network level defines when it is possible to receive packets from the higher level and when they may be trans-

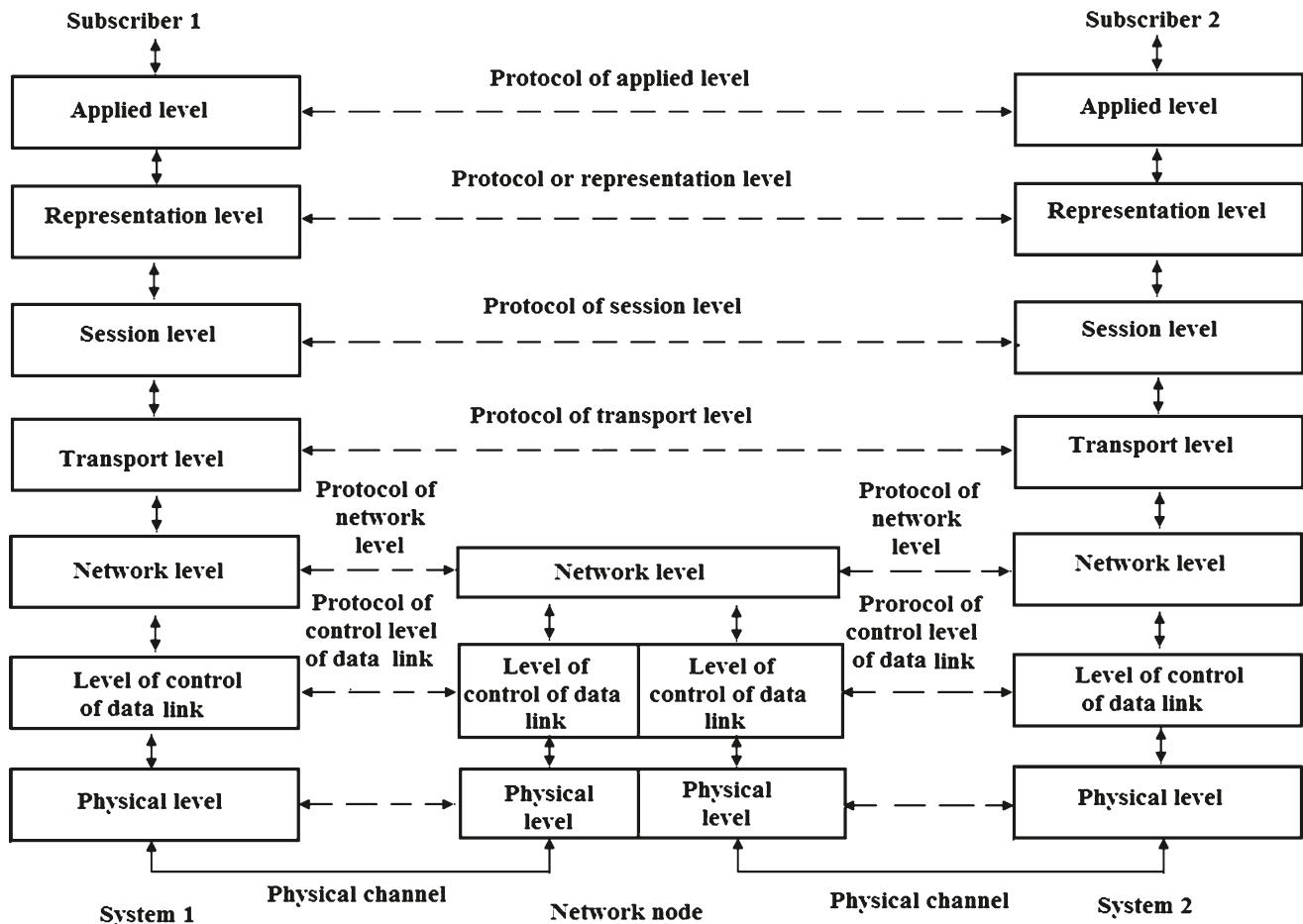


Fig. 1.10 Structure of the etalon model OSI

mitted in another network or the destination node. The packet is the information unit of protocols of the network level; therefore, sometimes one can apply the name of the packet level.

Other levels of OSI model are called the high levels, in which protocols define interaction between systems and do not connect directly with functions of the specific communication system.

The transport level is intended to translation of data flows from one port to another. The port is the logical channel end of data transmission, where operations on data transportation actually finish and computational processes begin. This level checks the error absence in received data, the packet location in the appropriate order, their completeness, etc. The transport connections are established and disconnected at the transport level, and packets are formed belonging to the transmitted communication session.

The session level serves for organization, support, and finishing of processes between applications, which distribute resources between sessions. After logical communication establishment between subscribers, the processes of the session level control the dialog between them.

The representation level realizes the common format of data exchange for two terminals on the base of the digital sequence formation standard. On this level, for known types of digital data, these characteristics are converted into the format, which allows the choice of the services' set on the applied level, for instance, the method of information ciphering.

The applied level is the highest level of the OSI model providing an access to OSI possibilities.

Operation of any digital system begins from the highest level from the point of view of the etalon OSI model. On this level, the applied data block passes from a subscriber, which is transmitted to the lower level adding a heading at each transfer. The heading contains the checking information for realization of this level's protocol. Such a data set is considered (together with the heading) as the data block. On the level of data link, besides the heading, the group of finishing bits (a

trailer) is added intended for error detection and correction. For instance, in the system of error detection with block coding CRC (cycling redundancy check), the computation by the encoder of parity bits lay as the basis.

On the receiving side, the reverse processes occur. At block appearance on the current level, it is unpacked, the heading and the trailer are eliminated, and received information is transmitted to the higher level.

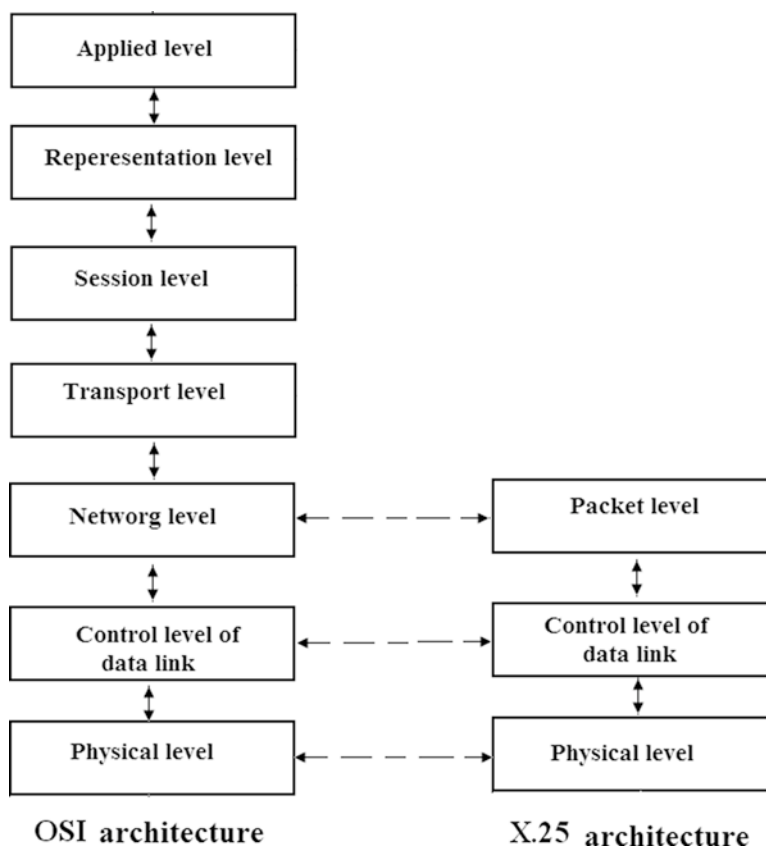
As it follows from the data packet exchange process consideration, which flows on various levels, the study, design, and optimization of the front-end belong to the physical level of the OSI model. *The physical channel is the most important part of the system, the absence of which does not allow the information transmission between two subscribers.* The higher levels only establish the exchange parameters, improve the received signal quality, present additional services, etc.

The X.25 Protocol for Networks with Packet Switching The perfection of the mobile communication networks and presentation of new services (mobile television, radio, multimedia) leads to the wide application of packet data transmission technologies. In the general case, networks of wireless communication may be divided into networks with *channel switching* and networks with *packet switching*.

In radio communication networks with multi-station access with frequency channel division (FDMA), in networks combining the multi-station access with frequency and time channel division (TDMA/FDMA) or with application of the extending sequence (CDMA) with code channel division, the information exchange between subscribers occurs in the some time interval, when the physical connection establishes. The typical example of such a network with *channel switching* is the network of the GSM standard. In networks with channel switching, the physical radio channel is reserved for subscribers during all connection period. However, even in this network, the control information exchange fulfills in the packet mode.

There are radio networks, which fully use the *packet mode*. In such a network, during information exchange, the physical connection between subscribers is absent: subscribers send the numerated data packets having addresses in the network. Packets are sent to the nearest network node, from which they are directed to the destination point on the dynamically appointed route. The network resource is used very effectively compared to the network with channel switching, since the chain of devices between the transmitter and the receiver (the necessary subscriber) is used only at the presence of the packet occurring in the cue.

Fig. 1.11 Protocol X.25 in the OSI architecture



The protocol X.25 is the most often used in networks with packet switching. This protocol has the level structure covering three lower levels of the OSI etalon model (Fig. 1.11).

Since operations of packet formation and addressing are performed with digital sequences created by physical sources, the lower levels of the etalon model are presented in the X.25 protocol.

The *physical level* defines the bit representation form and methods of connection and network-node synchronization establishment. Joint places of different digital and information flows are described by interfaces of electrical, physical, and procedure levels of the subscriber terminal and the network node.

The level of *data link* provides the synchronous operation of the transmitter and the receiver of terminals, as well as provides the error detection and correction.

On the *packet level*, the arrangement of the virtual channel in the network and the packet data transmission is provided.

Taking into account that sources have different bit speeds, the processing order of each signal is defined by LAP-B protocol (link access protocol) on the data link level. The frame structure in the LAP-B protocol begins and ends by the specific 8-bit sequence (a flag) used for receiver synchronization with the received frame. The frame address field, containing the address of transmitting and receiving stations, contains only one byte.

The control field identifies the type of the transmitted (received) frame: information, controlling, or non-numerated. The information field (besides arbitrary number of information bits) contains the CRC-field for the frame check on parity.

The packet level of the X.25 protocol initiates the virtual channel arrangement from the direction, which is defined by the terminal, of the special packet on the network level. This packet contains addresses of the source and destination point, as well as the request of connection realization. After receiving of confirmation by the calling terminal, the virtual duplex channel is considered as connected.

Interruption of information exchange occurs after sending to the network node by any of the terminals of the finishing packet and confirmation of its reception from another terminal connected to the virtual channel.

1.4.3 Methods of Duplexing and the Multi-station Access

1.4.3.1 Duplexing Methods

For systems of fixed and mobile radio and TV broadcasting, the unidirectional information transmission is typical, which turns out to be rather a rare way of digital and mobile system structure. In most cases, we need to organize the double-direction communication, for which it is necessary to create either the feedback channel or the duplex (double-directed) channel.¹

The double-wide line of public switched telephone network (PSTN), which can be used for double-direction communication arrangement, is the duplex channel analog. The possibility of organizing double-directional communication over a two-wire line of the PSTN is provided connecting subscribers through differential systems that implement a bridge circuit made with the use of lumped inductors. In radio communication systems, utilization of the whole frequency range for data transmission is possible, using the method of duplex transmission with the *time channel division* (time division duplex – TDD).

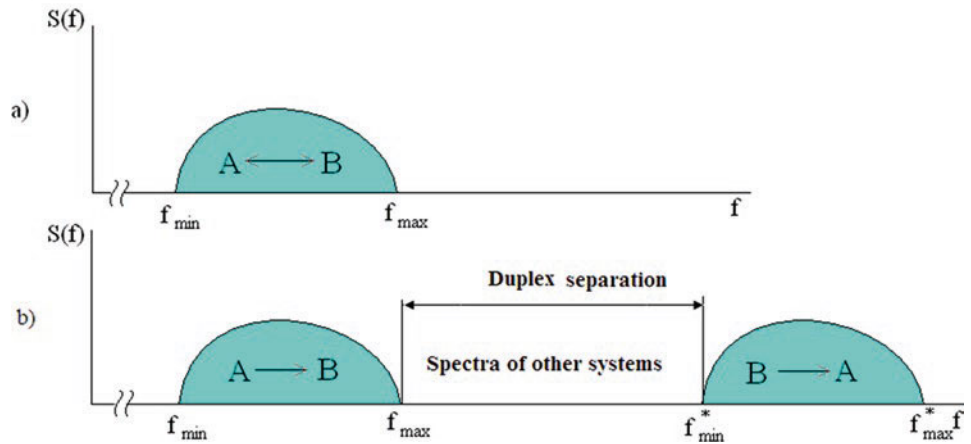
In the TDD method, the frame system is used, when the channel is in turn used for signal transmission and after some time interval for reception in the same frequency band (Fig. 1.12a). Most often this method is used *if signal propagation time is essentially less than the data block length*. Technically, this is implemented with the help of the switcher on MOS transistor in the input RF module (IRFM), which is inserted in the input of the transmitter-receiver (transceiver). To reduce the transmitter influence upon the receiver of the transceiver at direction switching, we can provide for the time protection interval. In systems of complicate structure, for example, local computer networks, for effectiveness increase, we may vary the time disposed to reception or transmission in each direction.

When using the *method of duplex signal transmission with frequency division* FDD (frequency division duplex) of transmission direction (Fig. 1.12b), the whole spectrum allocated to this system is divided between two opposite directions. The spectra of transmission direction should not be overlapped. This frequency band called the duplex separation can be used by other systems. Typically, the frequency band allocated for communication in each direction has the same bandwidth. This system of transmission band separation (channels of duplex communications) is similar to four-wire line in PSTN, and the advantage of such a method consists in the larger transmission independence in each direction.

¹ Under the term **channel** here and later, we understand the *radio channel* organized between antennas of the receiver and the transmitter, and under the term *radio front-end* we understand the radio path of the receiver between the antenna and the detector (demodulator), where the signal spectrum is located in the radio range.

Often, when studying *channel coding*, the term channel is understood as a part of the architecture of a radio system between the output of the transmitter encoder and the input of the receiver decoder.

Fig. 1.12 Spectra for duplex transmission with frequency separation (TDD method): (a) for both transmissions in the same frequency band and (b) for transmission via spaced frequency bands



The duplex channel division may be both frequency and temporal, which allows operators to use effectively the frequency resource. On one hand, this broadens operator opportunities; but at the same time, the formation of radio frames for transmitter and the algorithm for processing the received signal becomes more complicated. Therefore at the present time, the frequency division is the most distributed method of organization of cellular communication system and simultaneously more effective compared to TDD, from the point of view of frequency resource utilization. At frequency duplexing, the logical channel construction and signal processing algorithms simplify as well as the network cost decreases.

Usually, we use not individual methods of duplex transmission with time of frequency division of transmission directions, but instead their combination, taking into account used methods of multi-station access in the system (the approach to channel combining), for example, in the GSM system, TDMA/FDMA with FDD; in the DECT system, TDMA/FDMA with TDD; and in UMTS WCDMA system, CDMA with FDD, UMTS WCDMA with TDD, and TDMA/CDMA with TDD.

1.4.3.2 The Multi-station Access

The multi-station access (the system of channel combining) to the communication channels defines the total resource distribution method (for instance, the electromagnetic spectrum) between two ("point-point") or more ("point-many points," "many points-many points") subscribers. In wireless communication, the *frequency range is the total resource* allocated by the Administrative Agency: in the USA, the Federal Communications Commission; in Europe, the European Commission of Postal and Communication Agencies; and in Russia, the State Committee on Radio Frequencies of Russian Federation.

The most important issue for the development of communication systems is the effectiveness of allocated frequency resource utilization to obtain the communication services by the maximally possible number of subscribers. For this, the frequency range is divided into channels, which gives an opportunity to use communication services simultaneously by a large number of subscribers, which is possible for elimination of mutual influence of channels. This is performed by signal choice of separate channels, which decreases their mutual influence to the permissible level. On the receiving side, the task of channels division and useful signal extraction from the mixture passing through dedicated channels is solved by receiving antennas and devices in the front-end structure. The power vector components, which is described by three spatial coordinates and varied in time, acts to antennas, which transfer the EMF to EMF. Such a signal is called *vectorial*, and signals acting in the front-end are called *scalar*.

It is evident that utilization of the vector methods of channel division is based on antenna applications, which have spatial selection. Utilization of signals in the front-end, which have polarization, simplifies the task of vectorial channel division by application of antennas with good polarization characteristics.

Signals acting at the receiver input are random processes, which have the unknown in advance parameters (an amplitude, a phase, a frequency, and other parameters) of the carrier signal and the modulating signal defined by the modulation type and the message character. Utilization of the common resource for transmission signal of many sources simultaneously forms some total signal at the transmitter output called the *grouped signal*. Reconstruction of the useful signal at the receiving end is possible in case if it can be unambiguously determined from the grouped signal, which is achieved by analysis of this signal at some time interval. A possibility to signal separation assumes that measured parameter is linearly coupled with the grouped signal.

The Condition of Signal Orthogonality The grouped signal is formed from the useful signal and the adjacent channel signals, which, at their linear independence, essentially simplifies the procedure of channel separation. *The subset of linearly independent functions* called *the orthogonal functions* turns out to be the most simple from the point of view of the channel separation procedure, and therefore, they have got the maximal distribution in communication technique. Two functions $u_1(t)$ and $u_2(t)$ are called orthogonal at the time interval $t_1 \dots t_2$, if for them expression (1.8) is satisfied:

$$\int_{t_1}^{t_2} u_1(t) u_2(t) dt = 0 \quad (1.8)$$

The strict signals' orthogonality is not often achieved in practice, and therefore, the most often, one can use the quasi-orthogonal signals, of which the degree of orthogonality is defined from the equation

$$\int_{t_1}^{t_2} u_1(t) u_2(t) dt = \frac{1}{E_1 E_2} r \quad (1.9)$$

where E_1 and E_2 are energies of $u_1(t)$ and $u_2(t)$ functions in the interval $t_1 \dots t_2$, relatively, and r is some number, which is called the *normalized correlation coefficient* of signals, that always satisfies to the condition

$$|r| \leq 1 \quad (1.10)$$

The modulus of the correlation coefficient serves as a measure of signal orthogonality, and at $|r| \ll 1$ signals are considered as quasi-orthogonal.

An example of orthogonal signals is a pair of *time-infinite ideal* harmonic signals $u_1(t) = \sqrt{2} \cos \omega t$ and $u_2(t) = \sqrt{2} \sin(\omega t + \varphi)$, each with the unit power, and the energy of these signals is $E = T$, where T is the analysis interval:

$$r = \frac{1}{T} \int_0^T u_1(t) u_2(t) dt = \sin \varphi - \frac{\cos(2\omega T + \varphi) - \cos \varphi}{2\omega T}$$

The orthogonality condition, when $r = 0$, is fulfilled at $\varphi = 0$, which means that the correlation between the cosine and the sine of the same frequency is absent: i.e., the functions are orthogonal. At that, there are no other harmonic oscillations of the same frequency, which are orthogonal to them. If signal frequencies are different $\Delta f = f_1 - f_2 \neq 0$, then the minimal frequency interval, at which the orthogonality between them is saved, is $\Delta f = 1/T$. At the analysis interval T growth, the correlation coefficient decreases, but at fixed frequency separation (different from zero), we can choose from the condition

$$\omega T \gg 1 \quad (1.11)$$

such (large enough) value of T , at which two (sinusoids) harmonic functions can be considered as orthogonal.

Thus, condition (1.11) shows that at small frequency separation of sinusoid *segments* with T duration, when $\Delta f = 1/T$, their spectra are essentially overlapped, and their separation with the help of filters becomes impossible.

There are some basing methods of the multi-station access (they call them also methods of compaction, methods of separation or multiplexing), based on separation between stations of such parameters as the frequency, time, code, and space. The main task of the multi-station access is the allocation to each communication channel of one (or several) mentioned parameters at a minimum of mutual interferences and at adaptation of signal parameters to provide the minimal error probability of the received signal.

The Frequency Channel Division The very old access method (the method of channel compactness) is the *multi-station access with frequency channel division* FDMA (frequency division multiple access) (Fig. 1.13).

Filter application for separation of signals with the overlapped spectra lies in the basis of the *frequency method for channel separation*. Radio signals, which are represented by a sum of a large enough number of harmonic components, have no strictly limited energy spectrum, and to separate them with the help of channel filters is possible with some permissible influence of adjacent channels.

The frequency range allocated to a system is divided into several bands, in which each data transmission between subscribers is arranged [4, 19, 17]. Channel should be narrow enough in order to create the maximal number of channels, but at

that, they should provide the required quality of information transmission. This method is widely used in analog transmission systems and requires the presence in the receiver and in the transmitter of high-quality channel filters and separation of the protective frequency bands, which decrease the mutual spectra overlap of adjacent channels. The steepness slope of filter AFC defines the admissible separation between spectra of adjacent signals on the specified suppression of the spurious signal, and the bandwidth defines the useful signal delay. Filter application for near-laying frequency separation leads to signal delay appearance, and the larger this delay, the higher the suppression degree for the spurious signal.

In spite of the mentioned measures, including the frequency band restriction on the transmitting side, difference in received power levels may lead to penetration of power portion into the adjacent channel, which causes the essential decrease of the reception quality. To eliminate similar phenomena, it is necessary to use the permanent monitoring of the transmitted signal power.

Time Channel Division Another class of orthogonal signals, which are not time-overlapped, allows implementation of the *time channel division* (TDMA, time division multiple access). In this case, the full allocated band can be selected to the subscriber for signal transmission, but only in some periodically repeated time intervals. The structure of such signal ensembles can be described in the following manner (Fig. 1.13): the some time interval (*a frame*) is divided into n non-overlapped sub-intervals (*slots*). Slots are numerated from 1 to n within the periodically repeated frame, and each j -th slot of each frame is allocated for j -th channel signal transmission. Information of each channel is transmitted by periodically followed segments (packets). Signals of each channel occupy their own slot; therefore, they do not intercept in time, i.e., are orthogonal. At signal separation on the receiving side, we must know time positions only, which allow useful signal extraction from the grouped signal with the help of the switcher, which repetition frequency is specified by the clock rate of the synchronization system. The maximal possible number of simultaneously supported subscribers is equal to the number of time intervals in the frame. Factually, the subscriber number is slightly less than the slot number because some of time intervals are allocated for checking, synchronization, and control.

Since usually they use the signal digital form for data exchange, for its obtaining, the analog signal is exposed to the discretization procedure selecting the discretization frequency according to the Kotelnikov theorem with following quantization and replacement of obtained amplitude values by the coding combination. Thus, we must reduce the time interval duration allocated to the one subscriber to ensure the high rate of information exchange, simultaneously increasing the frequency

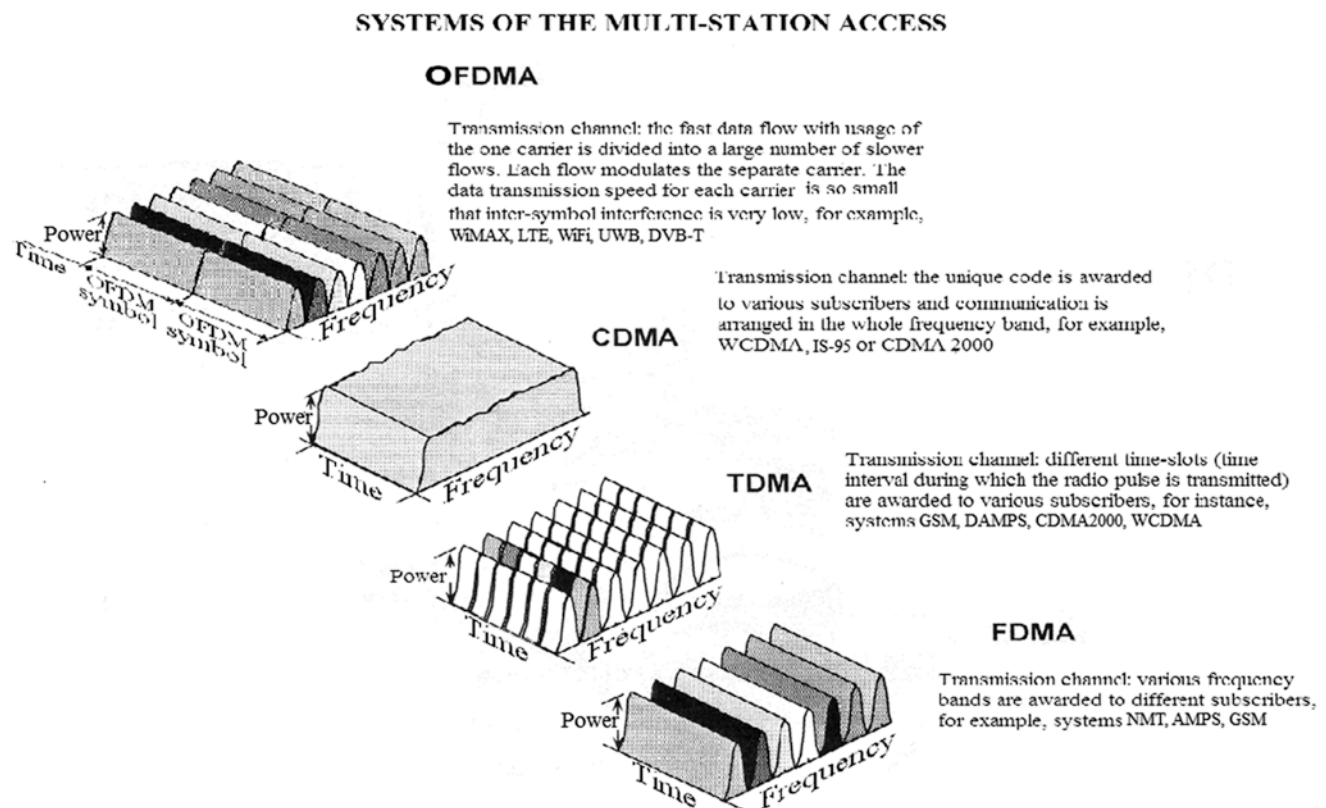


Fig. 1.13 The multi-station access with FDMA, TDMA, CDMA, OFDMA

bandwidth. As in the system with FDMA, it is necessary to create the protecting time intervals between the sequence of time slots, which reduces the mutual channel influence caused by the finite time of on-and-off switching and by the difference in signals' passing time between different subscribers, which are connected to the same base station. In the real frequency systems, one can use the combination of TDMA and FDMA methods (Fig. 1.13) as, for instance, in the systems of the GSM standard.

In the time multiplexing system, the information distribution of each channel in the transmitting side for separate packets, its extraction in the receiving side, and the formation of the continuous information flow occur under control of the synchronization system, which excludes the fall of received signal slots into the adjacent time interval.

The main disadvantage of communication systems with time multiplexing is the loss of communication at synchronization failure due to random or intentional interferences.

Signal Division According to the Form (Code Division) The method of code division multiple access (CDMA) has got a wide implementation [8, 9, 41] in the cellular wideband systems (cdma2000, WCDMA) and in systems of the wideband access (the standard of IS-95). In some cases, the access method led to the appearance of systems with the same name. Signals *orthogonal in form* are completely overlapped in frequency and in time and regarded to as the wideband signals (noise-shaped signal, NSS) (Fig. 1.13). In communication systems realized on this principle, each bit of the digital flow of initial data is replaced by the coding sequence with the length of 8, 16, 32, 64, etc. binary units called *chips*. This sequence is unique for each transmitter, and each of them is mutually orthogonal to any other one (the correlation coefficient is equal to zero). In practice, one achieves the minimal value of the correlation coefficient.

Modulation of the transmitter carrier by such a pseudo-random (expanding) digital sequence creates the noise-like signal (NLS) at its output, in which the spectrum is significantly wider than the initial signal spectrum. Such a method for radio signal formation is called the *method of the direct sequent spread spectrum* (DS-SS). At the receiving side, we know the transmitter code, which signals the receiver should receive. The current information sequence converted in the digital section correlator (in base band) into the digital one is compared with the etalon (reference) prespecified sequence. If the output correlator signal exceeds some specified threshold, then we may consider that 1 is received; if less than this threshold, it means that 0 is received. Transmitter signals with other codes are perceived by the receiver as the additive noise. Owing to large coding redundancy, the signal power at the receiver input is commensurable with the integral noise power. The transmitter can repeat the transmission of each bit of the digital sequence several times to reduce the error probability.

Besides the method of direct sequent spread spectrum, we may use methods of the *frequency hopping spread spectrum* (FH-SS) and the *time hoping spread spectrum* (TH-SS) (or *time jumps*), relatively. In the case of the spread spectrum by means of frequency jumps (frequency hopping), when several transmitters operate in the operating frequency band W , the central frequency f_c of each transmitter changes by a jump in the B_k band ($W \gg B_k$). A law of central frequency change of each transmitter is defined by specified for it the unique code sequence known at the receiving side in the coverage area of each cell. The frequency change character for one transmitter should be orthogonal by another transmitter, which excludes the simultaneous operation of both transmitters and reduces the mutual interference level and the danger of interception and listening. The FH-SS system, in particular, functions in the non-licensed ISM (industry science medicine) frequency range of WLAN network. The physical interface of such systems is characterized by the well distortion immunity, high system capacity, low power consumption, the medium communication distance, and low cost of the radio equipment.

Packet formats in DS-SS and FH-SS interfaces are different.

The *method of time hopping* TH-SS is similar to traditional system with time multiplexing TDMA, but moments of packet translation start are not strictly periodical but varying according to the law specified by the pseudo-random digital spread spectrum sequence. Usually, the code sequence defined the phase (position of the radio pulse leading end) of the translation start of the current packet. Such a method of system construction with the ultra-wideband was implemented by the Time Domain Co.

The signal, which is not useful for the specific terminal, after multiplication with the marker sequence of this terminal, remains wideband since the adjacent channel convolution does not happen, and the energy spectrum is formed at the multiplier's filter output, and this spectrum is defined by fluctuation noises in the channel.

Frequency Division by Means of Orthogonal Carriers The mechanism of *multiplexing by means of orthogonal carriers* (OFDM – orthogonal frequency division multiplexing) is the further development of code and frequency multiplexing. The principle of grouped information signal formation is based on the operation frequency range W separation into large number (up to several thousand) carriers [23, 31, 56, 62, 71, 72]. Several carriers selected from the whole set according to some definite law are allocated to the one radio channel (the transmitter – the receiver). Transmission is simultaneously performed on all carriers: the unified information flow is divided into several N sub-flows, where N is the number of carriers allocated to the transmitter. The carrier distribution can be dynamically changed during operation that essentially increases the flexibility

of the multiplexing mechanism compared, for instance, to FDM. The OFDM multiplexing technology is rather difficult from the point of view of technical implementation, but in connection with the significant progress in semiconductor technology, it is actively used in systems of IEEE802.11 and IEEE802.16 standards in promising networks of the 4th LTE generation.

Spatial Channel Division *The multi-station access with spatial channel division* (SDMA, space division multiplex access) uses the signal division in the space, when each wireless device can perform the transmission (reception) only in some spatial region free from influence of other sources. This method becomes actively developed with the appearance of methods of the adaptive power adjustment of transmitters in the subscriber and base stations as well as with application of antenna systems with the controlled antenna pattern.

Up to now, we discussed structures using the reserve possibilities of various algorithms for digital signal processing, which increase the effectiveness of wireless systems, a possibility for terminals to operate with different standards, etc. At the same time, there are other approaches to improve these indices.

1.4.3.3 The Transmission of Short Messages

When each subscriber has stable information flow (e.g., the digital exchange file or the facsimile transmission), the access methods based on the system reserve utilization turn out to be quite effective. Nevertheless, when we need to transmit information by packets, methods relying on utilization of reserve access methods may turn out to be extremely expensive for transmission of short messages. *Protocols of random access* provide flexible and effective methods of access control to the channel for short message transmission (SMS, short message service). Methods of random access give the freedom for each subscriber to provide the access to the network every time, when he needs to transmit information. As a consequence of such a freedom, these schemes are results of competition between subscribers for access to the network. Usually, used protocols of the random access can be distinguished as the pure ALOHA, the segment ALOHA, and the CSMA/CA.

In the *classic ALOHA scheme*, the some algorithm of the channel extraction is absent: the subscriber can transmit data in any time, which leads to collisions (simultaneous operation of several transmitters) and does not guarantee the data transmission to a recipient. Such a scheme well operates in networks with the weak loading, when the small number of transmitters, which transmit the small quantity of information in the time unit, acts in the network structure. The carrying capacity of such systems achieves a maximum at already 18% loading.

In the basis of the *segment ALOHA*, the algorithm of time division into the intervals-slots, which are equal to transmission time of the one packet, lays. Subscribers are synchronized by time slots, and when any subscriber has a packet, this packet is delayed and transmits in the next time slot. At time synchronization of time slots, the interval of possible collision decreases down to the one packet duration.

The multi-station (network) access with the carrier check (CSMA, carrier-sense multiple access) differs from the classical and segment ALOHA protocols by that the channel is listening by some device. If it is occupied (another device transmits data), then the transmitter transfers into the waiting mode until the channel will be free.

The CSMA/CA protocol (CA, collision avoidance) is the version of CSMA widely used in WLAN networks of the IEEE802.11 standard. Here, the definition of the channel state (busy-free) is chosen in the random manner in some time interval.

The operation principle similar to the CSMA/CA protocol is used in the method of the *multi-station access with detection of suppression* DSMA (digital sense multiple access), in which the channel occupation is defined not by the channel listening before information packet sending, but simply by sending the packet by the base station, in which the channel status is determined.

The most claimed in the modern wireless communication systems are methods of the multi-station access combining the centralized allocation of time intervals and methods of the competitive access. Distributed multiple access (DAMA) system is the example of using of centralized allotted slot method. The base station on the length of the frame allocates, as implemented in satellite communication systems, time intervals (mini-slots) during which all stations generate request for base station to reserve a time slot for data transmission. At that, there are collisions on the reservation stage, which can be resolved by the central station, extracting the successful requests, to which the message is sent with indication of access rights to further time intervals. Requests of other stations are ignored, and they repeat the request procedure on the time reservation. Such a scheme of the access regards to the request distribution method with explicit reservation.

The scheme of the *multi-station access with packets reservation* PRMA (packet reservation multiply access) is the effective variant of the TDMA system, which used the hidden packet reservation. As in the previous case, the base station transmits the list with time interval distribution, and, if there are free intervals, the device can use any of them in the random

manner for data transmission. At successful data transmission, the base station sends a confirmation to the device and reserves it for the new device. At unsuccessful attempt, the device may obtain on demand the reserved channel or to choose again the one of free intervals in the beginning of each cycle.

1.4.4 Systems of Fixed Radio Communication

Historically, the radio communication development began from assimilation of *low and medium frequencies* (LW, MW) for organization of radio broadcasting as the specific case of the radio communication system, when the single-direction communication was developed for transmission of speech, musical monophonic signals with the amplitude modulation application, which has comparably band-narrowity at simple construction of the receiver. On that time, the communication system can be considered as fixed. With development of techniques and assimilation of new frequency ranges, division into fixed and mobile communication became not so evident. For instance, at the end of the twentieth century and at the beginning of the twenty-first century, in Russia, the open system of the cdma2000 standard of the wireless mobile communication IS-95 for services of fixed communication was exploited. The systems of satellite communication, systems of personal radio communication, and systems of personal satellite communication are difficult to describe as systems of fixed or mobile communication. At present, under the term “mobility” we understand only the sensitivity of the receiving terminal to the motion speed, the complexity of its transition from the one service zone to adjacent one without communication interruption, etc., which is mainly determined by the characteristics of the terrestrial radio access network.

1.4.4.1 Decameter Communication

Assimilation of the *HF range* allowed (besides radio broadcasting systems) creation of radio communication systems as the means for main lines (national and international) and zone lines of the mobile communication of different purposes. Radio communication systems in the HF range can not only provide the fixed communication lines (main lines, zone lines, local lines, radio broadcasting, ground-based links of satellite, aviation, naval, etc.) but also arrange communication between ships, aircrafts, and other mobile radio services. Transmission of TV signals in mentioned ranges is impossible because each transmitter frequency spectrum occupies the frequency band of several megahertz. Configuration of HF communication networks can be different (radial, linear, star-shaped, etc.), and it is defined by subscriber requirements and their location in each zone to organize communication between points on “point-point” type.

Development of fixed HF communication of long distances with application of small power transmitters inside the zone is the alternative variant providing noticeable advantage (economic and technical) compared to utilization of huge radio centers. However, due to signal fading at ionosphere propagation and the presence of “dead-spots” (absence of direct visibility between transceivers) and interference influence, in some cases, the reliability of HF communication channel can be not sufficient to provide the required reception truth. Therefore, such communication organization may exist as reserved line for cable (wire), radio relay, or satellite communication system, including for the case when we use the mobile stations as the intermediate transceivers.

1.4.4.2 Broadcasting Networks

Development of the radio broadcasting network in the *very high frequency* (VHF) range led to the creation of stereophonic radio broadcasting networks and TV broadcasting networks. At present, the widely distributed network of cable TV and the actively developed network of wideband mobile communication systems replace the analog TV broadcasting. Transfer to the digital TV, which uses cable and radio networks, will allow essential decrease of the these network cost and the terminals' cost owing to application of multi-functional digital integrated circuits (IC) and simultaneously essential improvement of the received signal quality. The cable networks on the base of fiber-optical lines, which have huge carrying capacities, will very often permit to organize TV programs for wide broadcasting in the 3D format. At present, the ubiquitous implementation of Internet already solves the problem of TV signal transmission in these networks, and the video conference mode becomes the usual phenomenon.

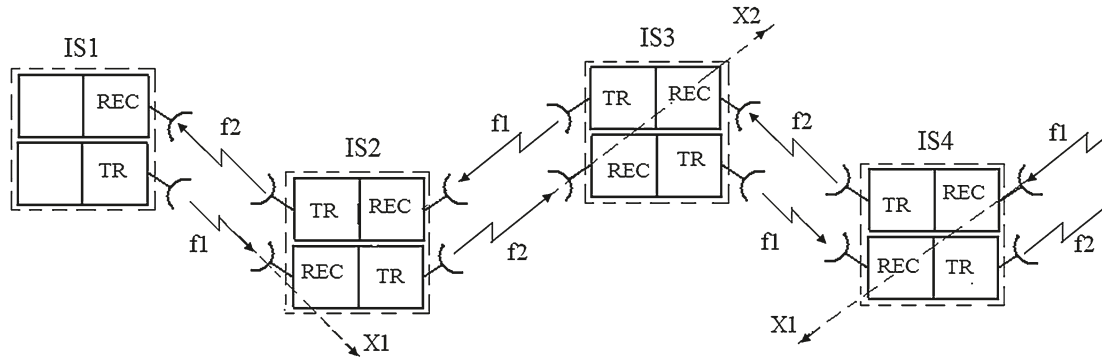


Fig. 1.14 The zigzag location of intermediate radio relay stations

1.4.4.3 Radio Relay Systems

Radio relay systems (RRS) of analogous and digital types are often the components of both fixed and mobile radio communication systems. Digital systems, which have the higher reliability and capacity, replace the analog systems used at the first stage of such systems' development. Traditional RRS application is main lines, infra-zone, and local radio relay lines (RRL). Estimating the potential possibilities of different frequency ranges in RRS, we can notice that for these lines we should use the wide ranges VHF, UHF, and SHF.

The radio wave propagation distance in these ranges is restricted by the direct visibility limits between antennas (cases B and C in Fig. 1.3). We can neglect the surface wave propagation in these ranges together with radio wave diffraction (bending on terrain imperfections). Received radio signals are amplified; if necessary, a part of channels can be extracted to transfer to the subscriber or to correct the received signals to increase the noise immunity. Radio signal propagation within the limits of direct visibility from one side is the factor, which restricts the distance between radio relay stations, but simultaneously allows decrease of the transmitter power (reduce the power consumption by RRS) and reuse of operating frequencies on other parts of RRL. The growth of the antenna height leads to insignificant increase of the distance between stations compared to the cost growth of construction. Depending on the terrain character, the distance between stations may increase by installation of intermediate stations on the natural elevations. Simultaneously, this may lead to signal diffraction and the arising of interferences from signals of adjacent station, if we use the same frequencies for reception and transmission (frequency duplexing). To eliminate interferences of this type, one may use the zigzag location of stations at RRL design (Fig. 1.14).

The signal from the intermediate station IS1 (Fig. 1.14) passes to IS2 and does not pass to the input of IS4 receiver and goes away in the X1 direction; also, the signal from IS2 at frequency f_2 passes to the input of IS3 and does not pass to the receiver of IS4.

To transmit the large information flows (several TV programs, organization of hundreds of telephone channels), several equipment sets are mounted at each radio relay station, which operate in various frequency channels.

At application in the digital RRS of the multi-station access with time channel division, the formation of the digital grouped signal defines the method of the network construction. The European standard (CEPT) of grouped signal formation unifies signals from different sources with various group speeds, with adding of additional bits for speed equalization forming the "plesio-isochronous" digital signal (the E format). The American standard (DS format) forms the grouped signal by synchronous group formation of encoded initial speech channels (the synchronous digital hierarchy). The principle of group signal formation on the base of the synchronous digital hierarchy is more flexible: it provides the direct access to the group signal of any level, in order to coordinate the E and DS formats, to arrange additional service channels owing to larger capacity.

The operating frequency range at RRL design is chosen, first of all, from the EMC conditions, as well as taking into consideration that with frequency growth, the signal absorption increases, and hence, the number of intermediate stations grows.

To organize the network according to the principle "point-many points," the local digital RRS of not-large capacity are built. The digital RRS have got also a wide application in the connecting lines between base stations of mobile communication systems. For the access to the main RRL line or to the fiber-optical communication line, one can build the single- or double-span RRL. The similar structures may be created for interface formation in the mobile communication systems between controllers of base stations and the central switch. RRL is used more and more actively for connection of the local wideband networks for the urgent connection with Internet (for instance, the equipment of the IEEE802.11 standard with 11a/11b/11g protocols with WiFi technology).

1.4.5 Systems of Mobile Communications

The Brief Review of Mobile Communication Systems Development of wireless networks of information transmission was the result of network technology symbiosis (local computer networks), which overgrew in Internet, of packet methods of signal processing, and revolutionary breakthrough in technology caused the creation of high-speed computers and assimilation of new frequency ranges.

In Russia, the main principles of the unified communication network (UCN) were formulated in 1962 by A.A. Kharkevich, and importance of digital methods of transmission and switching was noted for different information types in the digital form. Preconditions to UCN development caused by space investigation development lead to serious successes in computer technologies and to the appearance of satellite communication systems. However, the secretiveness of conducted investigations led in the past to the loss of leading positions in Russia in modern directions of computer technology development and communication engineering. Networks of wireless communication as the component part of the global network of information transmission had got a further development with appearance of the analog networks AMPS (advanced mobile phone service) and NMT (Nordic mobile telephone – the unified cellular communication standard for five North-European countries for speech message transmission). However, in spite of active implementation of analog systems in other countries as well, a series of shortcomings typical for them (the most important among them is that subscriber phones did not operate beyond the covering area of this system) led within the European Union to the development of standards of the unified mobile telephony.

Systems in which the input analog information (a speech, the TV signal, etc.) is initially converted into the discrete (digital) form are regarded as *digital* systems. Thus, input analog signals are digitized and processed by the digital methods in the part of the system (interleaving, scrambling, switching), which is called the digital section (base band), and then they again are converted into the analog form.

A choice of the *digital technology* of *modulating signal* formation together with *digital methods of transmission* is based at construction of all following communication systems with the mobile objects. The main advantages of digital methods used in the mobile communication systems are:

- Digital modulation permitting the efficiency increase of frequency spectrum utilization
- Digital speech coding with low speeds, which together with digital modulation methods allows transmission of several speech channels on the same carrier
- Higher noise immunity, which allows exploitation of digital systems under condition of stronger co-channel interferences and adjacent channel interferences
- The flexibly-varying bandwidth of the front-end, which increases the spectrum utilization effectiveness
- Introduction of new services, such as data transmission, authentication, speech ciphering, and other possibilities of the digital network with integration of ISDN (integrated switched digital network) services
- Growth of effectiveness of the access control and call, which is caused by flexible utilization of the front-end bandwidth but does not lead to essential increase of cell radii at growth of cell number

On the base of the criterion of complexity degree and the volume of rendered services, the mobile communication systems can be divided into:

- Systems of the personal radio call (SPRC)
- Personal networks PAN (personal area network), Bluetooth
- Ultra-wideband (UWB) networks
- Trunking systems (terrestrial trunked radio) TETRA;
- Interactive data entry network (iDEN);
- Systems of mobile (cellular) communication GSM, UMTS, cdma2000, PSC, DCS, LTE
- Systems of personal satellite communication PCS (personal communication system)
- Iridium, Odyssey, GLOBALSTAR
- System of wireless access to the local computer networks (wireless local area network, WLAN)
- WiFi (wireless fidelity)
- Metropolitan area network (MAN),
- WiMAX (worldwide interoperability for microwave access)
- Sensor networks ZigBee
- Global navigation satellite systems (GLONASS, GALILEO, GPS)
- Mobile TV and radio broadcasting systems (ATSC, DVB, EUREKA147, IBOS, DRM, PABIC)

Table 1.2 Main parameters of systems for mobile communication

System, method of multi-station access and duplexing	Frequency range of transmission; commun. channel (up/down), MHz	Number of narrow-band channel, s	Bandwidth of front-end, MHz	Data transmission rate	Receiver sensitivity of the subscriber station, dBm	Dynamic range, dB	Modulation type
1	2	3	4	5	6	7	8
Personal radio call [1]	30; 80; 160; 300; 450; 900	1	5; 6.25; 25 kHz	512; 1200; 2400 kbit/s	10 μ V/m		MSK, O-PSK
GSM900 [18] TDMA/FDMA with FDD	890–915/935–960	124	0.2	9.6 kbit/s (14.4)	–102	140	GMSK
GSM900 , E-GSM [2] TDMA/FDMA with FDD	880–915/935–960	174	0.2	9.6 kbit/s (14.4)	–102	140	GMSK
GSM1800 , DCS [2] TDMA/FDMA with FDD	1710–1785/1805–1880	374	0.2	9.6 kbit/s (14.4)	–102	140	GMSK
GSM1900 , PCS [2, 3] TDMA/FDMA with FDD	1850–1910/1930–1990	299	0.2	9.6 kbit/s (14.4)	–102		GMSK
CDMA2000 [4] UMTS IS-95 with FDD	824–849/869–894	41	1.23	<14.4 kbit/s	–101.1	80	QPSK
WCDMA UMTS with FDD [5]	1920–1980/2110–2170; 1710–1785/1805–1880; 2500–2570/2620–2690		5	3.84 Mchip/s	–126		QPSK
LTE E-UTRAN with FDD LTE Advanced [7] E-UTRAN with FDD with TDD	880–915/925–960; 1710–1785/1805–1880 3400–3500/3510–3600; 3400–3600; 2500–2690	15 kHz per one narrow channel	(1.4; 3; 5 10; 15; 20)	(Mbit/s) 300/75 (1000/500) (Down/uplink)	–102.7		OFDM/(QPSK, QAM). O OFDM/OQAM
Bluetooth , FH/TDD standard: 802.15.1 [8]	2400–2483.5 (ISM) 2445–2475 (Spain) 2446.5–2483.5 (France)	78 22 79/23 22	1	1 Mchip/s			GFSK
UWB , FH/TDD standard: IEEE802.15.3a IEEE802.15.4a [8, 9, 10, 11]	3100–10,600 <1000; 3000–5000; 6000–10,000; 2450	hundreds		1 Mbit/s 400 MBit/s	–75 –70		MB-OFDM CSS
RFID standard: ISO-IECCD 18000-6C. Version 2. Ic2, July.- 2005	860–960		100/200	500 Kbit/s	–70		FM0
DECT [1, 13] MC/TDMA/TDD	1880–1900	10	1.728 MHz	1152 kbit/s	–86		GFSK
WiFi [20, 66] standard: IEEE802.11, 802.11 802.11 g	5.15...5.35 GHz 2400 (2.4...2.483 GHz) 2400	64 3 3	3 22 312.5 kHz 22 MHz 22 22	6; 12;24;36 11;5,5 Mbit/s 2	–76		OFDM/(QAM, DSSS, FHSS, nQAM)

(continued)

Table 1.2 (continued)

System, method of multi-station access and duplexing	Frequency range of transmission; commun. channel (up/down), MHz	Number of narrow-band channel, s	Bandwidth of front-end, MHz	Data transmission rate	Receiver sensitivity of the subscriber station, dBm	Dynamic range, dB	Modulation type
1	2	3	4	5	6	7	8
ZigBee standard: IEEE802.15.4 [19]	868 902–928 2400–2483.5	1 10 16		kbit/s 20 40 250	–98		BPSK BPSK O-QPSK
WiMAX [21, 22] standard: IEEE802.16e	2000–11,000	–	1.25; 5;10;20	Mbit/s 136 (stat); 2 (suppr)	–91		OFDM(QPSK), QAM, OFDM/ nQAM
TETRA , TDMA/FDMA with FDD [164] iDEN , TDMA/FDMA with FDD	385–390/395–399.9; 870–876/915–921; 917–933/872–888 851–866/806–821	240 600	0.025 0.025	2.4– 28.8 kbit/s 64	–103		$\pi/4$ QPSK QAM, 4QAM
INMARSAT Iridium , TDMA/FDMA with TDD; CDMA	1621.35–1626.5 (int.) 19.4–19.6/29.1–29.5 GHz	1100		16 50 2.4; (4.8)			
GLOBALSTAR , UMTS IS-95 with FDD	1610– 1621.35/2483.5–2500; 5091–5250/6875–7055			2.4; 4.8; 9.6 Mbit/s			
Hot Bird	10,719–10,949			27.5 Mbit/s	–136 dBm/m ²		QPSK; 8PSK
GLONASS [16]	L1:1603.6875 L2:1247.3125 L3:1202.025		1.022 (CT); 10.22 (BT)		–161 dBW; –164 dBW		QPSK;
GPS [17]	L1:1575.42; L2:1227.6		20.46	600– 2400 bit/s	–166... –160 dBW		QPSK

Table 1.3 Letter designation of HF ranges

Range	L	S	C	X	Ku	K	Ka	V	W
Frequency band, GHz	0.8–2	2–4	4–8	8–12	12–18	18–27	27–40	40–75	75–110

The main indices of the most distributed systems for mobile communication are presented in Table 1.2.

According to the IEEE (Institute of Electrical and Electronics Engineers) classification, the high frequency range is divided into sub-ranges designated by letters (Table 1.3).

1.4.5.1 Systems of Personal Radio Call

Approximately, before the previous century middle, systems of stationary radio communications, radio and TV broadcasting, were the main and practically the only structures in Russia, which provided the information transmission between subscribers with the help of radio waves. Such systems, together with intensively developed cable systems, effectively function at

present providing the communication services for the main part of Russian population. The necessity of some information transmission to the separate subscriber, which is at motion, led to the development of the personal radio call systems (PRCS).

Systems of personal radio call (paging) are most often the systems of single-direction communications, at which the transmitted message arrives to the subscriber pager, informing him about the necessity to undertake some action or simply informing him about some events. In the most of cases, the double-direction communication is not required which essentially reduces the cost for its organization and allows more effective utilization of the radio-frequency spectrum (the one radio channel supplies the large number of subscribers) and is characterized by lower power consumption, higher ecological compatibility, and a possibility to cover the arbitrary large territories. For different territories, the systems of similar type usually use the meter and decimeter wavelength ranges usually near frequencies 30, 80, 160, 300, 450, and 900 MHz. The zone of reliable communication reduces with the frequency growth, which is caused by the less favorable conditions of signal propagation and by noticeable influence of weather conditions and the terrain relief, etc. However, frequency ranges 450 and 900 MHz are preferable for application in cities with compact planning from concrete buildings, because radio waves in these ranges have good penetration. In addition, the influence of atmosphere and industrial interferences in these ranges is less manifested.

The subscriber of personal radio call systems uses the miniature ringing receiver (pager) having the individual number (address). The call signal is transmitted from the central station transmitter to all pagers; however, only one of them is switched on, which is tuned on the definite frequency and has the appropriated address. The digital systems of personal radio call usually used the frequency carrier modulation, and if the radio broadcasting transmitters with FM are used for a call, then to eliminate mutual interferences, the relative phase shift keying is applied.

Interaction between transmitters is defined by their functional possibilities and used technology of communication organization. When using the echo-repeater technology, the base station emits signals on frequencies and at speeds allowing their reception by pagers and relays (echo-repeaters). After the message packet transmission ends, the base station is switched off by time interval equaled to the packet duration, while relays duplicate the received signal.

The cellular technology assumes relay utilization, which are equipped by microprocessors arranging the current monitoring of signal transmission to other cells and relay efficiency and relay signal exchange with the base station.

The “simulcast” technology is most often used for elimination of interferences from base station signals at overlapping of their service zones, when the signal in the mutual region arrives simultaneously owing to adjustment of the time delay from each base station.

At present, the paging systems have limited enough application. This is caused by the narrow list of rendered services and the violent development of cellular communication systems.

1.4.5.2 Systems of Cellular Communications

The Analogous Systems The development of radio telephone cellular mobile communication systems, which started with implementation of analogous communication systems (NMT, AMPS) functioning in the frequency ranges 450, 800, and 900 MHz with frequency modulation, reveals the typical features of mobile communication system exploitation. The specified frequency range was divided into sub-ranges with width of 25 (30) kHz (in the NMT system using the method of frequency channel division FDMA), while for duplex communication mode (FDD), the frequency range allocated for the downlink transmission (from the base station) was shifted by 10 MHz with respect to the range of received frequencies via uplink. Usually, each base station has the calling channel, the access channel, the combined speech-call channel, and some channels for speech traffic and data transmission. To increase the energy effectiveness of mobile communication networks and the *capacity* increase (the number of subscriber stations per 1 Hz and 1 km², which are supported by the system simultaneously), the supported territory is divided into cells. In connection with the frequency resource restriction, this leads to necessity of frequency reuse in the one cluster and, as a consequence, to arising of co-channel interferences.

Another problem caused by the need to increase the network capacity is reduction of the cell sizes, which leads to necessity of the dynamic power control of mobile stations reducing the co-channel interference level and weakening the influence of distance differences of mobile stations in the cell from the base station. This so-called near-far-effect is caused by imperfection of filters, which extract the separate channels. For near-located mobile stations, the side lobes of the signal spectrum fall into the filter bandwidth, which extracts the adjacent channel with near carrier frequency used by another station. The similar situation may arise at channel organization from the base station. All controlling signals are transmitted with the help of the MSK modulations (frequency shift keying with the minimal shift) directly in the speech channel with interruption of the subscriber signal. The procedure of connection realization, the connection transfer from the cell to another cell (a hando-

ver), and the cell of mobile stations and information upgrade about their location, which were used in the NMT system, were included in the digital GSM system standard, which has got the status of the worldwide system of cellular communication.

There are several reasons that led to the transition from analog systems to digital ones. The most important from them is the restriction of network capacity, which causes trouble in connection support, especially at decrease of the cell radius. When analog systems are exploited, the roaming (the subscriber transfer from one supported zone to another zone, which is located in the supported zone of another reference switcher (network)) is the rather complicated technical problem. This is caused by the standard difference, which defines the operation of mobile communication systems, while in the Scandinavian countries by the roaming prohibition outside this region even for NMT system exploitation, which significantly restricted their supported zone.

GSM The basing version of the GSM-900 [3, 4] system was introduced in the exploitation in the beginning of 1990s, and at this period, the development of its adaptation to the range 1800 MHz began. One of the most important modifications of the GSM system is the version operating in the 1800 MHz DCS-1800 (digital cellular system), and it can be concerned to a category of personal communication systems (PCN). The DCS-1800 system is a modification of the basic version of GSM-900 and does not always exactly correspond to the concept of personal networks, which is based on the ability to receive and initiate a call using a personal terminal regardless of its location, but this modification is close to the PCN category compared to other cellular networks. The system PSC-1900 developed in the USA is the further evolution of such networks. The base and mobile station of these systems have the lower transmitter output power and, hence, lower cell sizes. This leads simultaneously to increase of network capacity compared to the GSM-900 network. Another advantage of the DCS system is the possibility to provide the roaming within the country; even several networks exist in this country.

At *data transmission* in the GSM system, it can be considered as the network of radio access to other telecommunication networks, such as the public switched telephone network (PSTN), the circuit-switched public data network (CSPDN), or with packet switching. At standard bandwidth of 200 kHz allocated by one radio channel, as in speech transmission, this quantity is 9.6 kbit/s or in specific cases 14.4 kbit/s (the full-speed channel). This speed is usually not enough, for instance, for IP applications, which sharply increases the connection time and, hence, the data transmission cost. Depending on the channel state and requirements to service, the different coding schemes can be used, and at small interference level in the channel, the data transmission rate in 171.2 kbit/s can be achieved with combining of the eight time intervals. At such a packet method of traffic organization, the channel extraction is performed only by the reception and transmission time of the packet. In addition, channel asymmetry (the uplink requires a significantly lower transmission rate than uplink) allows other mobile stations to provide the channel during the time slots between the transmission of packets from the active AT.

Nevertheless, even this rate of data transmission is not very high compared with wire connection to the Internet network. The technology of enhanced data rate for global evolution (EDGE) allows increasing of the data transmission rate owing to the application of multi-level shift keying 8-PSK in high-speed modes. Another possibility for data transmission rate increase is the slow stick-slip frequency variation, which helps to eliminate the channel fading. This essentially reduces the co-channel interferences and, hence, increases the total system capacity. The new option, which was absent at speech signal transmission, is the channel quality monitoring; on the base of this information, the combination of modulation type and the channel coding is chosen. The estimation of the channel state allows maximization of the channel resulting capacity and the system as a whole.

Introduction of new services coupled with data transmission has required the increase of data transmission rate in the radio channel of cellular networks and improvement of the transmission characteristics. This commemorates the transition from second to third generation, which unites ground-based and satellite-based systems of personal mobile communication systems into the unified global telecommunication network. The development of programs for the creation of unified fundamental network providing the interaction of various interfaces was performed under the guidance of the International Union of Communication and the European Institute of Communication Standards. The result of cooperation was the appearance of *four standards of cellular communication of the third generation*.

The IMT2000-SC (IMT2000 single carrier) standard of *single-frequency system* with frequency channel division and frequency duplexing. This allows bringing together of GSM and D-AMPS standards, providing the rate increase of data transmission in downlink direction to 384 kbit/s and network creation with packet switching. The appearance of the GSM/EDGE technology with evolution of the GSM base station sub-system into the network GERAN (GSM EDGE radio access network) is the result.

The IMT2000-MC (IMT2000 multicarrier) is the *multi-frequency system* using frequency duplexing of reception and transmission, and it is the evolution of the CDMA2000 standard allowing in the 1.25 MHz bandwidth provision the data transmission with maximal rate of 9.6 (14.4) kbit/c, while at using the variable expansion coefficient allows arrangement of additional channels the transmission rate of 153.6 and 307.2 kbit/s, relatively.

IMT2000-DS (IMT2000 direct spread) is the standard of wideband system with the *direct spectrum spread* and with the frequency duplex. This is European development of the system with code channel division according to the UMTS (universal

mobile telecommunications system) program, known as WCDMA (wideband CDMA), which is called according to UMTS classification UTRA-FDD (UMTS terrestrial radio access, frequency division duplex).

The IMT2000-TC (IMT2000 time code) is the standard of the combined system with *code channel division* and time duplexing in the one frequency band. According to European classification, systems operating in this standard are called UTRA-TDD (UMTS terrestrial radio access, time division duplex). Information reception and transmission are fulfilled in the same band in different time intervals.

The main principle of network construction for the UTRA-FDD standard is to provide the maximal succession of GSM standard possibilities, which is provided by application of the common sub-system of services (the network core). Deployment of the UTRA-TDD network on top of the GSM network allows making common of the main mobility protocols: subscriber connection to the network, authentication of subscribers, their localization, etc. At that, GSM networks, which use the packet technology of GPRS and EDGE data transmission, are included as component parts of 3G (third generation) system. On the current stage of 3G system development, new technologies and services appear: HSDPA (high-speed downlink packet access), development of IP-transport networks in UTRA, utilization of IP-interfaces in GERAN, development of multimedia broadcasting, and organization of priority services.

CDMA, WCDMA (UMTS) In the system with code channel division CDMA, each information pulse is replaced by the binary code combination, which provides the orthogonality of transmitted coded sequences with the help of Walsh functions. This allows the most effective usage of the frequency range allocated to the communication system, in which all channels are transmitted simultaneously providing the highest noise immunity (all channels are transmitted with the same power).

At frequency division, channels of reception and transmission sections are separated by the protecting frequency interval, which reduces the power level of the transmitter signal at input of the receiving section to the value defined by standard requirements on the communication equipment (national or international) [9, 30, 56, 83].

In systems with time duplexing (TDD), the frequency of reception and transmission is the same, while separation of receiving and transmitting sections is performed using the different time intervals (time slots), for which the antenna switch (the *Tx/Rx* switch – transmitter-receiver switch) is used. In transmission more, an antenna is connected to the output of the transmitter power amplifier (during time of the one time slot) while at reception to the input of the low-noise amplifier (LNA) of the receiver.

Wideband systems of mobile communication were initially developed for military applications. From this, the main tasks are followed, which was solved at development of such systems: to hide the fact of digital data transmission, to complicate the procedure of information extraction from the radio signal, and to make difficulties in influence of intentionally created interferences on the received signal truth (the IS-95 standard).

Application of pseudo-random sequences as the signal modulating the harmonic carrier is the basis of solution of mentioned tasks. The pseudo-random sequence is the deterministic periodic digital sequence with the very long period compared to the message duration (the one element of the sequence). Such a sequence is not (and should not be) completely random, as it must be repeated in the receiver. In the ideal random (“white”) sequence with zero mean, the autocorrelation function is equal to zero at nonzero arguments. Hence, to distinguish the subscribers operating in the same radio-frequency range, it is necessary that each from them would have its own sequence, which differs from other sequences. Then, the comparison of the received sequence with any others besides formed by the specific subscriber, which sequence rule is known at the receiving side, will create zero mutual correlation function. In the open systems (CDMA2000, WCDMA), all subscribers use the same pseudo-random periodic sequence, which has for each subscriber the *identifiable time shift*. The most known types of pseudo-random sequences applied in CDMA systems are *m*-sequences, sequences of Gold, Kasami, and Walsh [2, 14].

The main advantages of wideband systems are caused by:

- Reduction of requirement to the decoupling level between beams of the multi-beam antenna (which is typical, for instance, for DECT systems, satellite systems, LTE, etc.). This allows application of the same frequencies in different beams and reduction requirements to EMC with other systems.
- Reduction of multipath influence by application of correlation processing of received signals.
- Application of diversity technique in reception and transmission.
- Application in time interleaving for struggle against fast and medium fading.

LTE The LTE system, which is actively implemented in Russia in commercial exploitation since 2013, at present, can operate in two frequency range: 880...915/925...960 MHz, which is allocated for mutual utilization by systems GSM, of the 2,500...2,690 MHz, which is allocated for mutual utilization by systems of the wideband access WiMAX, MMDS. In spite of potential possibility of utilization of frequency bands with widths 1.4...20 MHz, for the range 2300...2400 MHz by network LTE Advanced, it is permitted using the aggregation (summing) principle of base channels of 1,4; 3; 5; 10; 15; and 20 MHz; it is admitted the summary band up to 100 MHz in transmitting sets of subscriber and base station [31, 46]. LTE

systems depending on allocated frequency range can operate in pair (with frequency duplexing FDD) and non-pair (with time duplexing TDD) frequency ranges. At that, the LTE Advanced provides the downlink transmission rate of 1 Gbit/s, at uplink operation – up to 500 Mbit/s. Frequency ranges of LTE Advanced, GSM-900/1800 network will be used in LTE and LTE Advanced networks as additional at GSM-900.

Potential possibility to expand the channel bandwidth in the LTE systems gives the feeling to subscriber that they possess their “own” wideband channel, although factually they share it with many others. The specific subscriber considers the channel as “own” because the used traffic is usually non-stationary in time and claimed needs usually achieve at peak values of digital flow speed. The capability to dynamic integration of the channel to obtain the wider operating bandwidth or amplification increase is called the trunking effect. Narrowband systems, such as GSM, with channel width of 200 kHz, have no mode allowing to instantaneously render a large capacity, even if there are free channels. Another option of the LTE system is a possibility for the subscriber to formulate requirements on indices of rendered channels (time, traffic rate, etc.).

The features of the LTE system using the OFDM technology are a possibility to change the channel width without changing the base properties of the physical level: spacing of sub-carrier and the symbol length. Application of the different channel width indicates on the system flexibility compared to systems of previous generations.

To optimize systems using the OFDM technology, the subscriber is allowed to choose the modulation type and adjustment of the signal level on each sub-carrier. At that, the strongly damping sub-carriers can be eliminated from processing on the receiving side. Evidently that the adaptive signal processing both at transmitting and receiving sides should be performed taking into account the channel properties (its AFC), which assumes the feedback existence between the transmitter and the receiver.

Achievement of orthogonality between sub-carriers is the most important feature of the OFDM technology. The mutual orthogonality allows partially overlapping of adjacent signal spectra on sub-carriers without arising of interference distortions.

Application of frequency hopping (FH) together with OFDM allows also implementation of wideband systems. Owing to frequency hopping of the carrier oscillation, spread spectrum has some advantages compared to the method using the direct successive spectrum spreading (DS-SS). At that, for instance, the near-far problem is absent, the synchronization becomes easier in the system, the receiver implementation simplifies, etc.

1.4.5.3 Satellite Communication Systems

Universal Multi-functional Systems In accordance with the radio communication regulations [9, 41], depending on the ground-based station type and system destination, we distinguish the following radio communication services:

- The fixed satellite service providing the radio communication between ground-based stations located in the fixed points.
- The mobile satellite communication organized between mobile and fixed ground-based stations with participation of one or several satellites. Depending on the installation place, we distinguish the land, naval, and aircraft mobile satellite services.
- The radio broadcasting satellite service providing the direct TV and radio broadcasting signal reception by the population. We distinguish the joint and individual reception.

Depending on the transmitted information character, we may mark out the universal multi-functional systems, when the ground-based stations (GBS) exchange different information types (Intelsat, Eutelsat, etc.) and specific systems for transmission of one information type or several uniform information types (radio broadcasting systems, the direct TV broadcasting systems).

The satellite mobile communications cover 80% of the land of the Earth surface, while, in spite of the active implementation of the cellular communication systems, they cover only 20% of the land. The most wide application is found by the following satellite communication systems: “Inmarsat,” “Iridium,” “Globstar,” and “Turaia.”

Satellites have the unique possibility to provide the multi-station access mode (many points-point) and radio broadcasting mode (point-many points) for simultaneous transmission of speech, images, and data. Mentioned features can be used in the fixed and mobile communication for modes “many point-many points” for any subscriber group. Such an organization of communication leads to the double-jump transmission mode through satellite relay located on the geostationary orbit (GSO). In this case, satellites are located at altitude of approximately 40,000 km over the equator, which leads to the time delay about 500 ms for double-flight route. Such a delay is acceptable for low-speed data transmission and TV and radio broadcasting, but it is inadmissible for the main duplex speech transmission and for the interactive data transmission.

For small terminals of the mobile personal communication, the fully available network (many points-many points) with essentially less delays at transmission and reception of signals is created. Such a network is provided by the low-orbit (LO) and middle-orbit (MO) satellite systems.

Communication satellites in the wireless satellite communication systems can fulfill the role of international communication nodes providing the expanded possibilities of existing local networks or providing the communication in the mobile service network (e.g., the “Inmarsat” system). Satellite systems providing the direct connection between mobile terminals without participation of the ground-based gateways can be created or the hybrid systems operating in several standards.

The Inmarsat system, covering more than 98% of the Earth surface, uses eight main satellites and one reserve geostationary satellite. In addition, there are 30 ground-based gateway stations providing connection with PSTN or to another gateway station in the case of the token passing of the ground subscriber. The “Inmarsat” system initially was developed with the aim to provide communication with sea ships to achieve the safe navigation and arrangement of the first aid for distressed ships. At present, this system having the high reliability renders the full spectrum of telecommunication services for all markets. The satellite terminal includes an antenna and the talk-back-switching with the total weight approximately 2 kg and dimensions as the standard attaché case.

The Iridium system, applying 66 low-orbit (and 6 reserve) geostationary satellites, covers all Earth surface. Its commercial exploitation had begun in 1998. Besides the speech communication, this system allows providing the data transmission with rates 2.4 and 10 kbit/s. Depending on destination, the portable terminals, which are close enough in dimensions to the cellular communication telephones, are used as well as terminals mounted at sea ships and in aircrafts. Terminals can operate with different standards and systems (GSM, CDMA, AMPS). The “Iridium” system, in contrast to other satellite systems, permits realization of the low-speed global satellite paging.

The deploying system of mobile satellite communication “Globalstar” includes 48 low-orbit satellites providing 80% cover of the Earth surface; at present, it can be considered as regional. This is caused by the roaming absence between the gateway stations, as well as their agency-level belonging. The ideology of this system construction is based on cellular communication methods with mounting of base stations (relays) in the cosmic space. The equipment is represented by “Qualcomm” company developments. In the “Globalstar” system, the noise-like signals and the combination of multi-station access methods with code and frequency channel division are used. The “Globalstar” system provides the speech transmission with mean rate of 2.4 kbit/s and data transmission with the rate from 2.4 to 9.6 kbit/s. The paging and localization possibilities are provided. The geographic subscriber coordinates can be defined either on the demand from the ground-based interface station or by the passive method.

In the “Globalstar” system, the signal processing is not performed on the satellite, and inter-satellite communication lines are absent. The frequency ranges L (1.610...1.626.5 MHz “uplink”) and S (2.483.5...2500 MHz “downlink”) are used for communication with mobile terminals. The onboard complex, containing the receiving and transmitting active phased antenna arrays, forms 16 beams with bandwidth of 16.5 MHz each, providing amplification and pattern adjustment with many-cells ground covering. Each beam contains 13 channels using FDMA, which have the bandwidth of 1.25 MHz. In this frequency band, using the CDMA standard, 127 logic channels are formed using the orthogonal Walsh sequences.

The ground-based segment of the satellite communication system performs the ground network control and the orbital satellite group control. The interface station provides the information exchange in the covered sector, distribution of network resources, and terminal connection to the networks of public use. The parabolic tracking antenna with 5.5 m diameter for transceiver has the gain of 42.5 dB. The own noise temperature reduced to the receiver input is less than 190 K.

The subscriber segment of the interface station includes three types of terminals: stationary, mobile, and portable. Subscribers of stationary terminals operate in the “Globalstar” system only and have no exit into the cellular network or to networks of public use. Mobile terminals consist of handheld apparatus (the phone) or the automobile complex. Portable terminals are close in sizes to the cellular communication phones and can provide operation with the GSM, CDMA, AMPS, and PCS standards, besides the satellite mode.

For communication with the interface stations in the “Globalstar” system, the C range (5.090...5.290 MHz “uplink” and 6.875.95...7.052.9 MHz “downlink”) is used. Effective utilization of the allocated frequency band is provided owing to the frequency reuse and application of pseudo-noise spread orthogonal sequences on the base of functions Walsh. Several types of logic channels for transmission – the pilot-signal, the synchronization signals, the personal call, and information signal – are formed in the “Globalstar” system.

The Arabian zonal regional “Turaia” system of mobile communication uses the network of geostationary satellites for providing the mobile satellite communication. However, its application in the Russia territory is limited at present, which is caused by the presence of the only one gateway station. This system provides the speech and facsimile communication as well as data transmission with the rate of 9.6 kbit/s.

VSAT Station Networks Systems of satellite communication on the base of VSAT (very small aperture terminal) technology began its development from organization of private communication networks for speech message and data transmission

in simplex and duplex modes. Ground-based stations of satellite communication are mounted directly at subscriber place, and their location density in the territory can be rather different. The VSAT stations usually have no permanent maintenance personnel, and monitoring of its operation in the network is performed in the centralized manner. Parabolic antennas with diameter 1.8...3.5 m are mounted on these stations for operation in the frequency ranges allocated for fixed satellite communication: 6/4 GHz (reception/transmission), 14/11...12 GHz, and 20/30 GHz. Radio transmitters of VSAT stations usually have small output power (1...20 W depending on the station type) at information transmission rate not more than 2 Mbit/s.

The structure of VSAT networks can be different: from simplest ("point-point") to fully available. Networks organized on the "star" type, when the radial multi-directional traffic is organized between the central station with high transmitting power and remote peripheral terminals, which have no the mutual traffic, have got the largest distribution. This allows utilization of large-diameter antennas and the large-power transmitters in the central station, as well as the cheap peripheral stations with small antennas without the traffic speed loss (32...48 kbit/s).

In the network of "each-to-each" type, the communication between any stations is provided through a satellite in the "one-jump" manner, which is relevant for networks deployed in the regions, which are difficult of access, at small number of terminals.

Portable stations, which have the automatic mode as typical, include three main components: an antenna system, an outside block mounted directly on the antenna, and the internal block deployed in the subscriber premise. The outside block includes the low-noise amplifier, attached directly to the feeding antenna, and the receiver and transmitter frequency converters (on elements of antenna construction). The internal block includes the satellite modem, a controller, and interface boards allowing connection of the manifold equipment: a facsimile, a phone, a controller, and a personal computer.

Onboard stations are equipped by antennas, with dimensions of 0.5...1.5 m (depending on the frequency range) and solid-state amplifiers with output power of 2...5 W.

The modern ground-based stations provide the subscriber with the wide spectrum of rendered services: from the simple e-mail to arrangement of video conferences and the interactive TV.

1.4.5.4 Systems of On-Air Digital TV Broadcasting

The new stage of TV and radio broadcasting is the transition from transmission of analog signals to creation of digital systems of TV and radio broadcasting on the base of existing broadcasting networks. The digital television (DTV) provides the subscriber with principally new possibilities, which correspond to the digital format, and, first of all, it is an image of excellent quality. High-definition television (HDTV) is a new technology (a new standard for transmission and receiving television signals) with a resolution 1.125 lines, which is twice the value of current one. Transfer to the digital form of analog image and audio signal processing before its transmission via the radio or cable channel should be accompanied by more effective utilization of the frequency range (instead of the one analog channel, we have several digital channels) and by obtaining additional types of services.

The most widespread standards of digital TV broadcasting are ATSC (advanced television systems committee) in the USA and DVB (digital video broadcasting) in Europe. In Russia, we are going to use the DVB standard, which provides the transmission of video and audio data according to the standard on compression and reproduction (MPEG-2) and several qualitative levels. The audio can be coded with the help of other algorithms, for instance, Dolby AC-3.

The European System of Digital TV Broadcasting (DVB) The main tasks for the development of the digital broadcasting system in Europe were increase of the channel number with the standard resolution and development of not-expensive TV receivers [63]. It was assumed that the digital TV network under development will be the basis for the unified telecommunication system, which must transmit data of any nature. Comparing to ATSC, the DVB system turns out to be more flexible, and in the standard bandwidth of 8 MHz it can provide the operation with rates 4.98...31.67 Mbit/s, while the TV program number can be changed from 16 to 2. The DVB system may use channels with bandwidths of 6 and 7 MHz and simultaneously to transmit programs with low resolution (high reliability) or with high resolution (at decreased reliability). The specific feature of the DVB system is a possibility to receive the signal with specified quality at the receiver motion speed up to 300 km/h (a train, the inter-city motor-transport, an ambulance, a police, etc.).

The DVB standard is the set of specifications covering the cable DVB-C, the satellite DVB-S, and the ground-based DVB-T (terrestrial) broadcasting. The most complicate algorithms act at construction of ground-based digital TV broadcasting networks, which is explained by the most rigid operation conditions and requirements to transmission. The ATSC standard adopted in the USA has theoretically higher indices compared to DVB, close to limited ones; in practice, it is the weak competitor. The reason is mainly the necessity of the high SNR value for realization of theoretical indices, which is quite

difficult at using of the multi-position amplitude modulation with the partially suppressed sideband 8-VSB (vestigial side band modulation system for broadcast) in this standard.

The Russian Government adopted the order about preferable implementation of digital broadcasting system of the DVB standard. The base standard of coding and compression of audio and video information MPEG-4 is recommended instead of MPEG-2 Motion Pictures Experts Group, which was adopted till the present time. At present, we need to implement the 4K mode instead of multiplexing mode 2K COFDM (coded orthogonal frequency division multiplexing) of 1705 sub-carriers with development of the element base. Implementation of the second version of the DVB-S2 standard of the satellite digital broadcasting will permit to increase by 30% of channel capacity compared to the first version. The Russian Government Commission adopted in 2011 the DVB-T2 standard for ground-based digital broadcasting, which will lead to the system noise immunity growth and capacity increase.

The system of analogue direct satellite television broadcasting and communication, which existed in Russia since 1999, was replaced in 2005 by system of orbit satellite communication stations for direct digital TV and broadcasting (Orbit, Moscow, Ekran). This made it possible to provide coverage of the entire territory of Russia with satellite digital TV and radio broadcasting network. The development of terrestrial and cable network allows the transition of residents of the entire country to the digital TV and radio broadcasting system in 2017. From 2007, the company “NTV-Plus” began to commercial broadcasting of HDTV channels. The International Regional Radio Communication Commission adopted in 2006 the new frequency plan for systems of ground-based digital TV broadcasting systems DVB-T and audio broadcasting T-DAB and defined the new date for transfer to digital broadcasting: 17 June 2015. The ground-based TV broadcasting should be transferred in digital standards in 2012.

1.4.5.5 Systems of Digital Broadcasting

The record on CD with the 16-bit linear signal digitization with the sample frequency 44,1 kHz is the etalon of the audio-signal quality format. To transmit this digital signal, which has the rate of 1.5 Mbit/s, the signal bandwidth is 1.536 MHz, and the preliminary compression of the digital flow is used (up to 192 kbit/s, for instance, using the Musicam codec), but even in this case, the bandwidth is essentially higher than occupied by the analog signal.

At present, technologies of the digital radio, Eureka-147, IBOC, and DRM, have got the most widespread in the world. The Eureka-147 standard is oriented to wideband single-frequency relay networks, and in some cases it requires the revision of existing frequency distribution. IBOC and DRM systems are intended for operation in frequency ranges, which are used at present in the systems of analog broadcasting; at the nearest future, the DRM standard will be implemented in MW and SW ranges, which will permit to ensure the receiving signal quality comparable with analog FM broadcasting.

The Eureka-147 System The European standard of digital radio broadcasting known also as DAB (digital audio broadcasting) is in exploitation from 1997. Its later variant DAB+ takes into consideration the application features in the transmitter structure of more modern audio codec AAC+.

The procedure of group signal formation is rather traditional for modern communication systems. The audio signal is discretized with the frequency of 24 or 48 kHz and is exposed to the coding procedure with the help of the MPEG audio layer 2 codec. The DAB stream of audio frames is formed in the codec output including 1152 PCM samples at the total frame length of 24 ms for the discretization frequency of 48 kHz. The overhead information and information data (the weather prediction, the name of the program, the currency rate, etc.) is added to the audio frames. The symbolic sequence is exposed to randomization – equalization of the zeros and units number over the frame duration by means of its multiplication by the pseudo-random sequence, which provides the constancy of the average power. The obtained sequence is exposed to the convolution protective coding and then by the time interleaving. After adding of the overhead information (parameters of multiplexing, information about coding, etc.) and synchronization packets, the transmission frame is formed.

The transmission frame includes information from 32 sub-channels with width of 48 kHz each, in which all described operations are already performed. Each sub-channel with mentioned multiplexing parameters and synchronized with other channels is multiplexed with other channels and forms the group signal – the transmission frame. From this, according to the given algorithm, the slow digital sequences are formed, which compose the OFDM symbol. Each from OFDM symbols is separated from other symbols by the protective gap, which provides the mutual orthogonality of symbols and simplifies the fight against the inter-symbol interference caused by multipath properties. Then, the sequence of OFDM symbols is exposed by the quadrature modulation QPSK (quadrature phase shift keying). Each two bits of the OFDM symbol (besides the two first intended for the synchro-channel) correspond to the QPSK symbol described as

$$q_{l,n} = \frac{\sqrt{2}}{2} [(1 - 2p_{l,n}) + i(1 - 2p_{l,n+k})],$$

Table 1.4 Parameters of the transmission modes of Eureka-147 (DAB)

Transmission mode	I	II	III	IV
Frame duration, ms	96	24	24	24
Duration of null symbol, μ s	1297	324	168	648
Duration of the protective interval, μ s	246	62	31	123
Nominal maximal distance between transmitters, km	96	24	12	48
Nominal frequency range, MHz	to 375	to 1500	to 3000	to 1500
Duration of symbol informational part, μ s	1000	250	125	500
Total duration of symbol, μ s	1246	312	156	623
Number of orthogonal sub-carriers	1536	384	193	768

where $n = 0, 1, \dots, K-1$ is the carrier number; $l = 2, \dots, L-1$ is the current number of the OFDM symbol; K is the number of orthogonal symbols; and $(p_{l,n} - n)$ is the bit of l -th OFDM symbol. This is equivalent to the expansion of two bits into the quadrature components.

Then QPSK symbols are distributed according to some algorithm over K sub-carriers, which essentially increases the stability of informational blocks to the effect of the narrowband interferences. The interference influence damages only the small part of pulses distributed over the large number of sub-carriers, and single errors can be easily detected and eliminated by correcting codes.

Depending upon the accepted technology of modulated orthogonal sub-carrier formation, the circuitry of the software-algorithmic approaches with the help of the Fourier transform, the spectrum transfer of slow modulating sequences onto intermediate frequencies (sub-carriers), is executed. At the same time, in the quadrature QPSK modulator, the shift of all sub-carriers by $\pi/4$ is performed, and for information transmission to the receiver about the structure variation of the signal constellation, two symbols are allocated in the audio frame structure, which reflect the positions of the state vector in the complex plane.

The DAB standard describes requirements to the digital radio broadcasting network construction of various purposes: ground-based, cable-based, and satellite-based. Depending on the network destination, we anticipate four transmission modes providing the optimal structure and the duration of the transmission frame (Table 1.4)

This standard anticipates the formation on the physical level of three logical channels, which combine the group of executing functions:

1. The main service channel (MSC) intended for data and audio signal transmission.
2. The fast information channel (FIC) providing the transmission of the overhead and auxiliary information.
3. The synchronization channel providing the receiver and transmitter frame synchronization, the frequency control, the check of the channel state, and the transmitter identification.

As we see from Table 1.4, for the audio frame transmission, which is formed in the form of the OFDM symbol sequence, two of them form the sequence of the synchronization channel: first is the null symbol (always zero), and the second symbol carries information about the phase of the QPSK signal at the modulator output. Informational sequences (from MSC and FIC channels) in the arms of the quadrature modulator are converted into the symbol sequences, each of which includes two bits and is distributed over the OFDM symbols. Depending on the service zone size, we choose the duration of the protective interval and the number of OFDM symbols in packets of MSC and FIC channels.

Parameters of the audio signal structure are chosen by the optimal manner for the radio signal formation in the ground-based networks of UHF range (modes I, II, IV). Mode III is optimal for radio signal transmission in the cable and satellite broadcasting networks, as well as in the hybrid satellite-ground networks. Modes II and IV are optimal for operation in the local networks of the L range and in the single-frequency global networks.

In any of the mentioned modes, the occupied frequency band is 1536 MHz at the digital stream rate of 1.5 Mbit/s providing the transmission of mono-, double-channel mono- and stereo and joint-stereo signals.

IBOC Technology The main problem for Eureka-147 implementation is the occupation of the radio spectrum, on which this standard is oriented. In Russia, for UHF radio broadcasting, the ranges 65.8, ..., 73 MHz and 87.5, ..., 108 MHz are defined by standards and are reserved by FM radio stations for military and other needs. We do not expect of quick release of frequency ranges 30 MHz, ..., 3 GHz, and as a consequence of it, in the USA the IBOC (in band on channel) system is implemented, which allows the transfer from the analog radio broadcasting to the digital without allocation of additional channels. For two ranges allocated for radio broadcasting with AM (LW, MW, HW) and the UHF range with FM, the two

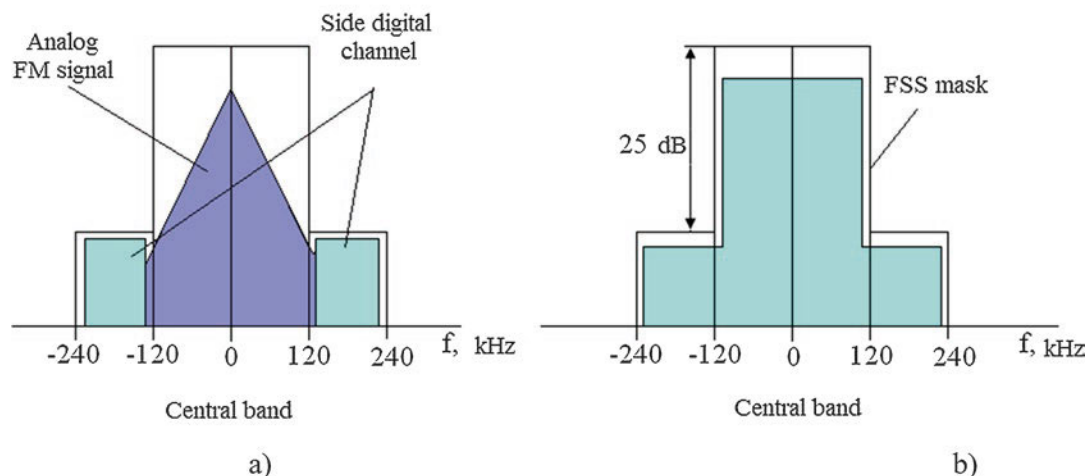


Fig. 1.15 The spectrum of the analog FM signal (a) and the signal mask according to FSS (b)

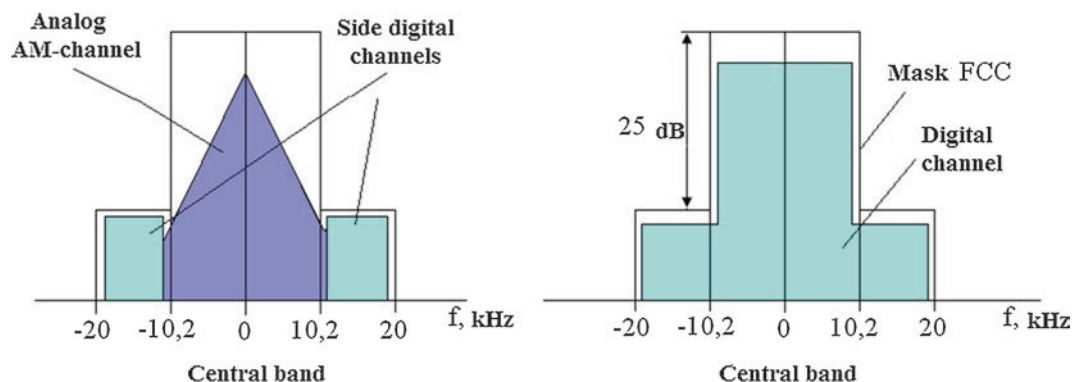


Fig. 1.16 The spectrum of the analog FM signal (a) and the signal mask according to FCC (b)

modes are anticipated: the hybrid and fully digital. The main band of the channels in the FM mode is approximately 240 kHz, while two side bands give two bands by 120 kHz (Fig. 1.15a). In accordance with requirements of FCC (Federal Communications Commission) of the USA, the signal level in side bands should be by 25 dB lower than in the main band (Fig. 1.15b).

In the hybrid mode, the digital broadcasting signals are transmitted in the side bands (Fig. 1.15a), while in digital mode they are transmitted in the main and side bands (Fig. 1.15b). The limit speed of the information flow in the radio channel is 54 kbit/s in side bands and 96...128 kbit/s in the main band. The whole band is allocated for transmission of analog signal broadcasting.

The analog broadcasting system with AM application, which exists in the USA, uses the radio signal that occupies a bandwidth of about 20 kHz at modulation frequency of 9 (10) kHz. When using the hybrid mode (Fig. 1.16a), the side bands of the digital signal are added to the main band (18...20 kHz).

In this case, the bandwidth of the digital channel partially overlaps the bandwidth of the analog broadcasting, but radio wave interference is rather low because this bandwidth is outside of the main band in frequency. In the digital mode (Fig. 1.16b), the signal occupies all frequency band in accordance with the FCC mask. Rates of information flow transmission for the hybrid mode are 20...48 kbit/s, while in the digital mode rates are up to 64 kbit/s.

To achieve the permissible broadcasting quality for narrow enough bandwidths (especially, for AM signal transmission), the coding PAC (perceptual audio coding) technology are used, developed specially for the digital broadcasting systems. The principle laid in this technology is based on the estimation of the channel width: at insufficient width, the main flow is transmitted with some loss of the audio signal quality; if the channel width satisfy the FCC requirements, the additional flow is

transmitted simultaneously, which improves the received signal quality. At information flow rates of 128...96 kbit/s, the quality of the received signal does not practically differ from the original signal: the signal from CD.

The IBOC broadcasting system uses the version of the OFDM modulation with preliminary encoding as in the DVB standard at CTV (color TV) signal formation. The audio signal is transmitted at the same time in the digital and analog form. After analog signal discretization, its audio encoding occurs, and after the convolution encoding and interleaving happen, then the digital sequence passes to the OFDM modulator.

A possibility to transmit the same program in the digital and analog form with some time shift is the feature of the IBOC technology. If the digital signal loss happens, the receiver, imperceptibly for the listener, begins the analog system reproduction that excludes the appearance of periods of unintelligibility known in digital systems. Utilization of such a technology makes expedient the transfer to completely digital broadcasting only in cases when the number of IBOC receivers achieve 80–85%.

The Worldwide Digital Radio The Consortium DRM (digital radio mondiale) is established for the development of the digital radio broadcasting standard in the range below 30 MHz at significant improvement of the audio quality, decrease of interferences, and keeping of the maintenance zone at essential decrease (up to six times) of the transmitter power. At that, the bandwidth of digital signal broadcasting in DRM is equal to the bandwidth of analog systems [44].

The DRM technology uses the method of the radio signal formation with orthogonal frequency multiplexing (OFDM) with the large (about 200) number of carriers with the preliminary quadratic AM (16- and 64-positions). The low-speed encoding of the analog signal is performed using the AAC (advanced audio coding) algorithm supplemented by the SBR (spectral band replication) mechanism for spectral band doubling, which significantly improves the received signal quality without growth of the flow rate. We know that the limitation of the digital flow rate at encoding leads to narrowing of the received signal spectrum: the high-frequency components disappear. The SBR algorithm allows recovering of the high-frequency spectrum part according to transmitted low-frequency part practically without increase of the flow rate (if after compression by the AAC encoder, the upper frequency of the transmitted signal was 6 kHz, then after SBR recovers, it increases to 15 kHz).

The system of the local broadcasting format **DRM+**, developed by the DRM Concern, and standard RAVIS (real-time audio visual information system) system [44] of Russian development support the subscribers by mobile audio-video broadcasting in two frequency bands in the VHS range.

The mentioned digital audio-video systems provide the radio broadcasting signals in I–II bands of the VHF range (65.8–74.0 and 87.5–108.0 MHz) for keeping of frequency distributions.

For systems of satellite digital broadcasting, the international commission on radio frequencies adopted the frequency band of 1452...1492 MHz (besides USA, Russia, Belarus, Ukraine), in which the development of the ground-based networks is possible.

The active development of the wideband DAB system Eureka-147 (digital audio broadcasting) is caused not only by rendering of high-quality digital broadcasting but by a large list of services (information services: low-speed mobile video, transmission of various fixed images, news and reference information).

1.4.5.6 Trunking Systems

Under the *trunking* term (trunk, beam, the communication channel), we understand the method of automatic distribution of restricted number of free channels among a large number of mobile subscribers [6, 7]. This method allows the effective utilization of the frequency resource in systems of mobile radio communication owing to the random access mode to the free channel. So, in the Russian “Altai” system of mobile radio communication, 8 radio channels of the 1 trunk had served about 200 mobile subscribers. At present, this method is used in networks of industrial-technical destination (agencies’ networks) and in the modern radio networks of general use (cellular networks).

Systems of mobile communication that supported a large number of subscribers on a large territory are usually built on the principle of radio-frequency reuse in territory service zones (cells, sites). Trunking networks usually serve one zone (site), since they *do not realize the relay-race principle* of subscriber transfer from one zone to another known in the cellular communication networks as roaming. The main trunking idea similar to arrangement of trunking channels in wired communication systems lies in the selection of one (from the limited channel number) channel to each subscriber during connection time. In addition to effectiveness increase of frequency resource utilization, this leads to conversation confidence increase and the quality growth of rendering services. In the trunking systems, the radio channel is not fixed for the specific subscriber, but he has got any channel free at this moment. Therefore, a number of simultaneously rendered subscribers in the trunking network are equal to the channel number.

It is clear that the dynamic channel allocation requires insertion of the device for the control of channel distribution into the system, and each mobile station should have the technical possibility to switch to any allocated channel frequency. If in the time of the largest load (TLL) all channels are occupied, then the network does not reject calls, but takes them in a cue

for service. Even in TLL, the channel allocated to the pair subscribers is not available for the other station, which requires the connection, which provides the conversation protection against listening. The dynamic rendering of channels also increases the network operation reliability and conversation confidence. A failure of one channel does not lead to access interruption of the subscriber group to the network resources, but reduces a few of the network capacity. The growth of network capacity can be provided by the limitation (by technical means) of the connection time to the channel (by limitation of the conversation duration).

In analog trunking systems, the speech message transmission is provided with usage of FM and multi-station access with frequency channel division (FDMA) at bandwidth of 25 or 12.5 kHz per one channel. At digital procession of the speech signal in the codec with linear prediction, the speech transmission is provided with rates 2.4...9.6 kHz in the same bandwidth.

The data transmissions (message about resource state, the subscriber status, short telegrams, etc.) in analog systems are usually transmitted via the control channel, while the long messages are transmitted via the operation channel with the modem usage. We can mark out two network configurations of stations on the attended territory: single-zone and multi-zone. In the single-zone network, all relays are located on the only base station. In the multi-zone variant, there are several base stations located on the attended territory, and communication with remote relays is performed via allocated communication channels (wired, relay).

Distribution of frequency channels between mobile subscribers offers two schemes: centralized and decentralized control. In the first case, trunking networks contain several relays connected between them by the unified control network. The channel allocation is performed on the base station. In systems with decentralized control, the subscriber stations continuously scan the operating channels in the search of the calling signals of the free channel.

There are several *standards of trunking networks*. Among them, the most important is the standard adopted by Ministry of Post and Telecommunication of UK, which describes the data exchange protocol between the base station (relay) and mobile stations: MPT 1327 (ministry post and telecommunications). Many large companies, which function relying on requirements of the MPT standard, developed trunking networks functioning in several configurations (ACTIONET, Nokia). Trunking networks on the base of the MPT 1327 protocol are characterized by the technology and maintenance simplicity and equipment availability. The shortcomings typical for them are low spectral effectiveness, comparably low immunity against the conversation listening, not high coefficient of frequency reuse, and small transmission rates.

The closed *digital system* EDACS (enhanced digital access communication system) developed for special purposes by Ericsson Co has the rights for equipment and the network structure. Advantages of such system are low time of the network access and the possibility of data and speech transmission via all channels.

The EDACS system exists in two versions: wideband (channel separation is 25 kHz) and narrowband (channel separation is 12.5 kHz) providing:

- Analog signal transmission
- Digital transmission of ciphered speech (in wideband system with the rate of 9600 bit/s)
- Data transmission (with the rate of 9600 bit/s or 4800 bit/s depending on the bandwidth)
- Connection with the TNGU network

This system can function in various configurations depending on the covering zone dimensions. The wideband version of the EDACS system can operate in frequency ranges: 136...174, 404...515, and 806...870 MHz. The narrowband version can operate in the range 894...941 MHz. The EDACS system is the first step of Ericsson Co. during the transfer process to the trunking system TETRA of the second generation.

The trunking system TETRA (*terrestrial trunked radio*) is the ground-based system of trunking radio communication developed within the European Union, and it is an open digital system, which eliminated the shortcomings of analog systems and approached the list of rendered services to cellular communication systems. Two families of developed standards specify the system parameters at speech and digital data transmission as well as the packet data transmission. At speech transmission the various connection variants are provided: individual connection; grouped connection; direct connection; grouped connection with confirmation; and wide broadcasting connection.

Data and speech transmission in the digital form is provided in accordance with adopted standards with the rate of 7.2...28.8 kbit/s (for coding protection absence). The data and speech digital transmission with channel switching is possible with the rate of 4.8...19.2 kbit/s (with minimal code protection). Different variants of data packet transmission in the mode "point-point" with connections or without them (in the standard format) are anticipated by this standard. The architecture of the TETRA system for different types of connected equipment and types of interfaces and connections is presented in Fig. 1.17.

The mobile station represents the final equipment of the radio channel (mobile termination, radio telephone) and terminal equipment, which permits the subscriber to transmit data.

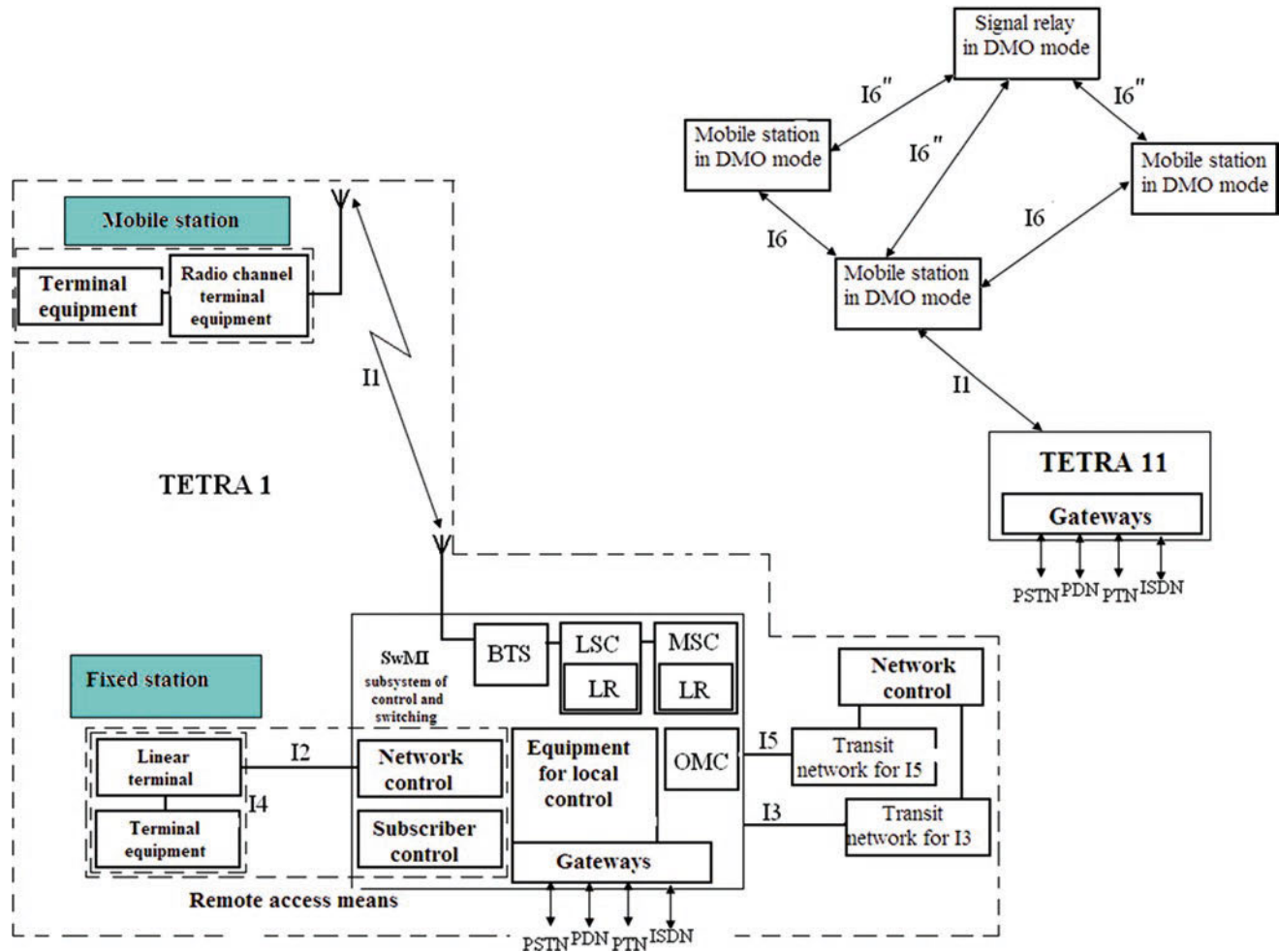


Fig. 1.17 The architecture of the TETRA system

The line (fixed) station includes the same equipment (on destination), but it connects to the sub-system of control and switching with the help of the ISDN channel. The line station can be used in the corporative networks as the dispatcher station.

The sub-system of control and switching (SwMI, switching and management infrastructure) includes base stations (BTS), the main switching center (MSC), the local switching center (LSC) with registers of localization (LR), and the switching center and operation maintenance center (OMC).

As we see from Fig. 1.17, the architecture of the TETRA trunking system allows connections' organization directly between mobile stations in DMO mode (direct mode operation) without application of the inter-network structure.

Using gateways of the control and switching of the TETRA trunking network, we can switch to the PSTN (public switched telephone network) or to PDN (public data network) and PTN (public telephone network). The TETRA technology allows connection to the digital communication network iSDN (integrated services digital network) and provisions the high-speed transmission of various types of data: texts, speech, video, etc.

At the physical level, the TETRA system has the following parameters:

- The bandwidth of the radio channel is 25 kHz.
- Time radio access system (TDMA) together with FDMA (four channels of speech transmission).
- When using more than the one time slot in the frequency channel with a bandwidth of 25 kHz, the data transmission rate is 28.8 kbit/s.
- The differential quadrature phase shift keying with the shift $\pm \pi/4$ or $\pm 3\pi/4$ ($\pi/4$ -DQPSK).

Information in the TETRA standard is transmitted by packets as in the GSM standard. The packet represents the physical content of the one time slot or the sub-slot. There is the variety from six packets of different destinations containing in the

middle the learning sequences. Owing to lesser quantity of time slots in the frame than in GSM, the data packets have the larger length and carry the larger number of bits. The system anticipates the amplifier linearization and control of VCOs.

The digital trunking system iDEN (integrated digital enhanced network) is a unique access platform since it combines together several different mobile technologies, which are based on the enhanced GSM. Services, which are integrated in iDEN, include the distribution system, the duplex telephone communication, the data transmission, and the short message service (SMS). The distribution system has the function of conference communication, when the group of subscribers may participate in conversation. The list of participants can be programmed for two-way connection, or it is possible to create specialized mobile radio system SMR, which has the feature that connection can be created on frequencies, which are available to the channel group physically belonging to subscribers localized in some region.

The telephone communication and data transmission are intended for provision of the traditional mobile communication. The SMS service enables the subscribers of iDEN network to receive messages containing not more than 140 symbols.

Application of multi-station access technology TDMA provides the iDEN system compatibility with cellular communication systems, which act on the base of the D-AMPS protocol. The iDEN system in the FDD mode separates the reception and transmission channels with the bandwidth of 45 MHz with the occupied frequency band of 15 MHz. For bandwidth allocated for 1 subscriber channel of 25 kHz, we may arrange 600 frequency channels in the operating band, which provides the capacity of such a system, which are 8 times more compared to the GSM system.

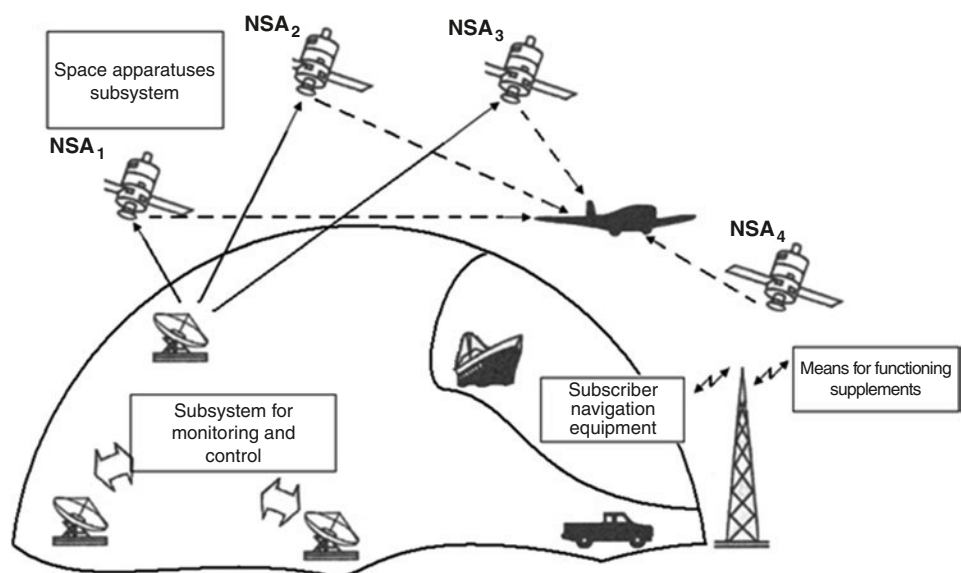
The iDEN system has many indices typical for cellular systems: the error correction, the packet mode for data transmission, and a possibility to arrange connection between the subscriber terminal and PSTN, as well as to receive and to send facsimile, to provide the exit in the Internet.

1.4.5.7 Global Navigation Satellite Systems

The ensemble of radio devices and auxiliary technical means for solution of the problem of mobile object positioning is called the radio navigation system (RNS). Systems for determination of fixed objects' location (the survey affixment) can be attributed to these systems. The auxiliary means can be located on the object or outside it, while different events of natural or artificial origin can be considered as information sources for solution of navigation problems. Navigation systems are concerned to the non-autonomous class if the artificially created fields and data are applied for determination of the object position.

At solution of the *positioning* problems, radio navigation systems use radio signals from the reference radio navigation points (RRNP) from external artificial sources with known coordinates. Methods permitting to unambiguously determine the current object location without attraction of data about positions in previous time moments (from the motion start) can be attributed to the position class, while all others are to the non-position class. The main RNS task under solution is the determination of the state vector (SV) of the object including spatial coordinates (x, y, z), axes components of the velocity vector, and the subscriber time scale amendment with respect to the system time scale. The mentioned SV parameters cannot be measured directly by radio engineering methods; therefore, the subscriber measures the indirect parameters: the time of radio

Fig. 1.18 GNSS structure



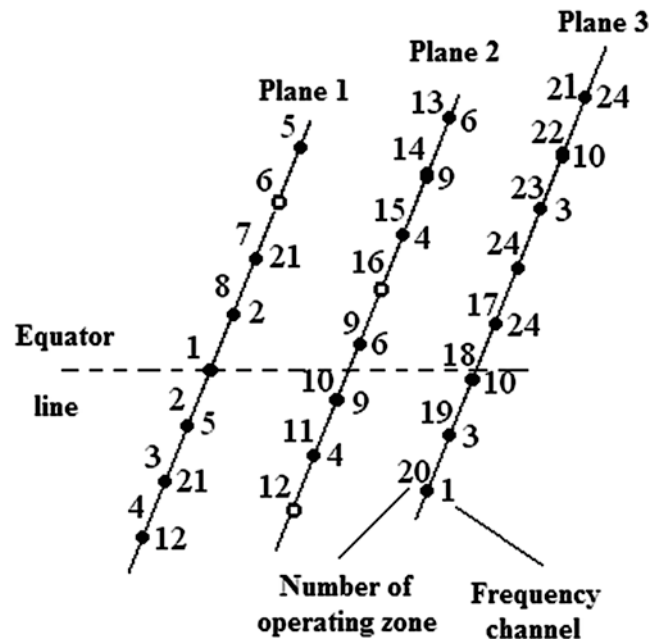


Fig. 1.19 The orbital group of navigational space apparatuses for GNSS

signal arrival, its phase (phase difference) with regard to the reference oscillator signal, the Doppler frequency shift, an amplitude, etc. Measured radio navigation parameters are compared with the functionally connected navigation parameters, which characterize the positioning angles with respect to RRNPs, the distance (or distance difference) with respect to several RRNPs, etc. Using a priori known navigational functional dependences between navigational parameters and the state vector as well as coordinates of reference radio navigation points, we can determine components of the state vector.

The main difference of global navigational satellite systems (GNSS) from the ground-based ones consists in the fact that the role of reference navigation points is played by navigational space apparatuses fulfilling the autonomous orbital motion. This leads to necessity of inclusion in the GNSS structure (in addition to the radio navigational ground-based base stations) of the ground-based control center, which provides the monitoring and prediction of the orbit parameters of navigational space apparatuses (NSA), the monitoring and correction of onboard time etalons, the monitoring of the onboard equipment, etc.

Accordingly, in the GNSS structure (Fig. 1.18) three main sub-systems are included:

- Navigational space apparatuses
- Ground-based complex for monitoring and control (GBCMC)
- Subscriber equipment (SE)

In order to increase the measurement accuracy due to application of the differential methods, the means of functioning supplements are included into the GNSS structure.

The NSA GLONASS Sub-system The design orbital group of navigational space apparatuses of GNSS should contain 24 NSAs located on the three circular orbits: 8 NSAs on each orbit (Fig. 1.19).

GLONASS satellites are located on three circular orbits with orbit planes' incline of 64.8° and longitudes of orbit planes' interception on the equator line of 120° . The altitude of NSA location is 19,100 km over the Earth surface, which corresponds to the nominal orbital period of 11 h 15 min 44 s and realizes of the non-synchronous mode (in contrast to the GPS system). In this mode, the circular orbit track slowly moves with respect to the Earth surface (in a period of 7 days 23 h 27 min 28 s) providing practically the same orbit disturbance (on average) for all NSAs.

Application of autonomous navigation technology, which allows inter-satellite measurements of distances and velocities of the relative displacements of the NSA group, will permit to supplement and to correct information passed through the ground-based monitoring-control stations and to provide the system operation ability at information absence. For conduction

of inter-satellite measurements, transmitters with the power of 60...100 W will be used operating in the range 2,2 GHz and providing the information transmission rate of 500 bit/s with application of the code signal division.

With renovation of acting NSA group, we simultaneously solve a series of tasks, which provides the higher positioning accuracy (up to 2...3 m), which requires:

- Instability reduction of the onboard synchronizing device to 10–14
- Transition to unrequested measuring means for amendments of the onboard time scale
- Expansion of the measuring stations' network
- Putting into operation of functioning supplements' means

The GPS NSA Sub-system Formation of the global navigation satellite system GPS, which is regulated by the Interface Control Document GPS [32, 45], having the status of double purpose (military and civil), had finished in 1996.

The regular GNSS group consists of 24 NSAs located on 6 circular orbits with the altitude approximately 20,000 km, with planes inclined at an angle of 55°. The NSA orbital period is 12 h 00 min and realizes the synchronous mode in GPS. At the stage of GPS development and exploitation, its navigational space apparatuses were modernized, which at present allow the autonomous operation of the NSA sub-system up to 180 days using the AutoNav system. The AutoNav technology is realized at NSA location of devices for inter-satellite data exchange, which permits to fulfill measurements of mutual distances and have the autonomous synchronization of the onboard time scale with specification of the orbit parameters. In autonomous mode, the equipment provides the GNSS operation with the regular group not less than 16 NSAs without significant accuracy decrease (the rms error is no more than 16 m at coordinate determination during specified time) with keeping of characteristics for functioning stability. If the AutoNav system is absent, the mentioned characteristics may be kept on the period of not more than 2 weeks. The further development of inter-satellite measurement system can be its inclusion into the ground-based station system, which will allow the essential increase of calculation accuracy using the ephemeris time (uniformly running time without account of the inconstancy of Earth rotation) under conditions of restricted number of used ground-based control stations.

The GALILEO NSA Sub-system This sub-system is developed under patronage of the European Union and European Space Agency. The orbital GALILEO group consists of 30 NSAs on three medium orbits with an altitude of 23.616 km and incline of 56°, which provides the best observability in the Northern Hemisphere. In prospects, one is going to use in the GNSS system (besides medium-orbit NSAs) also the geostationary navigational space apparatuses. The GALILEO sub-system, besides the orbit NSA group, as in the above-considered, includes objects of the ground-based infrastructure (the control complex, the subscriber equipment). The expected system accuracy at the mentioned NSA group model should be 4 m in the horizontal plane and 8 m in the vertical plane for the confidence interval 0.95.

The GALILEO project is oriented on the active interaction with other systems, which solve the similar tasks. So, the installation on GALILEO NSA of relay station for Cospas-Sarsat service is planned, which will widen its possibilities, especially, in northern regions that are difficult to access. Subscribers of "Globalstar" and "Orbcom" networks equipped by mobile terminals with built-in GPS receivers will be able to interact with ground-based radio navigation systems, for example, "Loran-S." We can use interaction with "Inmarsat-3" system, which transmit data about the state of GLONASS NSAs and GPS groups.

The operation of the GALILEO system provides:

- A public service for providing navigation and determining the coordinates of moving objects with data transmission through the open service (OS) mobile systems
- The service providing the motion safety (Safety-of-Life Service), on demand by naval and avian agencies
- The Public Regulated Service (PRS), which provides the navigation data to police, the civil defense, the safety communication, etc.
- The commercial service ensuring the paid service to registered subscribers in addition to OS service on motion control.

The combined receiver for GLONASS, GALILEO, and GPS can simultaneously operate and use for navigation and location determination more than 30 NSAs, which allows to perform time-navigation determinations under any conditions (shading in cities, mountains, forest regions).

Principles, Algorithms, and Processing Devices for Navigation Information *The construction principle of the NSA signal search and detection algorithm* in receivers of the subscriber equipment, which are developed as multi-channel GLONASS systems, begins from estimation of signal parameters, which are received in each channel. Characteristics of 12–16 channels

are estimated in comparably simple GPS receivers and several hundreds and more in multi-system (GLONASS, GPS, GALILEO) receivers. At first stage, the pair of values of $\{F_D, \tau\}$ is estimated: the Doppler frequency shift and the time delay, which correspond to a maximum of a posteriori probability. A priori values of measured parameters are considered as independent and having the uniform continuous distribution density. The search in the discrete totality ($i \in N$) of parameters $\{F_{D_i}, \tau_i\}$ combination, which exceeds the threshold values, allows to state that the detected signal parameters $\{F_D, \tau\}$ belong to some space region (a cell), which is adjacent to the point $\{F_{D_i}, \tau_i\}$. The cell dimensions are defined by the maximal permissible radio navigation error.

The signal search and detection are solved by optimal filtering, which is performed in the subscriber equipment with the help of multi-channel correlation processing of received signals after radio signal demodulation. After processing in quadrature channels, the total signal passes to the threshold device, where the decision on signal detection is made on the base of comparison with the threshold value with transfer to the filtering mode and to the navigation information decoding mode, or the search procedure is continued if the signal is not detected. Correlation processing of the multi-channel signal on the interval equaled to distance-measuring PRS duration allows SNR increase at the receiver output from -30 dB to about 0 dB at the correlator output (the number of PRS symbols for GLONASS is 51 symbols; for GPS the number of symbols is 1023, which is the SNR growth of $27 \dots 30$ dB). The further SNR increase up to $7 \dots 10$ dB is provided by application of the non-coherent accumulation during $3 \dots 5$ samples of the total square of orthogonal components from correlator outputs.

Obtained radio navigation parameters: the pseudo-delay τ_{ps} and the pseudo-Doppler shift $F_{D,ps}$ are primary results, which are the random quantities allowing determination of the *subscriber state vector* by statistical methods. There are different methods for estimation of state vector parameters, one among them is the widely used single-step algorithm based on the independent processing of the variety of values obtained in each time moment. Coordinates of NSA and the subscriber are described by the nonlinear equation system, the solution of which is based on the primary measurement results for all visible NSAs, the number of which can be more than minimally necessary (four). If obtained estimations significantly differ from a priori values, then the results obtained are used as primary at the current iteration of the navigation problem solution by the method of successive approximations. The described algorithm does not take into account the results of previous observations (coordinates of the velocity vector, information about oscillator frequency drifts and time scale offset, etc.), which reduces the measurement accuracy.

The double-step algorithm of signal processing performs estimation of NSA signal parameters at the first stage by independent channels of the navigation correlator. The primary signal estimation (pseudo-delay, pseudo-Doppler frequency shift) on each NSA under processing at the second step of its processing in the navigation processor forms estimations of state vector components. This is provided by the least-squares method or by solution of expanded Kalman filter equations. The state vector includes 11 components: 3 spatial coordinates, 2 estimations of the reference oscillator parameters (time scale and frequency scale), 3 estimations of the velocity vector components, and 3 acceleration vector components.

The single-step algorithms are the alternative of double-step algorithm for signal processing in the subscriber equipment in the case when the tracking process for radio navigation signal parameters is excluded, and we use the filtering of state vector components from all varieties of simultaneously observed NSA signals. The advantage of such an approach for signal processing is the possibility of efficiency keeping of the global navigation system at moments of shading of the one or several satellites till the moment until errors of extrapolate estimations of the state vector do not come out from the admissible limits. Recovering of the necessary NSA signal level allows practically immediate inclusion of received signal parameters into the process of the state vector calculation in the subscriber equipment.

The Accuracy of Navigation-Time Measurements The potential accuracy of the pseudo-random delay estimation by the envelope at the matched filter (correlator) output is mainly defined by the received signal's duration and shape. For standard pseudo-distance measurements in the GLONASS and GPS systems, as the first approximation, the rms estimation value is $1 \dots 10\%$ from the duration of distance-measuring PRS, which in the terms of the real physical length is $3 \dots 30$ m. The potential pseudo-distance estimation measured on the base of the carrier phase is defined by the spectral density of phase noises of the NSA onboard synchronizing device and is not more than 0.1 radians, which corresponds to the rms error not more than 3 mm.

Real values of measurement accuracy in GNSS are larger than mentioned potential estimations, which is caused by a series of factors:

- Influence of NSA equipment architecture, the ground-based navigation control complex, and used processing algorithms of navigational signals
- Parameters of the signal propagation medium from NSA to the subscriber equipment
- Parameters of the subscriber equipment

The mutual position of NSA and the subscriber provides the significant impact on the accuracy of navigation-time determinations.

The frequency-time support of the NSA onboard equipment is defined by the accuracy of collation procedures between the onboard time scale and system time scale. At that, the arising phase offsets of distance-metering codes and time marks lead to error at measurements of radio navigation parameters and, hence, the value of navigation-time determination. For the GLONASS system, the correction of the onboard time scale with respect to the system scale is performed at the shift of 10 ns. At onboard oscillator instability 10^{-13} , such a discrepancy arises already after 2 h after correction, and after 24 h it is 25 ns. For NSA with cesium etalon, which has a relative instability of $5 \cdot 10^{-14}$, the error of the NSA onboard time scale drift during 24 h is 10 ns. The inaccuracy of NSA orbit parameter calculation, which is caused by the unpredictable deviation of the real orbit from the extrapolated orbit, also leads to the appearance of the rms error in accuracy of the distance measurement, which is 1.5...4 m.

For high-accurate radio navigation measurements, the radio wave refraction and the inconstancy of its propagation speed in the ionosphere, as well as multipath property resulting in received signals interference, will give the significant contribution in the total results error. To decrease the rms error caused by delay in the ionosphere, we calculated the duration of the real delay during navigational measurements, using the method of redundant measurements. Application of the even approximated enough model for description of the ionosphere state, for instance, the eight-parameter model obtained on the base of statistical analysis, allows reduction of the ionosphere error by 75% for the GPS system. The model parameters calculated in the ground-based control center are inserted into the NSA navigation message for transmission to the subscriber. It is rather difficult to take into consideration the ionosphere influence on the navigation measurement errors in the GLONASS system using the single-frequency channel for the separate NSA, and it requires the attraction of additional information from other sources and results of redundant measurements.

Enough exactly known and predictable influence on the permittivity of troposphere layers of the atmosphere permits to reduce the residual equivalent error of the distance measurement to about 2 m.

The signal interference caused by multipath properties of NSA signal propagation has an influence on the object positioning accuracy. To reduce this influence of the reflected signals, it is expedient to use the wideband signals with the sharply expressed peak of the auto-correlation function and the low level of side lobes of this function. Application of antennas with polarization allows simplification of direct signal extraction. The angles of signal arrival, which are defined by NSA elevation over the horizon, have the serious influence on the distortion character and level of received signals.

The relative position of the subscriber equipment and the operating NSAs influences on the exact determination of subscriber equipment location. The minimal value of the error at location determination with usage of distance-measuring methods corresponds to locations of operating NSAs, for which one NSA is in the zenith and three others are uniformly located in the horizontal plane. At that, it is necessary to take into consideration that, in practice, it is not expedient to use NSAs located at elevations less than 10° , since its reduction increases the negative influence of atmosphere properties and the multipath conditions on the measurement accuracy.

Relative Measurements and the Differential Correction The above-considered factors influencing on the navigation measurement accuracy show that the modern subscriber equipment, in the standard mode, provides the equivalent distance error of 5...10 m in the rms value. This corresponds to calculation of the absolute coordinate values with accuracy of 10...30 m under condition of the optimal NSA location on the orbit, which is insufficient for a whole series of navigation applications.

The methods of joint processing of measurement results, which are obtained from several spatial-spaced points of the GNSS operating zone, allow increasing of navigation-time determination accuracy.

Since the subscriber should often know the relative (but not the absolute) location of the object, for this one applies principles of differential processing of measurement results with further correction. Carrying out of relative pseudo-distance measurements, at least, from two space points allows elimination of the correlated error components (atmosphere, synchronization of onboard time scales) and increase of the measurement accuracy owing to phase methods application at joint processing of navigation-time determination results. On the base of difference of the pseudo-phase first estimations, we form the nonlinear equations, permitting the linearization owing to smallness of the pseudo-distance measurement error compared to NSA distance.

However, application of the phase measurement method is accompanied by arising of obtained result ambiguity, which appears at the base line length exceeding the wavelength and is caused by the cyclic character of the phase-time characteristic. To solve the ambiguity, we can often use methods based on redundant measurements and using the signal processing, which arrive from different directions. At implementation of the phase method of relative redundant measurements, we may face with the necessity to perform measurement on the base of combined constellation of satellites (GLONASS and GPS) or to use the second frequency channel.

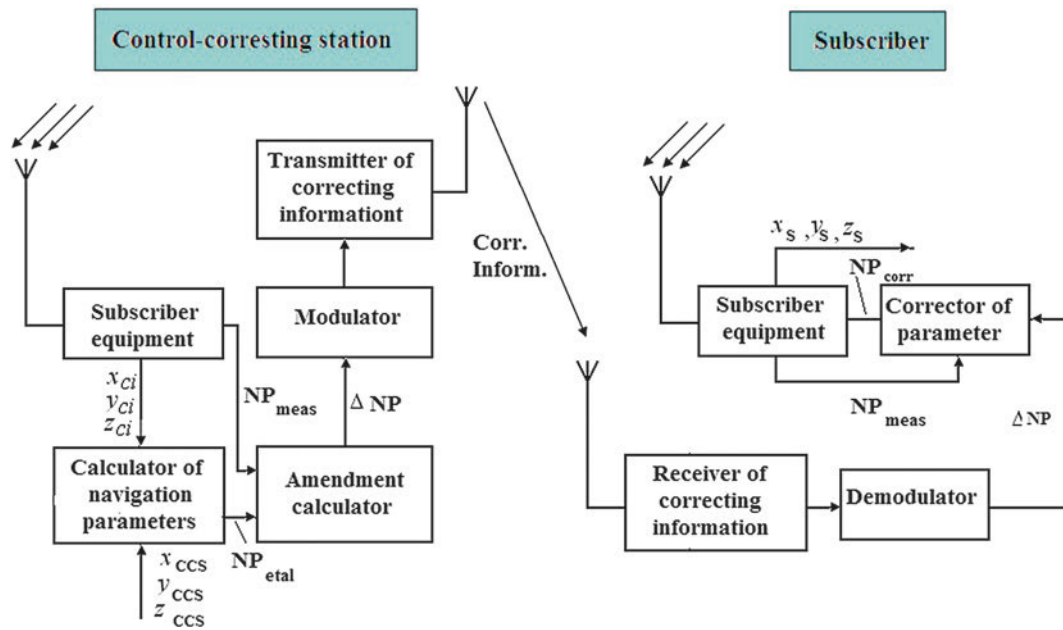


Fig. 1.20 The structure of the differential correction sub-system

Application of relative measurement methods allows determination of the ground-based point position with the error less than 0.1 m for geodesic network creation of different purposes. At that, the tracking accuracy for the carrier phase can achieve several millimeters and fulfill measurements of basing lines at the distance up to 1000 km with the accuracy of several centimeters. Processing algorithms used for measurement results (real-time kinematic) allow conduction of measurements and processing in real-time mode.

Methods of *differential correction* are based on using one or several points, in which measurements are provided as in the reference points. Amendments to results of navigation-time determinations are sent to the subscriber and used for specification of own absolute coordinates.

The complex of technical means realizing the mode of differential correction is called the differential sub-system. Depending of the covering zone, we distinguish the local differential sub-systems (LDSS), regional differential sub-systems (RDSS), and wide-zone differential sub-systems (WDSS).

The structure of the differential correction sub-system (Fig. 1.20) includes the control-correction stations (CCS), the channel of correcting information transmission (CCIT), and the equipment for reception and processing of correcting information in the subscriber equipment.

Geodesic coordinates of the control-correcting station $x_{ccs}, y_{ccs}, z_{ccs}$ are known with high accuracy and are used as the navigation etalon parameters (NP_{etal}) in the subscriber equipment. The precision subscriber equipment has the low level of the random error, which allows calculation with its help of primary values, for instance, of pseudo-distance, navigation parameters of the current measurement (NP_{meas}). According to known values of control-correcting station (CCS) coordinates $x_{ccs}, y_{ccs}, z_{ccs}$ and obtained from NSA information, calculated using the uniformly running time (ephemeris time, ET), x_{ci}, y_{ci}, z_{ci} , we calculate the etalon value of parameters NP_{etal} , which is compared to the measuring value. The difference between etalon values and results of current measurement $\Delta NP = NP_{etal} - NP_{meas}$ is the impact, which forms amendments transmitted to the subscriber as a part of correcting information. The subscriber introduces amendments to the value of measured navigation parameters NP_{meas} according to the known value of correcting information. Thus, we form NP_{corr} calculating on its base parameters of the state vector.

The accuracy of navigation parameters, using the method of differential correction, depends on errors of the subscriber equipment for the etalon point and the object, on the distance between them, and the dependence of auto-correlation function versus the equivalent errors of the distance and time. The differential correction provides the maximal effectiveness at subscriber equipment distance from the CCS less than 500 km, when the correlation in space and time remains between measurements, performed from different point of the space.

The *local differential sub-systems* provide the maximal operating distance of the control information transmitter not more than 200 km and are used in the control systems of the ground transport, geodesic, and other directions. The accuracy of

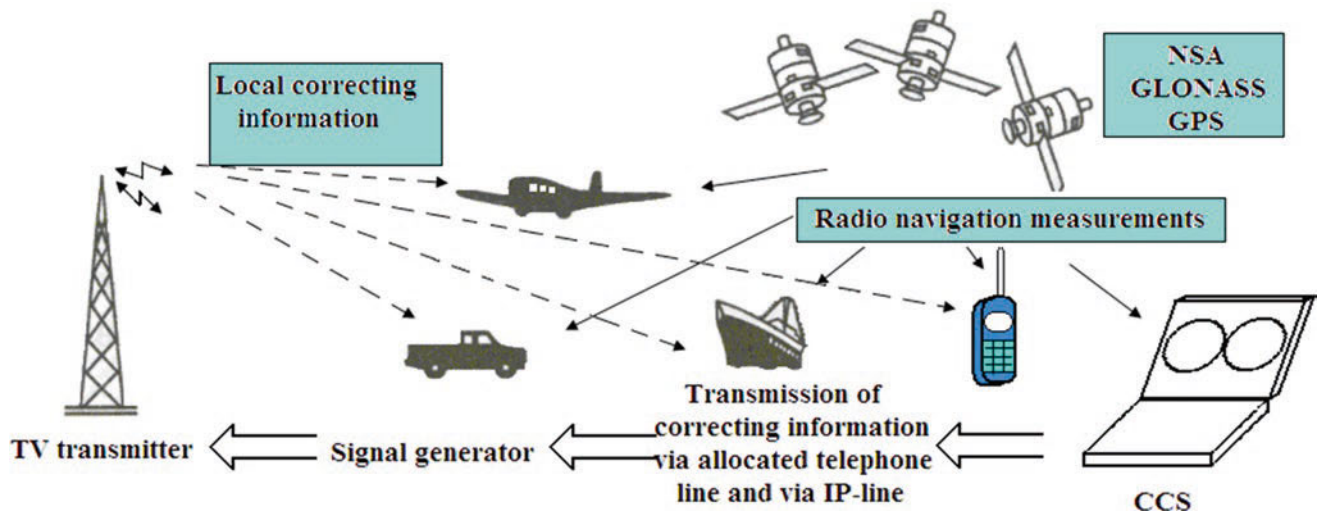


Fig. 1.21 The structure of local differential sub-system

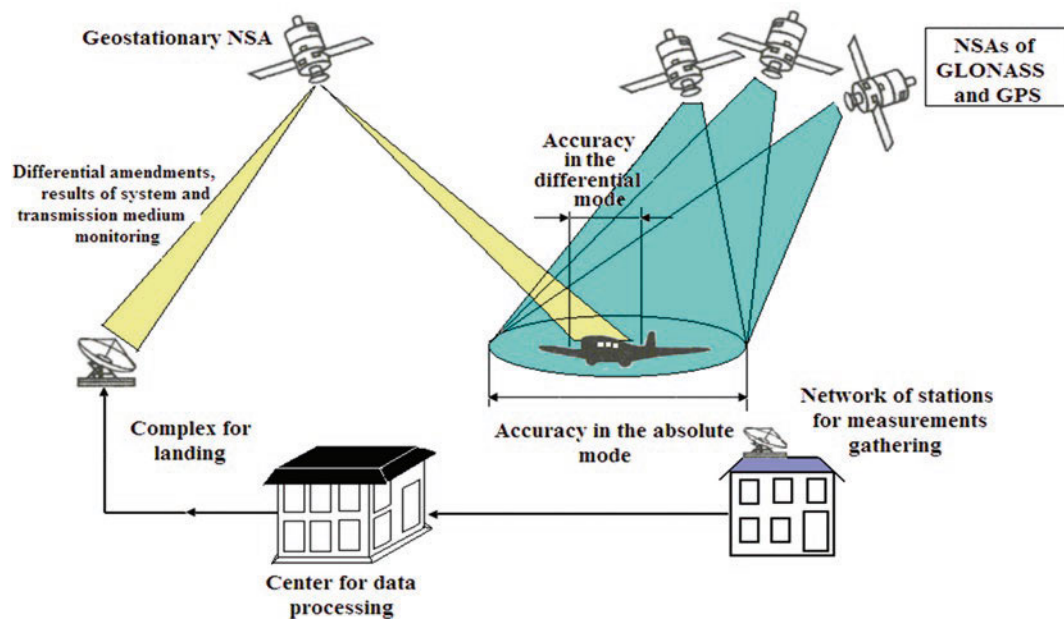


Fig. 1.22 The conception of the information gathering system

coordinate determination provided in such systems is not worse than 1 m in the plane and 2 m in azimuth for utilization of transmitters for data transmission from regional TV or FM radio broadcasting (Fig. 1.21).

Utilization of local differential sub-systems for monitoring and motor-car motion control is widely used in European countries. For this, the sub-carrier modulation of UHF radio broadcasting channel is used. Local sub-systems for aero-navigation purposes use the 8-PSK modulation in radio broadcasting transmitters of the range 112...118 MHz for creation of the low-speed channel of correcting information. In the highly complicated conditions for providing of instrumenta-

tion landing, the radio channels in the range 5...5.25 GHz are organized with joint usage of the local and wide-zone differential sub-systems.

Regional sub-systems unite the local differential sub-systems in the unified network, which allows provision of operation distance increase up to 500...2000 km. Several tens of control-correcting stations (CCS) are included in this sub-system, which are included also in the unified network of the system for information integrity monitoring. Regional sub-systems provide the location accuracy of the order of 1 m for distance 1000 km to the object.

The *wide-zone system* of differential correction and monitoring (DCMS) developing in Russia should provide the high-accuracy navigation not only in the Russian territory but in all zones of economical and other interests of Russia. The conception of DCMS development is going to use the following:

- The existing group of GLONASS NSAs together with geostationary space apparatuses
- Channels of geostationary and GLONASS NSAs for amendment transmission
- Creation of the network in the Russian territory from 10 to 20 autonomous stations for information gathering
- New methods of measurement processing and obtaining of correcting information in the asynchronous mode, which will simplify the network equipment for information gathering and reduce the total system cost.

In this case, the information gathering system does not produce the correcting information as the control-correcting stations, but only performs measurements of navigation parameters over all visible NSAs (Fig. 1.22).

The preliminary accurate affixment of coordinates of information gathering stations is not required in contrast to traditional control-correcting stations. In addition, the base stations are anticipated in the structure of the wide-zone system, which are located, geodesically affixed, and equipped by their own etalons of time and frequency.

In such systems, for the masses of subscribers, we can expect to have an information about troposphere and ionosphere amendments (instead of calculation in the subscriber equipment), which will permit to increase the accuracy of the single-frequency equipment. At that, a possibility is keeping to use results of autonomous double-frequency measurements for correction of atmosphere errors. For subscribers, who have the standard equipment and are included in DCMS, this will permit to make corrections of final measurement results (a velocity and coordinates). The residual error of subscriber coordinate determination is 33 cm (in rms value), and taking into account the geometric factor and multipath properties, we shall have 1...2 m for the area 500 × 500 km.

Checking Questions

1. Give the definition of interference in radio channel.
2. What is the estimation of the reception noise immunity in analog and discrete radio systems?
3. Enumerate the types of spurious impacts, and denote the most typical among them for cellular communication systems.
4. What is the source of lumped interferences? Denote the approaches for struggle against them.
5. Denote the sources of the pulse interferences and methods for struggle against them.
6. Show features of the energy spectrum and its phase characteristic starting from the nature of fluctuation interferences.
7. What is the difference of the radio channel properties having the “memory” and “without memory”?
8. Which kinds of distortions of the single pulse are arisen in the radio channel with the memory and numerate the methods for struggling against them?
9. What is the reason of etalon model appearance for interaction of OSI radio systems? At which level does the signal exchange occur between the base station and the subscriber terminal?
10. What does the term “the radio access methods” mean? Give the most widespread methods of the radio communication systems and explain their differences.
11. How much channels are formed in the physical channel of the GSM900 standard?
12. Which frequency band is allocated to the single connection (subscriber) in the GSM900 and WCDMA (DAMPS and cdma2000) standards?
13. Relying on properties of radio signal propagation in the free space, explain the reasons for a choice of operating frequency range, for example, for the cdma2000 system.
14. Compare methods of radio signal formation in the mobile systems of GSM, WCDMA, and OFDMA standards.
15. Give the comparative analysis of the navigation GLONAA and GPS systems in constructive structure, the number of navigation space satellites in the group, and the modulation type.

Chapter 2

Systems and Networks of Wireless Communication

This chapter is devoted to systems and networks of wireless communication. At first (Sect. 2.1), destination and features of wireless systems are discussed. We describe the personal wireless networks, local wireless networks, networks of the municipal level, and global networks of mobile communication. Section 2.2 is devoted to detailed description of personal wireless networks. Systems Bluetooth, sensor networks with its most important parameters, ultra-wideband communication systems, and systems of radio-frequency identification are analyzed. In Sect. 2.3, local networks of wireless communication of different standards including DECT standard are examined in detail. Section 2.4 is devoted to municipal networks with wireless access for various standards. Systems of cellular communication are described in Section 2.5 with examination of features of different generations, including 5G, of communication systems with the emphasis to used standards.

2.1 Destination and Construction Features of Wireless Networks

Systems of wireless communications are quite manifold in their purposes, methods of construction, ways of the multi-station access (the channel division), a type of transmission signals, etc. The main purpose, which was pursued at development of both open and closed systems of wireless communication, consisted in providing of radio communication with mobile objects. The further perfection of signal transmission approaches, the transition to development of digital transmission systems, the development of jam-resistive types of modulation, and methods of encoding provided a possibility of high-speed data transmission and multimedia information.

The cellular method as the most flexible turned out to be the basing approach for the creation of the mobile communication networks of various purposes, the frequency range and frequency bandwidth allocated to the single channel. This method provides any configuration of the covering zone for nonuniform traffic density inside of this zone.

Depending on the operation distance and the transmitter power, the wireless networks can be divided (Fig. 2.1) into personal (Wireless Personal Area Networks, WPAN), local (Wireless Local Area Networks, WLAN), metropolitan (Wireless Metropolitan Area Networks, WMAN), and global (Wireless Wide Area Networks, WWAN), so-called networks of (mobile) cellular communication.

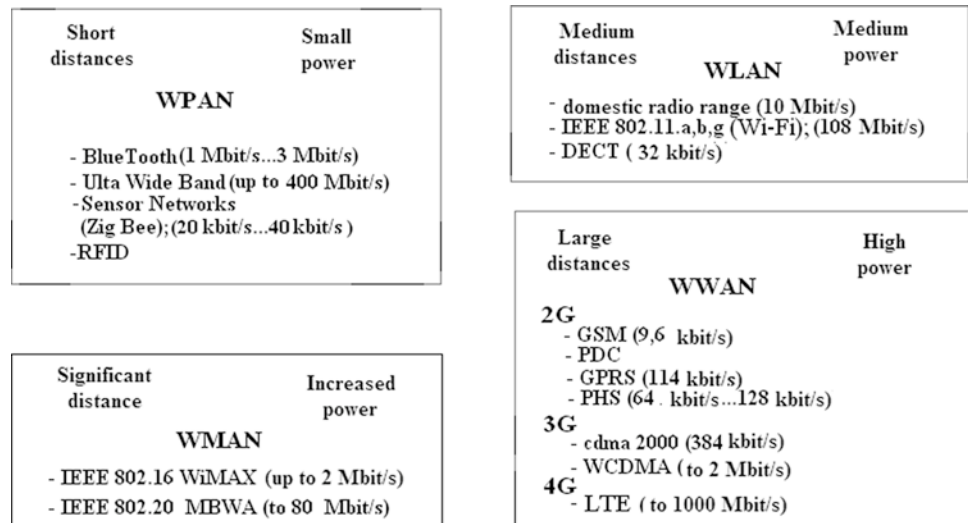
2.1.1 Personal Wireless Networks (WPAN)

Personal networks (WPAN) are intended for interaction of the domestic equipment. The operation radius of the receiving-transmitting equipment is approximately 10 m, and in the narrowband standard (IEEE 802.15, Bluetooth), the transmission speed is 1 Mbit/s. Bluetooth provides the information exchange between personal computers, mobile phones, printers, digital photo cameras, etc. free of charge at small distances between devices. The transceiver implemented on the single chipset can be built into cellular phones and computers and function in two modes as a sensor with low power consumption and as the high-speed relay.

At realization of ultra-wideband (UWB) femto-networks for high-speed data exchange, the transmission rate is 400 Mbit/s at the distance between objects less than 1 m and 110 Mbit/s for up to 10 m.

Depending on the protocol, sensor networks are used for: tracking of displacement of different objects, reading of sensor indications, changing of operation modes of devices connected to the network, etc. The data transmission rate is 1.2...250 kbit/s at the distance between network nodes up to several hundreds of meters (usually restricted by direct visibility limits).

Fig. 2.1 Types of wireless networks



Intensive development of the *Internet of Things* (IoT) as the global system of information exchange not only between people but also between machines (M2M, machine to machine) as subject of such a network led to the interest to sensors. Elements (sensors) of the environment, state monitoring of objects, their displacement in the space, and certification are included into the network of radio-frequency identification (RFID), which serves for wireless goods and other subject identification, for fixation and data correction about them, and for remote delivery of these data on demand. The system includes two types of radio devices: tags and readers. Each accountable article of the goods unit (an object) is supplied by a tag containing information, which provide its identification (information about place of production, object location, its movements, etc.). Readers in the RFID systems provide the remote data readout located in the tag but also perform the correction of data based in a tag. The operating frequency region, in which RFID networks are built, can be from tens of MHz to 900 MHz.

2.1.2 Wireless Local Networks (WLAN)

In the usual local computing network, its elements – servers, terminals, printers, and another periphery – are connected between themselves by copper or fiber-optical cables, which, at any changing of the network configuration, requires the additional cable laying. This increases expenses on the network exploitation and restricts terminal spatial displacements. *Wireless local networks* (WLAN) are intended for the *digital data exchange* between computers, i.e., between premises, and have such advantages as:

- The network flexibility, when network nodes can be located at essential removal (up to 100 m) and outside of the direct visibility zones.
- The simplicity of network planning, while networks of “ad hoc” type (special networks) with the episodic access of the arbitrary structure do not require reconfiguration. The structured network having the wire infrastructure, which connects it with other networks, is another type of WLAN. Points of high-speed wireless access to Internet (WiFi hot spots), which provide the stations (nodes) connection of the local network to Internet, are important components of such a network.

The network can be both autonomous and connected with networks of any other types including wired. In many cases, the WLAN plays the gateway role between a computer, the smartphone, and Internet, which requires ensuring of these networks’ compatibility for subscriber applications.

The WLAN represent a large choice of radio technologies, transmission rates, frequency ranges, and possibilities of application. The first standard of wireless access to *computing networks* **HIPERLAN** (**HIgh PERFORMANCE Local Area Network**), adopted by the European Institute of Telecommunications Standards (ETSI) in 1995, defined requirements to the interface compatible to the OSI standard. The **WiFi** network, which unifies IEEE 801.11b and IEEE 801.11g standards, is the specific case of WLAN. Local networks are arranged in the non-licensed frequency range **ISM** (**I**ndustry **S**cience **M**edicine) 2.4 GHz intended for use in industrial manufacturing, medical organizations, and scientific labs. The physical level interfaces of HIPERLAN and WiFi standards coincide. The necessity of data transmission rate growth in WLAN connected with assimilation of infrared frequency range in the region of 5 GHz leads to possibility expansion of these networks on the physical level.

Improvements in noise immunity are provided by the frequency bandwidth increase, which is occupied by the radio signal with modulation application based on the spectrum spread by the method of direct sequence DSSS or multiplexing by means of orthogonal carriers OFDM. The last spectrum spread method for radio signal is used with restricted number of sub-carriers in the IEEE 802.11a standard for the frequency range 5.1 GHz, as well as in many other systems.

Networks of DECT (Digital Enhanced Cordless Telecommunications) standard, which is the perfected standard of digital wireless (cordless) communication, allow subscribers to provide the connection between the telephone (mobile station) and the base station connected to the PSTN or to office ATS. The operation distance of cordless phones is restricted by several hundreds of meters and assumes the low mobility of the subscriber. It was a base for development of systems of the *subscriber radio access WLL (Wireless Local Loop)*. Application of **PMP (Point-to-MultiPoint)** technology permits to provide subscriber connection through the central station with PSTN. The central station is connected with the central radio station, which transmits and receives signals from the terminal radio stations. Radio signals, which are received by the terminal radio station, are distributed over subscribers with the help of copper subscriber lines, to which the usual phone apparatuses are connected. At application of the DECT technology, the radio lines can be used with provision of the communication distance to the terminal radio station not more than 5 km.

2.1.3 Wireless Municipal Area Networks (WMAN)

Networks of mobile communication are subject to influence both inter-system and intra-system interferences, which leads to validity degradation and to data transmission rate decrease. To increase the noise immunity of networks with the radio access, we can apply some approaches; among them there are the most widespread:

- Application of wideband signals.
- Spatial-temporal signal processing.
- Jam-resistive data coding.

Networks with the radio access are called wideband, in which for data transmission the wideband signals are used. In such networks, the radio signal occupies the frequency band, which significantly exceeds the message spectrum. For example, if you want to transmit a speech message with a frequency band of 3.4 kHz (a standard tone frequency channel), a frequency band with a width of 1.23 MHz is allocated for its transmission in the radio channel of system CDMA2000. At that, the applied additional modulation does not connect with data, but it is used only to increase the receiver messages' truth. At the same time, if for TV image (bandwidth of several MHz) transmission we use the radio channel with the width of several MHz, then such a system is not wideband.

Municipal networks are implemented on the base of IEEE 802.16 standard. The equipment of this IEEE 802.16 standard **WiMAX (Worldwide Interoperability for Microwave Access)** provides the operation radius up to 10 km. At present, the frequency range used for such systems is 2...11 GHz, which allows (after widening of operating frequency range to 10; 20; 25 MHz) increase of the system noise immunity. Application of the OFDM technology improves the effectiveness of struggle with the signal multipath and decreases the distortion influence caused by delays in the radio channels and, accordingly, increases the distance to base stations. The growth (compared with WiFi) of the number of sub-carriers, each of which is modulated by digital sequence, which is extracted from the initial sequence, and has the essentially lesser rate (the larger message duration), allows spectrum narrowing on each sub-carrier, providing at that their orthogonality. Applied types of the BPSK, QPSK, and multi-position QAM modulation provide the transmission rate of 136 Mbit/s with fixed objects and 2 Mbit/s with mobile objects.

With the development of the IEEE 802.16 standard for mobile application, in 2008, the standard **MBWA (Mobile Broadband Wireless Access)** was adopted. This standard anticipates the system operation in the licensed frequency range (below 3.5 GHz) with the operating frequency bands of 5, 10, or 20 MHz in the mode of packet switching at the maximal data transmission rate up to 80 Mbit/s. At that, the support of any standards and Internet applications should be provided, including the speech transmission through IP channels at speed of object motion up to 250 km/h.

Some companies including Qualcomm Corporation promote the technology of wireless high-speed access using the Flash-OFDM technology, which already got implementation in some countries in Europe. Its special feature is to ensure the relay transmission of mobile subscriber between the base stations interacting with it without breaking the connection (handover) even at a speed of up to 250 km/h. Another direction of technology development to achieve indices of the MBWA standard is utilization of adaptive antennas and adaptive methods of modulation (the iBurst technology). Application of progressive technologies allowed the transmission rate growth of up to 24.4 Mbit/s in downlink at operating channel bandwidth of 5 MHz and the increase of a radius of operating cell to several kilometers.

2.1.4 Global Networks of Mobile (Cellular) Communication (WWAN)

The cellular networks of mobile communication passed its development from the analog systems to digital ones, and then to wideband ones, increasing the list of offered services. The need to increase the transmission rate of the speech signal in the global network of the GSM mobile communication led to changing of signal transmission technology: the transfer from the channel switching to packets' switching. The next step in development of the cellular communication systems was the transfer from the utilization of narrowband systems to the wideband systems (cdma2000, WCDMA), which improved the effectiveness of spectrum application and increased the energetic potential of the link.

It should be noted that the mentioned separation is rather conditional, since, for instance, to WPAN we attribute systems with the operation radius up to 10...15 m, while the WLAN systems are intended for organization of networks with device remoteness within the limits of the local zone up to hundreds of meters, while on the base of IEEE 802.11 standard of local wireless DECT networks, we can successfully construct networks of the municipal scale.

From the comparable analysis of services rendering by the acting local networks (IEEE 802.11), radio access networks (IEEE 802.16 and IEEE 802.20), and the cellular communication networks (2G, 3G, and 4G), we can see the evolutionary development of the cellular communication networks. They provide the constantly growing spectrum of proposals but simultaneously require the infrastructure development of the networks with radio access, improvement of interconnection protocols, etc., which require significant financial investments. The rigid alternative of the cellular networks is WiFi networks (IEEE 802.11 standard), which can be autonomous and connected to any networks including wire network. These networks processing the small distance between subscriber terminals on the WiFi radio links at that, providing the data transmission rate of 54 Mbit/s, at present are used for creation of local networks and networks of access in Internet. WiFi networks include subscriber terminals and access points, which replace the base station. Access points provide the information exchange between subscribers of the specific case and external networks (PSTN, ISDN, network of radio access WiMAX). In the WiFi network, subscriber terminals often take the form of wireless adapters embedded in personal computers or connected to them through USB ports. For the simplest autonomous networks, when only two nodes of the network (operating radio stations) of access points are used, this is not required.

For large number of nodes, one usually uses the ad hoc mode, in which each node maintains the communication only with another (one) node. In such a mode, the minimal quantity of equipment is required, but the network lost its effectiveness and requires the introduction of special access points (allocated radio station) at operation inside bulky constructions. This is caused by essential attenuation of electromagnetic waves in the operating range of the WiFi systems in reinforced construction, which complicate the network structure making it porous (cellular), each of which includes the group of radio stations. Such a structure allows obtaining of the non-limited service zone as well as relay transfer of the mobile subscriber between subscriber terminal at its motion in the network (roaming) through the radio station of the access point (relay).

High-speed networks of mobile communication, which form the radio channel with the constant bandwidth, have the principal different methods of group signal formation (multiplexing). In 3G, 4G, and especially 5G systems, digital signals are generated from different sources.

To networks of municipal scale, we can attribute TV and radio broadcasting, trunking networks, and cellular communication. As to the IEEE 802.16 standard, introduced as a standard of regional communication, at present, networks of WiMAX standard are considered as the component part of the global system, which is competitive to systems of cellular communication of fourth-generation LTE.

2.2 Personal Wireless Networks

2.2.1 Networks of Bluetooth Standard

The system **Bluetooth** of the IEEE 802.15.1 standard uses the non-licensed frequency range of 2.4...2.4835 GHz (ISM range). This universal interface provides the wireless communication in the network of "ad hoc" configuration (equal with equal) with portable electronic devices located at small distances (communication of mobile phones, portable computers, headphones, printers, projectors, etc.). The variety and differences in properties of devices including in the Bluetooth system assume that connection should support the speech and data transmission and the transceiver should be miniature with low power consumption.

The type of connecting devices and their location in sufficient proximity assume the variety of connections in the "ad hoc" network and potentially lead to strong mutual influence. The absence of coordination between connections causes application

of wideband radio technologies. Two technologies of spectrum spread – DS-SS and FH-SS – are satisfied to these requirements. However, DS-SS turns out to be essentially more expensive and more complicated to implement; therefore, in the Bluetooth system, the technology of spectrum spread with spasmodic frequency hopping (FH-SS) and with duplex time channel division (EBB) is used. In system, at the time slot length of $625 \mu\text{s}$, we may arrange 79 frequency channels, each with bandwidth of 1 MHz. During time of each slot, the signal occupies one of these channels for transmission of the single packet as the request to the command from the leading device, which operates in accordance with the definite scheme of retuning. The reference Bluetooth network is called the pico-network. The total number of stations, which forms the pico-network, is not more than eight. Pico-networks may be partially overlapped, and separate stations may be simultaneously included in the structure of several networks or to temporarily displace from one network to another. The sequence of frequency hopping indicates the identification device number, and the system time indicates the hopping phase (the own phase shift is conferred to each station of the single pico-network). In different pico-cells, the different frequency retunings are used. Frequency hopping sequence should correspond to several requirements: they should have good statistical properties and stability to interferences; we must have much enough of these sequences. These contradictory conditions lead to the non-complete mutual orthogonality of sequences. The non-complete orthogonality can be compensated by application of coding and the **ARQ** (Automatic **Re**Quest for repeat) technology. Packets located in the consecutive time slots are transmitted using the Gaussian frequency modulation (GFSK) with the index $\beta = B_G \cdot T = 0.3$, where B_G is the bandwidth of the Gaussian filter. The maximal rate of data transmission is 1 Mbit/s.

The Bluetooth station can be in four modes: waiting, connection, parking, and analysis. In the waiting mode, the station is activated each 10 ms through each 1.28...3.84 s and scans the FH-channel in waiting of the paging signal. After pico-network identification, the controlled station can be switched in the parking mode, when it interrogates the FH-channel more rarely than in the waiting mode. In the analysis mode, the station scans the FH-channel more rarely. Station position in the listed modes, when the station only for the short time becomes active, essentially saves the consuming power from the power source.

The transfer from the waiting state into the connection mode occurs after the station in the single interval of 10 ms will detect in the FH-channel the access code representing the unique sequence (station identifier), which contains 32 frequencies. The calling station, which knows the identifier of the called station, remains in the waiting mode and repeatedly transmits the access code, transferring during 10 ms on 16 different frequencies. If the answer from the called station is absent, after several attempts, the calling station transmits the access code on remained 16 frequencies. If the calling station gets an answer from the called station in the form of the access code, then it transmits the synchronization packet in the frequency hopping mode **HFS** (Frequency **Hop** Synchronization), which contains an identifier and the sequence of hopping frequency retuning for both stations.

2.2.2 Sensor Networks

The IEEE 802.15.4 standard defines the physical layer (level) and the access control to the array of wireless personal networks and is the basis of sensor network protocols [43].

Construction of different level networks and the promising projects of the global network structure of the mobile communication make the development of technologies and technical equipment of wireless connections to the networks of the mobile communication a rather relevant task. Developing devices of wireless connections can be used in networks of the mesh type and for IP related to various standards (IEEE 802.15.4, IEEE 802.11), which operates in different frequency ranges and applies various protocols.

The main applications of wireless connection technologies are in the spheres: portable domestic devices, remote control of devices, domestic medical devices, control of power consumption, wireless audio devices, safety, and radio-frequency identification.

Depending on the applied technologies of connections and data transmission rate, the transceivers, which are usually realized as the single-chip structure, provide different communication distances between subscribers (Fig. 2.2 on the example of IC from Texas Instruments).

The Texas Instruments company is one of the greatest IC manufacturers for wireless communications.

The ZigBee standard has got the largest distribution on the domestic level. The main idea of this standard is to offer the lower levels of wireless personal networks, which are oriented to low cost, and the low data transmission rate (in contrast to WiFi) between devices at the absence of basing structures at the low level of power consumption.

The ZigBee standard is based on the radio protocol IEEE 802.15.4 [43] and provides the creation of static wireless networks with large number of subscribers (to 65,536) and the low time of their connection to the network (less than 30 ms). It

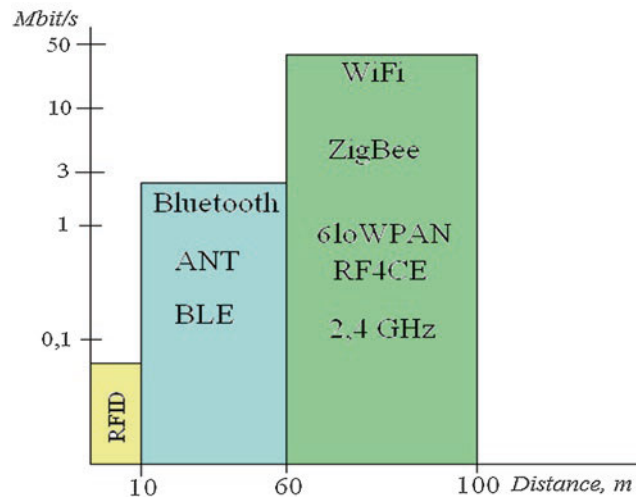


Fig. 2.2 Regions of various data transmission rates as a function of the distance

is assumed that ZigBee will be claimed where there is no necessary in the high rate of data transmission (to 250 kbit/s), and the traffic volume is small. In particular, it is optimal for systems in which a significant number of wireless sensors are used in the doze mode. One of the most important functions of the standard IEEE 802.15.4 network is the operation performing in the mode of real-time scale by means of time slot saving and prevention of the simultaneous access and with complex protection of networks. The network structure elements should realize the control functions of energy consumption and provide the device transfer from the active mode to the “sleep” mode and return into the active mode according to the calling signal from the base structure, as well as to provide checking of the connection quality. Devices, which are compatible with the IEEE 802.15.4 standard, can use one of the three possible non-licensed bands for operation:

- 868.0...868.6 MHz (Europe, the single communication channel)
- 902...928 MHz (Northern America, up to 30 channels)
- 2400...2483.5 MHz (used all over the world, more than 16 channels).

Initial standard version (2003) anticipated the operation in two frequency bands of 868/915 MHz and 2450 MHz with operation of the wideband modulations with the direct spectrum spread (DSSS) with the further transfer to the more effective modulation approaches: the binary phase shift keying (BPSK) and quadrature phase shift keying (QPSK). The transmission rate in the first two ranges was 20 and 40 kbit/s, while in the range of 2450 MHz achieves 250 kbit/s. To arrange the high-rate data transmission by the IEEE 802.15.4a standard, the following technologies found application: UWB (Ultra-WideBand) and CSS (Customer Switching System). Ultra-wideband technology is used in the following frequency ranges: below 1 GHz, 3...5 GHz, and 6...10 GHz, and the CSS technology is used in the band of 2.4 GHz of the ISM range.

Introduction in 2009 of IEEE 802.15.4c and IEEE 802.15.4d standards increased a number of available physical layers (levels) including the 780 MHz range with quadrature phase shift keying (QPSK) and with multi-phase shift keying (M-PSK) and in the 950 MHz range, the Gaussian frequency shift keying (GFSK) or the binary phase shift keying (BFSK).

Networks of ZigBee IEEE 802.15.4 protocol are realized in any area, which requires the wireless connection:

- Automation of operation of residential and industrial buildings.
- The system of ecological information gathering.
- Medical instruments, patients’ monitoring.
- Logistics, tracking of freight displacement.
- Sensor networks, active radio-frequency identification.
- Monitoring and control of power consumption.

Networks of the IEEE 802.15.4 standard can be constructed on the equitable structure or according to the star. At any structure, at least, one coordinator should be in the network, i.e., the network is formed by the devices’ group, each of which has a unique identification number. The peer-to-peer networks can form the arbitrary structures and are intended for the

creation of the so-called self-resetting network of the mesh-decentralized network technology, which reminds of Internet in the operation principle. It permits to find independently the new ways of signal transmission in the network in case individual nodes interrupt their operation.

This standard defines two types of nodes at construction of the sensor networks: the full function device (**FFD**) and the device with light functions. In the first case, the node can perform functions of the coordinator of personal networks and as the general node. For this communication model, it provides the interface with other devices and with the equipment of higher level (to act as the coordinator). In another variant, this is the node with light functions, the medium resource and requirements to the network, which provides the connection with the coordinator only.

Their interconnection with each other is provided by creation of the simple conceptual network. The structure of network layers is described by the OSI model, which standardizes the physical level, and interconnection with the higher layers is realized through the logic connection of the MAC sub-level.

Information transmission is performed by fragments, the type of which is defined as: data, confirmation, a flag, and the fragment of commands of the access mechanism. Flags are used at the super-fragment slot structure, which defines boundaries of its structure providing the synchronization of other devices and the network configuration. The data transmission to the coordinator requires the flag synchronization usage, which is performed by the **CSMA/CA** (Carrier Sense Multiple Access with Collision Avoidance) mode. The data transmission from the coordinator is usually accompanied by requests to devices, and after obtaining of the flag, which indicates the end of fragment transmission, the coordinator confirms the request and sends informational packets.

The example of creation of the sensor network of residential premises of the “smart home” is the network on the base, for instance, of CC2520 and CC2530 transceivers implemented in **SoC** (System-on-Chip) technology of the Texas Instruments company. The premise input door is equipped by the sensor of the access check, which controls with the help of the single electronic key the following: the input door lock, control of signaling system, and the illumination. Sensors located in different points of the premise can perform the following using the safety system: control by illumination as well as change the illumination brightness by complex devices and illuminating systems. The signal from this sensor and/or from the remote control panel permits to put in action the device for blinds control. The sensor of environment parameters fulfills the permanent monitoring of external temperature, moisture, atmosphere pressure, and illumination, which permits optimization of the domestic system operation taking into consideration external factors. Using the network of wireless sensors, the optimal temperature with energy saving is supported. The system of the fire safety provides the monitoring of temperature and smoke formation in the premise by means of sensors connected to the illumination nets. The DSL modem is the connecting element with the Internet for remote monitoring and control of the domestic network. Any variation of early-chosen values not from the control panel located in the premise through the GSM portal is sent to the subscriber in the form of a message about the operation condition change of as the alarm signals on the mobile devices of the cellular network.

The *6LoWPAN* protocol describes the open standard and is intended for operation arrangement in mobile communication networks with low energy potential under the IPv6 control. The 6LoWPAN technology operates in networks of the mesh types and allows creation of large-scaled networks with IP nodes addressing. This technology works at different physical levels in frequency bands up to 1 GHz and on frequency of 2.4 GHz over the IEEE 802.15.4 protocol level and provides an application of the IPv6 addressing type and creation of the IP content.

The single-chip two-mode IC CC2567 is the highly integrated module (Table 2.1), which uses opportunities of the Bluetooth chip and the transceiver ANT+ having a very low power consumption: **ULP** type (Ultra-Low Power).

Table 2.1 Parameters of several ICs

IC type	Technology	Operation frequency MHz	Transmission rate, kbit/s	Receiver sensitivity, dBm	Suppression of adjacent channel, dB	Suppression of channel, dB	Bloc-king, dB	Modulation
CC2530	ZigBee	2500	32/64/256	−97	49	57	57	
CC2560	Bluetooth	2500	2178	−93				
CC2567	Bluetooth	2500	1000	−93	5 (±1mRII)			
CC2540	Bluetooth	2500	1000					GFSK
CC2533	RF4CE	2500	250	−97	49	57	57	
CC8520	PurePath	2500	5000	−83				
CC1020	Sub1-GHz	402/470, 804/940	2,4	−116		49	78	OOK, FSK
				−118				
CC1100E	Sub1-GHz	< 1000	1,2...500	−112				MSK, FSK
WL1271	WiFi	2400	65,000	−76				
TMS3705	PaLFI	0,1342	8					Ask

Owing to the utilization of the *ZigBeeRF4CE technology* in the single-chip IC of the CC2533 type, it provides fulfillment of the two-way communication functions at minimal power consumption. At that, the communication distance also increases, the reliability grows, and the possibility of remote control is saved for objects, which are outside of the direct vision zone. Devices implemented on CC2533 are used for the checking and accounting of the energy consumption and for control of TV systems and CD players and being built into glasses, for formation of the three-dimension image.

The *PurePath technology*, at utilization of two and more CC8520 devices, allows creation of the wireless audio network functioning in the range of 2.5 GHz and providing the quality of CD audio. One device fulfills the functions of the signal source and others the functions of receivers. The built-in audio-network protocol with error correction and masking at data transmission provides the compatibility with other wireless systems, owing to the utilization of different techniques: the adaptive jumping change/adjustment of the channel frequency, the anticipatory error correction, and buffering and recurring data transmission.

The single-chip transceiver with FSK/GFSK and OOK (**On-Off Keying**) amplitude shift keying, which is realized on the IC of CC1020 type, has the operating frequency band in 12.5 kHz per one channel and high selectivity (up to 71 dB at offset of ± 2 MHz). It is used in the narrowband radio communication systems (great communication distance, semi-duplex communication systems, communication systems with jumping frequency change), wireless safety systems, and telemetric systems, for premises automation.

The integral circuit CC1100E with high density of components is the transceiver with the operating band in the non-licensed frequency sub-ranges in ISM range below 1 GHz, which has small output power of the transmitter. It has low cost, small current consumption in waiting mode (200 nA), and low transfer time into the active mode (240 μ s).

The *passive low-frequency interface* **PaLFI** TMS37157 allows realization of two-way communication on small distances without application of power source on one of the sides. The part of energy radiated by the base station is the power source for the receiver and permits it to radiate the response signal. Depending on system parameters (the antenna dimensions, the power radiated by the base station), the receiver can provide the necessary power to the micro-controller MSP430 on the distance of 1.5 meters. Such facilities of IC allow its application in devices requiring the periodic reconfiguration or actualization, in medicine as biosensors, as well as the recharging device on the receiving side during communication sessions.

2.2.3 The System of Ultra-Wideband (UWB) Communication

The *ultra-wideband (UWB) communication* at small distance between equipment inside the premises (less than 10 m) fundamentally differs from other radio-frequency systems because it provides the wireless connections without carrier frequency utilization [70].

Signals, which have relative bandwidth equal or approached to 1, are called ultra-wideband ones. The relative frequency bandwidth is defined as:

$$y = \frac{f_{\max} - f_{\min}}{f_{\max} + f_{\min}}, \quad (2.1)$$

where $f_{\max}, f_{\min}$ are the highest and lowest components of the received signal spectrum, usually on the level -3 dB. In this classification, under narrowband signals (systems) we understand those signals, which have $y < 0.01$, the wideband signals have $0.01 < y < 0.25$, and the ultra-wideband signals have $0.25 < y < 1$. A special feature of ultra-wideband (UWB) systems is the network structure that provides communication at a short distance between objects inside the room (less than 10 m). The radio interface of a terrestrial network is fundamentally different from other radio frequency systems, since the radio channel uses a noise-like signal created by a sequence of short pulses. Depending on the method of ultra-wideband signal formation, we may distinguish:

- Signals having the carrier frequency containing up to four periods of harmonic oscillation per transmitting signal.
- Sequences of very short pulses (less than 1 ns) having the non-sinusoidal shapes (video-pulses, pulse of exponential shape, Gaussian pulses).
- Multi-frequency signals formed by fragmentation of an initial signal into several partial.

As we see from the described classification, the concept of “carrier frequency” frequently corresponds to the average frequency (simple mean) of the transmitting signal.

The simpler criterion of signal broadbandness was adopted by **FCC** (Federal Communications Commission), according to which it is accepted to consider as ultra-wideband all signals having bandwidth of more than 500 MHz on the

level -10 dB. In this case, f_{max}, f_{min} are the highest and lowest spectrum components calculated on the level -10 dB, and the average frequency is defined as $f_{aver} = \sqrt{f_{min} f_{max}}$.

The European Electronic Communications Commission (ECC) allocated the radio-frequency spectrum from 3.1 to 10.6 GHz for ultra-wideband systems (any radio transmission technology with the occupied spectrum more than 20% from the central frequency, but not less than 500 MHz).

Telecommunication systems using the ultra-wideband signals have a series of advantages compared to existing telecommunication systems. Devices that form the radio signal in UHF range have a complicate construction, high cost, and high consuming power. Besides, they are less reliable compared to UWB devices, which operate at levels below the noise levels of traditional radio systems, have a minimal number of external components, and do not demand special requirements to equipment parameters. This allows significantly simplification of constructive solutions not only to satisfy the physical layer protocol (PHY) but in order to create, for example, the UWB transceiver with the flexibly adjusted in-out controller in the single device, which realizes a wide field of applications (the large range of rates, a variety of channels, registration of localization with an accuracy up to several centimeters, etc.). The network realizing on the base of UWB can function as the personal network (WPAN), as the local computing network (WLAN), and as the global mobile network (WWAN). Since the emitting power level does not exceed the level of the power spectral density, which is defined by the appropriate regulations as the admissible, i.e., not exceeding the level of natural interferences of an environment, this allows UWB systems to operate in frequency ranges used by other radio systems (Fig. 2.3).

The digital flow rate in UWB systems should be essentially increased compared to existing systems, which is caused by the large bandwidth (the Shannon formula):

$$R = B \log_2 \left(1 + \frac{P_s}{P_n} \right), \quad (2.2)$$

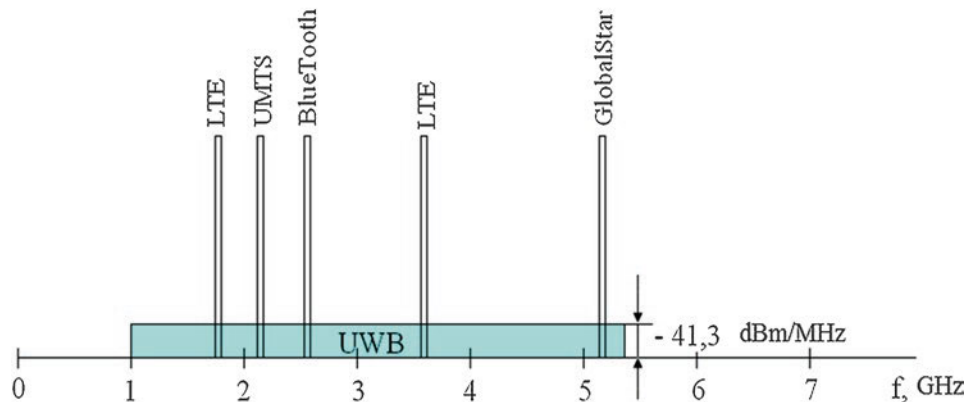
where B is the signal band, P_s is the signal power, and P_n is the noise power, since theoretically the upper limit of the capacity increases with bandwidth growth. Together with carrying capacity growth of UWB systems, we can simultaneously organize a large number of channels in the operating frequency band.

Another important advantage of UWB systems is the relative simplicity of a struggle with the multipath property distortions by means of the simple increase the pulse repetition interval, which simultaneously leads to reduction of the information transmission rate.

Secretiveness of information transmission in UWB systems and their stability to suppression predetermined its application in systems of special military destinations. In these systems, the impact of narrowband signals onto the useful signal is strongly relaxed due to the wide operating frequency band, and the pulse interference impact is relaxed due to its localization in time and relatively due to the small probability of coincidence with pulses of the useful signal.

The main distinguishing property of UWB systems is the signal shape used for information transmission. If in radio communication systems, which are considered earlier, the harmonic carrier oscillations are used providing the effective power transmission to the subscriber, the wavelets are the main class of signals for UWB systems since such a function is the eigenfunction of any linear medium. The wavelet which is a small wave (small oscillation) and the function $\psi(t)$ oscillating in the vicinity of the some point are effectively used at digital signal processing in radio physics, radar technology, and radio vision [4]. In the ultra-wideband telecommunication systems, the bipolar Gaussian pulses

Fig. 2.3 Frequency positions of various radio systems
Upper limit of the capacity increases with bandwidth growth. Together with carrying capacity grows of UWB systems, we can simultaneously organize a large number of channels in the operating frequency



$$s(t) = 2\sqrt{e}U t f_0 \exp\left[-2(\pi f_0 t)^2\right] \quad (2.3)$$

are the most distributed models of the transmitting signals, where f_0 is the average frequency and U is the signal amplitude.

Using the normalized frequency value f_0 through the relative pulse duration $\tau_0 = 1/f_0$, Eq. (2.3) can be presented in the form

$$s(t) = \frac{2\sqrt{e}U\pi(t-\tau_0/2)}{\tau_0} \exp\left\{-2\left[\frac{\pi(t-\tau_0/2)}{\tau_0}\right]^2\right\}, \quad (2.4)$$

describing the shape and the normalized pulse spectral density (Fig. 2.4).

Another radio signal model offered by Gauss and used in the UWB systems, which can be also attributed to the wavelet functions, well describes the real signals (2.3):

$$s(t) = U \exp\left[(-a\pi/2)t^2\right] \exp(j2\pi f_0 t) \quad (2.5)$$

where a is a parameter defining the envelope shape and also the effective pulse duration.

The signal model shape and its normalized spectral density for $a = 2$ are presented in Fig. 2.5.

As in the spectrum in Fig. 2.4, this model does not contain the DC component, and the larger part of power is concentrated near the f_0 frequency. The absence of the direct connection of the effective pulse duration τ_0 and the average spectrum fre-

Fig. 2.4 The time shape (a) and the normalized pulse spectral density (b) of the bipolar Gaussian pulses

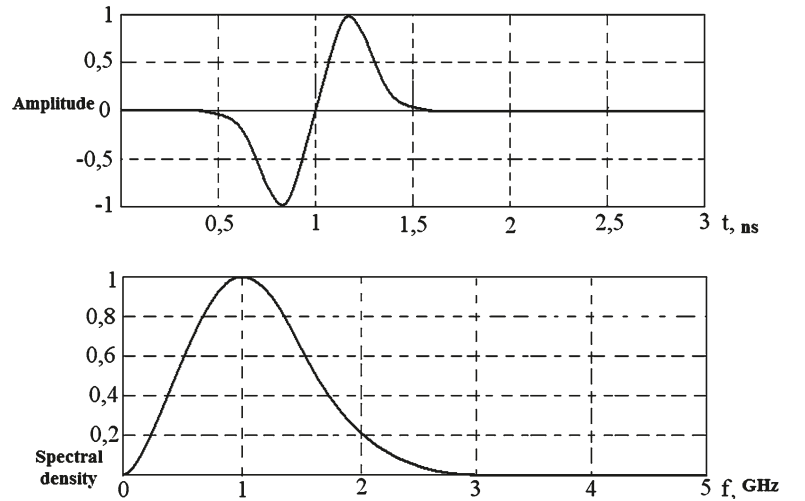
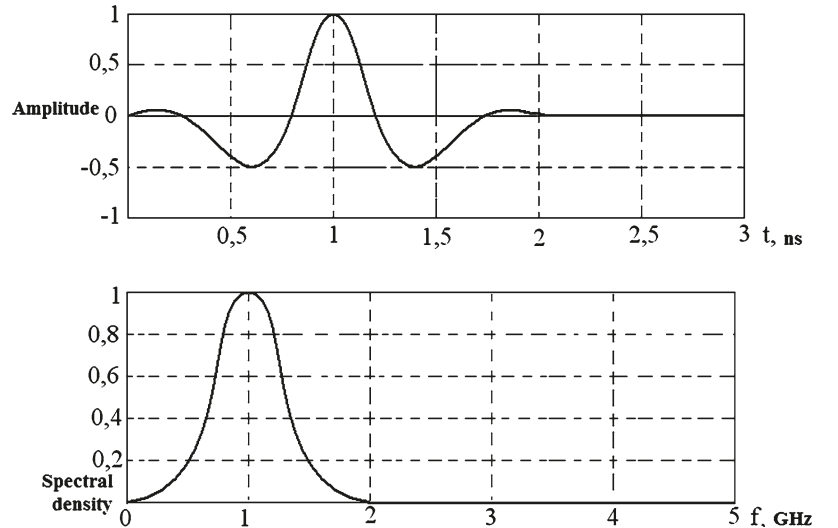


Fig. 2.5 The signal model shape and its normalized spectral density for $a = 2$



quency f_0 allows application of this model for description of UWB systems, which use the orthogonal frequency multiplexing (OFDM).

Ultra-wideband systems can be divided according to:

- Destination.
- Type of the transmitting signal.
- The method of radio-frequency spectrum usage control.
- Transmission rate and communication distance.

Depending on the UWB system *destination*, we can mark out the low-rate lines with a large operation radius and high-rate lines used for communication organization on the *small* distances (e.g., the system of wireless USB standard fulfilling the interface functions for computers, mobile phones, computer periphery, and devices of domestic electronics).

Another type of systems with a small operation radius intended for information transmission or for control signals transmission, to which we can refer, for example, the safety systems, medical applications (heart operation and other organs monitoring), wireless driver “assistants,” etc., is classified according to the transmitted signal type.

Systems that are intended for object localization determination use the principles of the ultra-wideband radar technology. On the base of this technology, it is possible to develop devices for image transmission at buildings and installations, construction, and repair.

Telecommunication UWB systems can be used for recognition of marks, identification cards, etc. during the inspection of logistic escort schemes of the moving equipment.

Dividing of UWB systems according to the *type of transmitting signal* (pulse of multi-frequency) most often determines their construction. *Pulse* signal represents the sequence of ultra-short (down to units of nanoseconds) bipolar pulses, which are modulated in time for the position, the pulse duration, the amplitude, or the phase. *Multi-frequency* UWB systems use the signal representing a sum of several orthogonal signals with application of frequency-divided channels, which are comparably narrowband.

Systems using pulse signals can be divided into systems with the constant period of pulse repetition and with its modulation. The example of the UWB system of the first type is the system [4], in which for “1” transmission of the monocycle is used while for “0” transmission of the opposite signal is used. Such systems are widely used in radar technology providing the advantage by 3...6 dB compared with other systems under conditions of the multipath propagation absence. At the peak factor less than 3, this allows decreasing the transmitting power and a possibility to apply ICs on the base of the CMOS technology, which have a low level of consuming power.

UWB systems with *varying interval* of pulse repetition include systems with pulse position modulation (**PPM**), when the single symbol is transmitted with the help of the single pulse and their variety **TH-UWB** (**T**ime **H**opping **UWB**) with jumps in time and **DSS-UWB** (**D**irect **S**pread **S**pectrum **UWB**). In this case, several pulses of transmitting data, which have nanosecond or sub-nanosecond duration, correspond to the single symbol. In order to reduce the discrete components’ level in the spectrum of the UWB signal, we can use the additional encoding of time intervals by pseudo-random codes. This allows the fulfillment of channel. Для уменьшения уровня дискретных составляющих в спектре UWB сигнала применяют division in systems with the wideband, similarly to the CDMA technology.

According to the *method of spectrum spreading*, UWB telecommunication systems can be divided into systems **FHSS-UWB** (**F**requency **H**opping **S**pread **S**pectrum **UWB**) and systems with orthogonal frequency multiplexing **MB-OFDM** (**M**ulti**B**and **O**FD**M**).

In systems with frequency hopping, data are transmitted in sequence in different frequency bands of the operating range in the form of short single pulses or signals with orthogonal frequency shift keying (OFDM) or linear frequency modulation (**LFM**). When using LFM, all frequency band is divided into N -sub-bands, in each of which there are N signals (codes) containing N elementary signals. The specific feature of such elementary signals is the continuity of the signal envelope in time and symmetry of this envelope spectrum with regard to $f = 0$.

Ultra-wideband telecommunication systems, which are used in the OFDM technology, are characterized by the structure similar to the narrowband systems, but in contrast to them, they may have significantly a larger offset between sub-carriers. The peculiarity of UWB systems with OFDM modulation is the excellent beam resolution and, relatively, the steadiness to the multipath propagation, which is accompanied by a large signal peak factor.

According to *their destination*, UWB systems can be divided into:

- Communications systems and measurements systems.
- Systems for detection (the radar of sub-surface sensing, systems for internal organs inspection, for locality monitoring, for objects detection in walls and through walls).
- Radar systems on the transport platforms.

Systems intended for data transmission and measurements can be used both for private utilization (organization of the domestic network, interfaces of multimedia devices) and for business (wireless local networks with increased capabilities, wireless personal networks). Such systems form the main quantity of used UWB systems and operate in the spectrum segment below 10.6 GHz, together with other functioning of more powerful systems in the same segment.

Radar transport systems operate in the range over 20 GHz, which is free enough from the point of view of electromagnetic compatibility (EMC).

Division of UWB systems into *low*-, *medium*-, and *high-speed systems* is rather conditional and is caused by the access methods applied in each groups, the methods of generation, and the processing of UWB signals. Low-speed signals (to 100 kbit/c) are oriented to application in telemetric and telematic systems, intended for data acquisition from sensors, which are remotely to distances of several hundreds of meters in networks of various destinations. Medium-speed (1...100 Mbit/c) systems with operation radius of 10...100 m, which form a competition to wireless local area networks (WLAN) of IEEE 802/11 a/b/g standards and have a lower power consumption and lesser mass and dimensions, provide the transmission secretiveness. High-speed systems (over 100 Mbit/s) with operation radius of 10 m are used for the exchange of large data arrays between computers, video cameras, digital video-photo cameras, etc. and between mobile and fixed devices on the wireless technology. Such systems have a lesser power consumption and capability to transmit information through the walls at larger distance between objects compared to infrared and laser systems.

Unwanted emission is the main factor, which determines the communication distance and the transmission rate and, accordingly, the system class. The limit power (P) created by the electric field of the isotropic radiator (in $\mu V/m$) at the distance r (in m)

$$P = \frac{E_0^2 4\pi r^2}{\rho_v} \quad (2.6)$$

is the value of limit the emission in the non-licensed frequency ranges, where $E_0(\mu V/m)$ is the intensity of electric field measured at the distance $r(m)$ in the sphere radius, in which the field is measured, $\rho_v = 377 \text{ Ohm}$ is the electric resistance of the vacuum.

In accordance with FCC requirements, the electric field strength is measured in the frequency area above 3.1 GHz at the distance of 3 meters in the band of 1 MHz, and the component of $500 \mu V/m$ corresponds to the limit value of the spectral power $P_{allowed} = -41.3 \text{ dBm/MHz}$ (Fig. 2.3). For UWB systems, the maximal admitted value of the average transmitter power is $P_{Tx} = 1.48 \cdot 10^{-4} \text{ W}$ (at transmission by the pulse signals in bandwidth of 2 GHz and the signal spectrum bandwidth not less than 2 GHz).

Comparison of different systems of the radio access with telecommunication UWB systems shows that the transmission rate in the pulse ultra-wideband systems depends on the modulation type, the period duration, and the signal duty-off factor, as well as on the signal base. So, at pulse duration of 1 ns, the duty-off factor of 2 and using the PSK2 signals, the transmitting rate is $R = 500 \text{ Mbit/s}$, while the application of the complex signal in the form of the sequence of ultra-short pulses with the base 127 decreases the rate to 4 Mbit/s. So, great decrease in the transmission rate leads simultaneously to appearance of advantages peculiar to systems with complex signals: a possibility to apply the CDMA technology, steadiness to distortion due to the multipath propagation, etc. The application of MB-OFDM technology ensures the highest transmission rate, which can be achieved in systems with the radio access.

The spectral effectiveness, i.e., the ratio of the transmission rate (bit/s) to the frequency band occupied by a signal, is the important index of telecommunication systems. Telecommunication UWB systems, at high transmission rate (more than 100 Mbit/s), which have a wide bandwidth, usually show spectral effectiveness not more than a unit.

The operation distance of the UWB system mainly depends on the limit value of the emitted power, and for pulse systems, it is comparable with systems of wireless radio access of the Bluetooth and ZigBee standards, while for multi-frequency systems with the OFDM technology, it achieves hundreds of meters (as in systems IEEE 802.11b/g).

Comparing two types of UWB systems, we can note that it is typical for pulse systems to have a high transmission rate and the simplicity of generation and processing of signals at high steadiness to the multipath propagation and the large number of subscribers. Simultaneously, it is peculiar for such systems to have a low spectral effectiveness at restricted operation distance and at synchronization complexities.

For multi-frequency systems, procedures of signal generation and processing are applicable, which are peculiar to narrowband systems, the larger transmission rate and operation distance, the high enough spectral effectiveness, and the spatial capacity. Typical shortcomings of such systems are the equipment complexity, the high peak factor, and the complexity of synchronization systems.

The main destination of these promising systems is the creation of unified communication systems with network architecture inside the buildings, which provides access and information exchange between devices without adding attraction of experts for its adjustments, and the possibility to depict, to edit, and to listen to the received information. In the existing

infrastructure, the three main equipment categories can be distinguished: computers, multimedia centers, and the mobile devices. At that, personal computers together with a periphery and also with modems, gateways, and routers are the primary elements of centralized networks; the entertaining cluster (the domestic movie house, audio and video equipment) and mobile devices (smartphones, notebooks) should have the possibility to easily displace inside the building.

2.2.4 Systems of Radio-Frequency Identification (RFID)

Under the term *radio-frequency identification (RFID)*, we understand systems using radio waves for reading of the unique identification number (a tag) of the object or a person. These systems are concerned with the automatic identification (auto-ID) technology including optical readers (recognition of text and bar code) and biometrical technologies (identification by fingerprints and by voice). The RFID system [18, 51] enables to read and transmit data in the computer system in real-time scale without participation of a person.

A variety of rendering services used by this technology and frequency ranges essentially complicate the standard and interface protocol formation. The International Organization for Standardization (ISO) together with the International Electrotechnical Commission (IEC) developed a series of RFID standards for automatic identification and checking of objects ISO/IEC 1800 worldwide. It includes seven main radio frequencies of the RFID technology worldwide beginning from 135 kHz to 930 MHz, including general parameters of the interface for worldwide adopted frequencies. The Auto-ID Center developed the electronic product code (EPC) providing the product tracking in the unified logistic system using different types (classes) of tags. At that, each from the five following types of tags should be more complicate than the previous. Assimilation of the microwave range for utilization in RFID systems, which are included into the global network EPCglobal, led to the development of specifications intended for execution of operations in the distribution networks possessing a wide set of applied software.

The basis of the RFID technology is application of a tag or a transponder (*transmitter-responder*) containing some information (for instance, about a product, manufacturer, the realization period etc.), which are transmitted to a reader during the interrogation. Usually, a tag includes two components: an integrated circuit for information storage and processing, modulation/demodulation of the radio signal, etc. and an antenna executing the signal reception and transmission. At radio signal transmission by a reader, the transponder (a tag) receives it and generates the output signal (reflects the radio signal). At reading of the tag, the data transmission occurs from its memory into a computer where it is processed to the form suitable for perception by a consumer.

Radio-frequency identification, as a matter of fact, on the IC of TMS3705 type is the integrated transceiver intended for support of the semi-duplex mode **DX (Half-DupleX)** relays (transponders) or the mode of full low-frequency duplex (LF dual interface). The transmitter of this identifier has the loop antenna in the form of a ring with the symmetric feeder or the ferrite antenna. Directional properties of this antenna allow provision of signal reception on distance of up to 10 m. The transmitter of the RFID transceiver uses the 100%-amplitude shift keying **ASK (Amplitude Shift Keying)**. On the receiving side, a digital frequency demodulator is used, the output signal of which is simultaneously a control signal for the auto generator that changes its frequency (VCO). RFID system is constructively made in the form of a miniature device in waterproof package, and it is located on the band with fixing layer. Application area is the following: tracking for containers (freights), for animals, industrial automatic, and check of the access (including the motor-car immobilizer).

The receiving part of transceivers of the considered means of wireless connections, as we see from Table 2.1, is implemented on the super-heterodyne architecture or on the direct conversion architecture. Single-chip IC of transceivers usually used the OFN-40 package with dimensions of 6 × 6 mm.

2.3 Local Networks of Wireless Communications

2.3.1 Wireless Communication Local Area (WLAN)

A local computer network, which includes the following elements: servers, terminals, printers and other peripherals connected to each other by copper or fiber-optic cables, is the prototype of modern “mining”. Such a network including a group of closely located computers integrated between themselves and connected to the Internet network through **GPON (Gigabit Passive Optical Network)** significantly increases the computing process speed. The splitting of big data arrays into blocks (block chains) facilitates this. The essential shortcoming of such networks is the fact that at each variation of the network

configuration, it requires laying of additional cables. This increases the cost for networks' exploitation and restricts the terminal displacement in the space.

In contrast to these, the wireless local networks (WLAN) are intended for *digital data exchange* between computers, i.e., between premises, and they have such advantages as:

- The network flexibility, when network nodes can be installed on a significant distance (up to 100 m) and outside zones of the direct vision.
- The simplicity of network planning, while networks of the “ad hoc” type with the episodic access of the arbitrary structure do not require the configuration. Another type of the WLAN is the structured network having a wired infrastructure, which connects it with other networks. Points of the high-speed wireless access to the Internet (WiFi hot spots) are the important components of such networks, and they provide connection of local network stations (nodes) to the Internet.

The network can be both autonomous and connected with networks of any other types including wired. In many cases, the WLAN plays a role of the gateway between a computer, a smartphone, and the Internet, which requires the compatibility of such networks for subscriber applications.

WLAN represent a large choice of radio technologies, transmission rates, frequency ranges, and possibilities of applications. The first standard of the wireless access to computing networks **HIPERLAN** (**H**igh **P**ERformance **L**ocal Area Network), which was adopted by the European Telecommunications Standards Institute (ETSI) in 1995, defined requirements to the interface commensurable with the OSI standard. The specific case of WLAN is the **WiFi** network, which unifies the IEEE 801.11b and IEEE 801.11 g [10] standards. Local networks may be organized in the non-licensed frequency range ISM 2.4 GHz indented for utilization in industrial manufacture, in medical institutions, and in scientific laboratories. Interfaces of the physical level of HIPERLAN and **WiFi** standards coincide. The necessity of the data transmission rate increase in the WLAN related to assimilation and mastering of infrared frequency range in the region of 5 GHz leads to the widening of these networks' possibilities on the physical level. To increase the noise immunity of signals, we can apply spreading of the frequency band occupied by the radio signal with utilization of modulation based on the spectrum spread method of direct sequence DSSS or on multiplexing by orthogonal carriers OFDM. The last method for spectrum spreading of the radio signal is used with the limited number of sub-carriers in the IEEE 802.11a standard for the frequency range of 5.1 GHz, as well as in many other systems.

The technology of the *local wireless network* WLAN for communication is intended for creation of the computing network, which has the terminal mobility and a possibility to plan easily and to change its configuration. Application of the radio channel for data transmission instead of cable network leads to the growth of the error probability up to $10^{-4} \dots 10^{-3}$, the increase of the unauthorized access danger to transmitting information, the higher cost of these networks' equipment. For prevention of the error appearance, we can use the **FEC** (**F**orward **E**rror **C**orrection) technology.

Classifying the WLAN according to configuration of creating networks, we can mark the structured networks and networks of the “ad hoc” type.

The structured network has the wired infrastructure, which connects it with other wired networks. The important component of such networks is the presence of access points (ports) providing the interaction of wireless terminals with wired network infrastructure (Fig. 2.6).

In most cases, radio communication between wireless terminals is carried out only through the radio access point, which acts as a base station in cellular networks, providing control over data transmission. Only the signal processing procedure is retained for terminal devices (computers). This allows prevention of collisions at access to the network under the condition that access points control the channel. However, such networks have no flexibility due to attachment to the wired network.

The computing WLAN of ad hoc type (Fig. 2.7) has no wired infrastructure, and each computer plays a role of the network node. Interaction between them is fulfilled permanently until terminals are in the zone of radio access (approximately, 100 m), and at some interface complication, it provides communication with other networks. In this case, both transmitting stations send pseudo-random signals generated by the positive sequence method at the input of modulator or using the method of hopping the frequency of carrier wave, which has different phases.

2.3.1.1 Network of the HIPERLAN Standard

For WLAN technology, independently on the applied network configuration, the problem of hidden station (node) is typical, when some station is situated at the same time in two networks or the direct connection with the station belongs to another network. Another solution to the problem of the appearance of a hidden station is the use of spectrum expansion technology

Fig. 2.6 Interaction of wireless terminals with wired network infrastructure

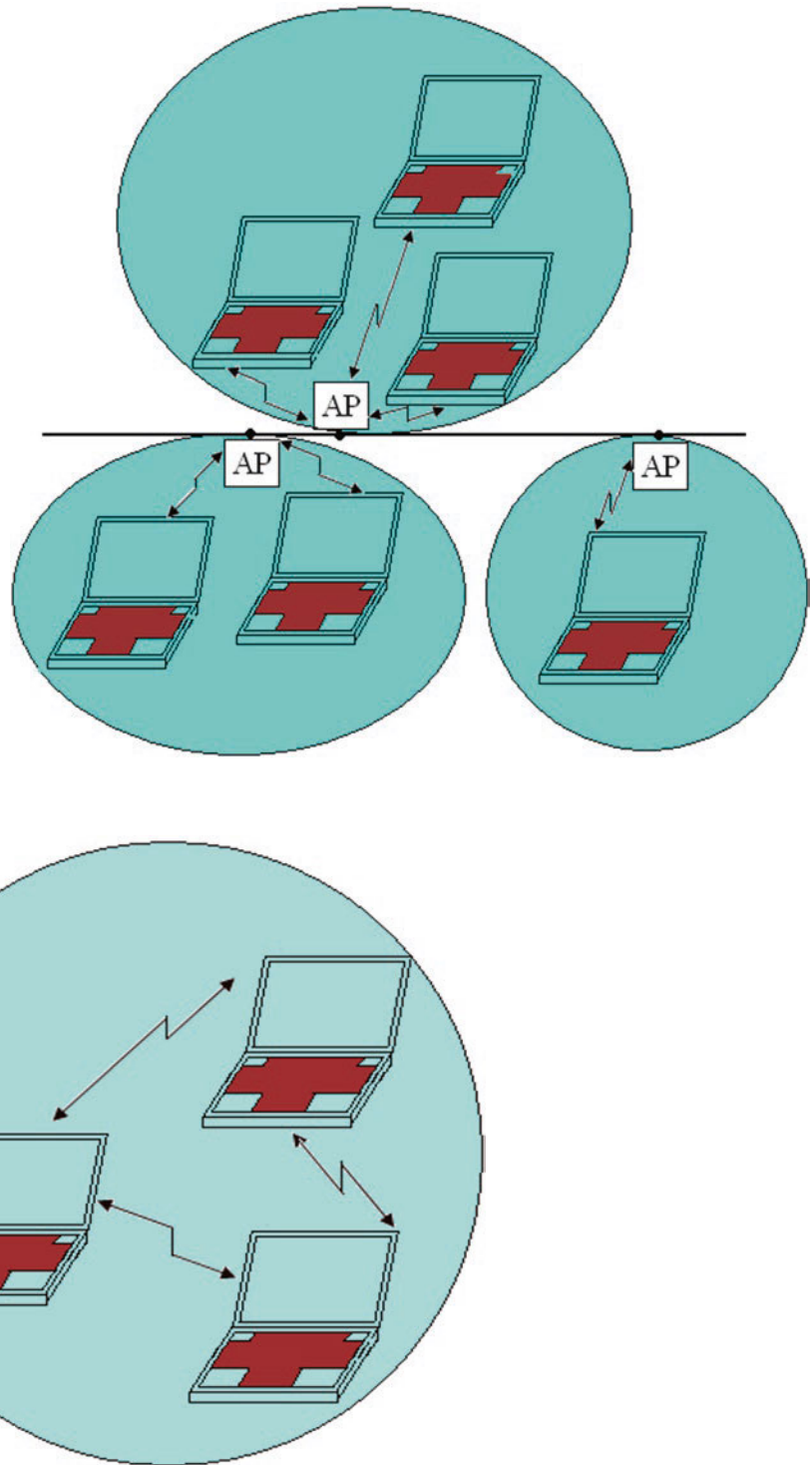


Fig. 2.7 The computing WLAN of the “ad hoc” type

at the physical level (in the structure of OSI reference model). The problem of hidden station can be also solved on the physical level using the technology of spectrum spreading. In this case, both transmitting stations send pseudo-random signals (the direct sequence method or the frequency hopping method is used) with different phases. Owing to practically their full mutual orthogonality, they are easily detected in the receiver using the correlation processing methods of the RAKE receiver. In this case, no preliminary requests on transmission are required.

The HIPERLAN/1 network operates in the range of 5.15...5.3 GHz divided into five frequency channels. Data transmission can be performed in synchronous and asynchronous modes with the maximal transmission rate of 23.5 Mbit/s. The maximal speed of terminal motion can be 5 km/h. Data transmission via the radio channel is performed using Gaussian PSK modulation with minimal frequency shift of the carrier oscillation (GMSK). The filter characteristic satisfies the relation $BT = 0.3$, where B is the signal spectrum width on the level -3 dB and T is the modulation period. Owing to the multipath property and the high transmission rate, the intersymbol signal interference arises on the demodulator output. Distortions of such a type are minimized by application of the adaptive corrector. The corrector used in the receiver of the HIPERLAN/1 standard constructively is made on two filters: the linear filter and the filter with deciding feedback. The linear filter forms the signal in its output including transients reflecting the impact of several last pulses. Then, on the base of these symbols' decoding, the filter with deciding feedback synthesizes a signal caused by intersymbol interference and subtracts it from the processing signal in given moment.

The HIPERLAN/2 system is used in two configurations: for office and for household. The office system represents the access network containing several access points, which are connected with the reference (base) networks. Each access point supports the definite quantity of mobile terminals coupled with it. Basing service zones of adjacent networks can be partially overloaded. At that, terminal movement between service zones of different access points is possible. Construction of the HIPERLAN/2 system in the household configuration usually realizes in the network of "ad hoc" type. Such networks can include several sub-networks, each of which operates on its own frequency under control from a controller (dynamically chosen terminal).

The allocated frequency range containing 17 channels is divided into 2 sub-ranges with the following carrier frequencies: 5.18...5.2 GHz and 5.50...5.70 GHz; the offset between them is 20 MHz. Application of the high-effective OFDM technology allows formation of the channel for the high-rate data transmission in the range of 5 GHz, each of it includes 52 sub-carriers with offset of 312.5 kHz. The spectrum of such OFDM channel is uniformly distributed around the carrier frequency and has width not less than 16.25 MHz. The transmission of digital data is organized in the form of OFDM packets (symbols). Each packet begins from a preamble containing the definite quantity of short adjusting OFDM symbols, after which several long adjusting symbols follow. During information transmission, four sub-carriers are used as pilot: via them the known informational symbols are transmitted. These symbols located in the beginning of packet are used for the AGC system, for the primary correction of the carrier, and for synchronization. The pilot symbols located inside the packet are used for correction of the phase shift and the final correction of carrier.

2.3.1.2 Networks of the IEEE 802.11 (WiFi) Standard

The local networks with the Ethernet technology (the IEEE 802.3 standard, the cable local computing network), wireless local networks (WLAN) with the family IEEE 802.11 [10] standard, which start their development, have got intensive development with unification under the single trademark WiFi (Wireless Fidelity). The main difference of such networks, for instance, with respect to the local networks of the WiMAX standard consists in utilization of the non-licensed frequency range. In the IEEE 802.11b standard, the well-known methods of spectrum spreading in the cellular communication systems of the third generation are used, namely, frequency hopping (FHSS) and the direct sequence (DSSS) at the exchange rates up to 11 Mbit/s. The data transmission at rates of 5.5 and 11 Mbit/c occurs with application of modulation by means of complementary coding sequences **CCK** (Complementary Code Keying).

The standard IEEE 802.11a describing the operation in the range of 5 GHz, adopted simultaneously with the IEEE 802.11b standard, assumed the application of principally another mechanism of modulation/multiplexing, namely, the frequency division of orthogonal carriers (OFDM). This method was already widely applied in networks of digital TV broadcasting DVB. The main development directions of the IEEE 802.11 standard are development and implementation of new technologies, which increase the data exchange rates, an assimilation of new frequency ranges, and creation of the mobile version (IEEE 802.11p) under condition of assignment of guaranteed communication quality **QoS** (Quality of Service).

The high carrying capability of the network is achieved by utilization of the MIMO technology with eight spaced antennas on the receiving and transmitting sides, the frequency band spreading up to 80 MHz, instead of the standard band in 20 MHz, and utilization of the OFDMA technology of the multi-station access to the channel with division for orthogonal frequencies, as in the IEEE 802.16 protocol.

The network equipment of the IEEE 802.11 protocol obeys two lower levels of the interaction model of the open systems (OSI): physical and channel. The physical level defines the approach to operation with the transmission medium, the transmission rate, and the methods of modulation. At the channel level, **MAC** (Media Access Control) principles are specified, on the base of which devices share the general channel, approaches for connection to the access points and their identification, and mechanisms of data protection. Since the IEEE 802.11 standard was developed as the "wireless Ethernet," in it methods of packet formation are remained, which performs their coupling with each other.

IEEE 802.11 standards as well as the HIPERLAN/1 standard for WLAN satisfy to the OSI model on the physical level. The physical layer of the IEEE 802.11 standard assumes the possibility of organizing radio channels in the ISM (Industrial, Scientific, and Medical) range of 2.4...2.4835 GHz and in the infrared range (5.150 ... 5.350 and 5.785 ... 5.825 GHz). Three different methods of transmission are used – FH-SS, DS-SS, and OFDM – with the basing rate of data transmission in all interfaces of 1 Mbit/s. The data transmission with the rate of 2 Mbit/s is permitted.

The WLAN can be configured as “ad hoc” network and the network of structured type. The most important block of the structured IEEE802.11 network is the basic service area **BSS** (**B**asic **S**ervice **S**et). This is the group of stations that can be connected to each other, while remaining under the control of one of them (Fig. 2.8).

In networks of IEEE 802.11 standard, the subscriber terminals (ST) are connected with each other (Fig. 2.8) through access points (AP), but they cannot perform the information exchange with each other.

The basing service zones are connected through the access points with the distribution system, which is connected through the portal (interaction block) with external IEEE 802.x networks. The distribution system can be constructed on the base of any wire or wireless network and provides the matched BSS interconnection and simultaneously increases the service zone.

Networks of the IEEE 802.11s standard can form the mesh networks by including in their structure of specific devices: points of the mesh network (MP), which can interact with each other and support the mesh services (Fig. 2.9).

Fig. 2.8 The basic service set

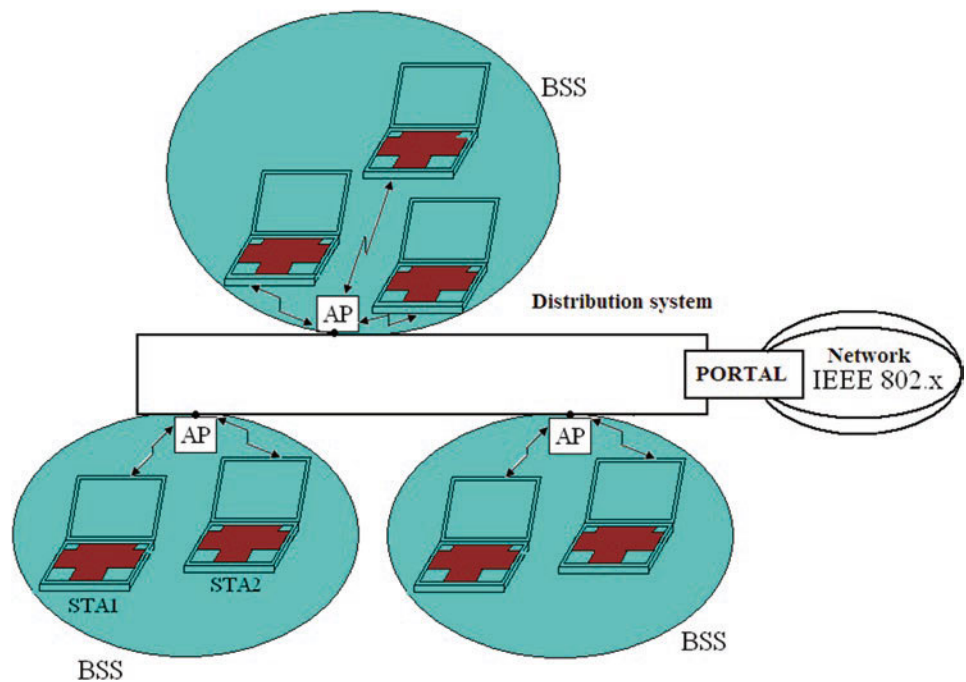
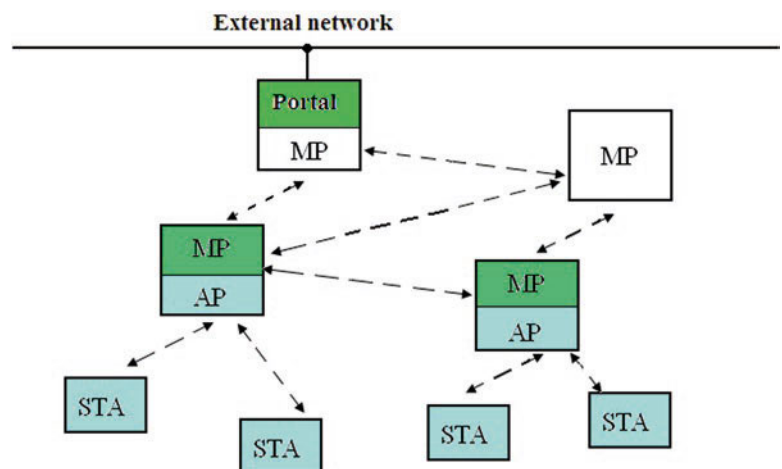


Fig. 2.9 Networks of the IEEE 802.11s standard



The single device can combine several functions: the portal of the mesh network MP is simultaneously the node of the mesh network and can provide the connection with external networks. From the point of view of high-level protocols, the mesh network is equivalent to the widely broadcasting Ethernet network.

The physical interface **FH-SS** (Frequency Hopping Spread Spectrum) provides the technology of signal spectrum spreading by variation of the signal carrier frequency, which is used for transmission via the wideband radio channel in the range of 2.4 GHz. During signal transmission, the transmitter jumps over one operating frequency to another. The minimal retuning speed in the USA is 2.5 jumps per second. The mobile terminal recognizes the frequency retuning speed during the connection with the specific access point. The scheme of frequency retuning is chosen to make minimal of the frequency number in different basing service zones. In Europe and the USA, in the ISM range, 79 frequencies are allocated with the step of 1 MHz. For the basing rate of 1 Mbit/s, the double-level GFSK modulation (Gaussian frequency shift keying with minimal frequency shift) is used, while for 2 Mbit/s the four-level GFSK modulation is used.

When using the technology of *signal spectrum spreading by the direct sequence* (**DS-SS**), each information bit is sent on several (usually, 11 with the Barker code) frequencies simultaneously. At that, for each following bit, the new frequency set is used, which is chosen according to some digital key, which is necessary to know on the receiving side for message decoding. The expanded sequence with the rate of 11 Mbit/s modulates the carrier according to **DBPSK** (differential double-position phase shift keying) or **DQPSK** (differential quadrature phase shift keying) schemes forming the spectrum with width of approximately 22 MHz.

Necessity of increase of data transmission rates in WLAN resulted in the appearance of new standards HIPERLAN/2 and IEEE 802.11a, which operate in the 5 GHz range. Physical levels of both networks are very similar and provide the following operation modes:

- Centralized, when access points are connected with the reference network; communication between the mobile terminal and other devices is always arranged through the access point.
- The direct communication, when terminals directly exchange data with each other under the control of the central controller.

The structural diagram of the IEEE 802.11a system using the OFDM technology is presented in Fig. 2.10, and the procedure of signal processing is closed to that described for the HIPERLAN/2 system.

The input data flow (bit) is exposed to randomization (multiplication by the pseudo-random sequence (**PRS**)) with the repetition cycle of 127. PRS is formed by the generator with the specific polynomial $G(x) = x^7 + x^4 + 1$. After scrambling (randomization), the data flow passes to the convolution encoder with the direct error correction (**FEC**). Depending on the chosen data transmission rate, the coding rate r can be: 1/2, 2/3, or 3/4, i.e., each input bit can create on the encoder output,

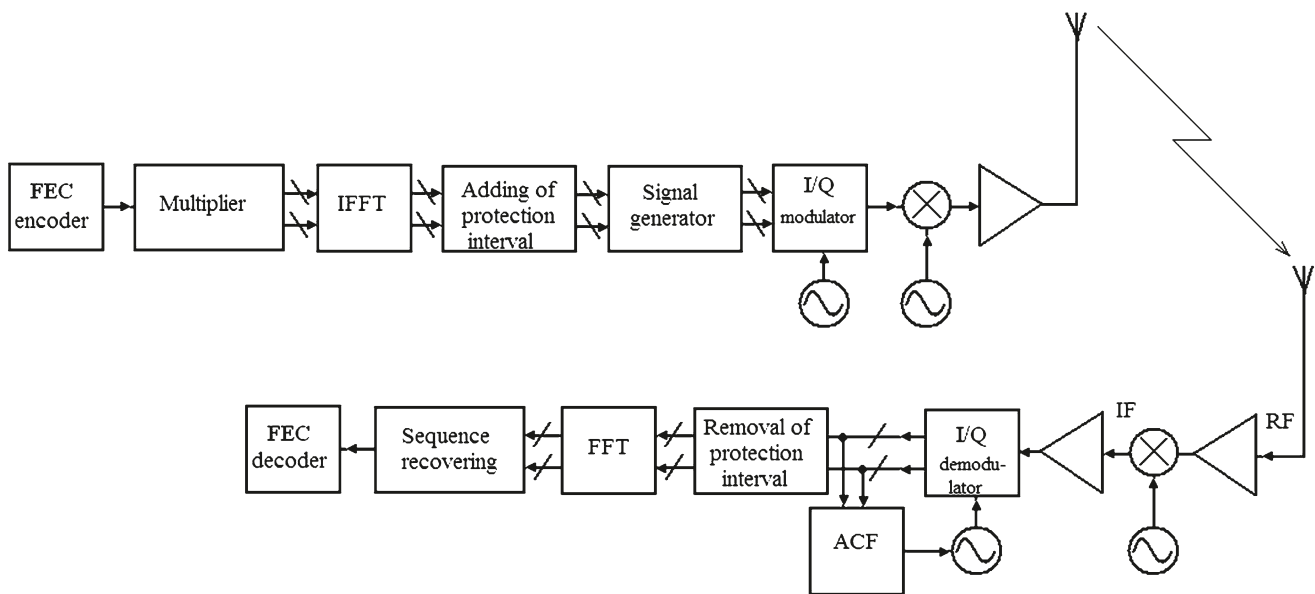


Fig. 2.10 The structural diagram of the IEEE 802.11a system

for example, two bits: $r = 1/2$. To obtain the other coding rate values, we can apply the “prick out” procedure, when the output sequence formed the separate values from two sequences on both outputs of the encoder.

In the transmitter’s multiplier, the coded flow of bits changes its order, which is defined by an algorithm of OFDM symbol formation. Depending on the bits’ number, the whole coded sequence is divided into blocks, in each of which the bit number is equaled to the bit number N in the OFDM symbol. Bits of each block are numerated and are exposed by interleaving in the matrix $16 \times N/16$. After that, the adjacent bit rearrangement is performed to be alternately in higher-order and lower-order bits. Necessity of such rearrangement is caused by that the transmission reliability of symbols of lower-order is lower compared to the higher-order transmission and their separation decreases the error grouping probability.

The OFDM technology applies the simultaneous information transmission on a large quantity of sub-carriers, which constitute a channel. At the channel width of 20 MHz, 52 sub-carriers are used in the channel (their factual separation in the channel is calculated using 64 sub-carriers $\Delta f = 20 \cdot 10^6 / 64 = 312.5$ kHz for simplification of the sub-carrier frequency grid formation with application of the inverse fast Fourier transform). Sub-carriers are modulated by quadrature phase and amplitude modulation BPSK, QPSK, 16-QAM, and 64-QAM with the position numbers 2, 4, 16, and 64, relatively.

The OFDM symbol is formed by a variety of all sub-carriers acting simultaneously on the interval $T = 1/\Delta f = 3.2 \mu s$. Its information capacity is defined by the modulation type and the number of used sub-carrier. So, in the IEEE 802.11a standard, from 52 sub-carriers, only 48 are used for data transmission, and the others are pilot. From here, the information capacity of the OFDM symbol is from 96 to 288 bits ($48 \cdot N_b$, where N_b is the bit number on the single sub-carrier).

Application of the OFDM technology essentially decreases the data transmission rate on the single sub-carrier, which allows to add into each OFDM symbol of the protection interval, which is a quarter of the information interval duration: $T_{prot} = T/4 = 0.8 \mu s$. Then, the overall duration of the OFDM symbol is $T_{sym} = T + T_{prot} = 4 \mu s$. The initial part of the symbol of the modulating sequence T , which is subjected to greatest distortions, is replaced by a copy of its final part and becomes the protective interval T_{prot} . This ensures mutual orthogonality of modulated subcarriers along the length of the symbol. Therefore, its location at beginning of the OFDM symbol protects the received line-of-sight signal from the character received with a delay, since the beginning of the delayed (reflected) symbol falls within the protection interval, preserving the mutual orthogonality of direct signal and its copies.

The digital signal consisting of two orthogonal components is transferred in I/Q modulator onto the appropriate sub-carrier; after that, the spectrum of the resulting grouped signal is transferred in the operation range by the frequency converter and is amplified in the transmitter power amplifier.

In the receiver section, the operations, which are inversed to appropriate operations in the transmitter section, are performed. In addition, the low-noise amplifier (LNA) is included at the input of the front-end for achievement of required receiver sensitivity and for suppression of spurious conversion channels.

In spite of high transmission rates in the cellular networks of 4G (LTE) and in networks of the IEEE 802.11n standard (up to 300 Mbit/s), investigations are performed on development of ultra-fast WLAN (up to 500 Mbit/s) in the frequency range up to 6 GHz.

In Russia, investigations are performed on creation of ultra-fast mesh networks (development of software and technologies) with expected information transmission rate up to 1 Gbit/s in the frequency range 60 GHz.

2.3.1.3 Systems of the DECT Standard and the Cordless Telephony

DECT standard. *Systems* of the DECT (Digital Enhanced Cordless Telecommunications) standard [68] developed for speech signal transmission inside premises find a wide application in regions with very dense communication traffic (railroad stations, airports, etc.) as the *component part of cellular communication systems or for organization of wireless access to network of data transmission* including to the local computing networks. Such application of DECT standard systems became possible owing to the information flow high rate and the wideband properties of radio channels.

Micro-cell organization of the DECT system allows elimination of the frequency planning and the relay transfer of the mobile terminal, which decreases their cost and simplifies the channel control. Frequency planning in such networks turn out to be inefficient due to difficulties in predicting the propagation of radio waves and large spread in the level of interference acting in the channel. In this connection, the procedure of automatic adaptive channel distribution is used. Significant growth of the switching frequency of basing stations in the micro-cells requires application of fast-acting switching algorithms for switching centers; therefore, the control is transferred to the mobile object. The mobile terminal performs the permanent measurement of the input signal level and makes a decision on switching, while functions of fulfilled decision registration are entrusted with the base station and the switching center, which excludes its overloading.

In the DECT technology, the direct radio access method (**RAM**) is used at operation on several sub-carriers with time channel division and with time duplexing (MC/TDMA/TDD). The applying algorithm of the dynamic choice and channel

selection allows development of pico-cell elements in the strongly overloaded (and even in hostile) medium and effective application of the same frequency range for operations of several operators.

Basing principles laid the foundation at development of the DECT standard which allows provision of:

- Stable functioning of the network components even in aggressive radio media.
- The dynamic choice of necessary frequency band.
- Network self-organization, which excludes a necessity of frequency planning.
- The subscriber mobility within the network (the speed is less than 30 km/hour).
- Flexible and reliable identification and addressing.
- High quality of speech transmission comparable with the quality provided in wired systems.
- High data transmission rates (to 2 Mbit/s).
- Transmission of speech, data, and multimedia in different communication sectors.

Possibilities of systems of the direct radio access standard allow arrangement of wideband communication networks with mobile objects of different purposes for various signal types.

Micro-cellular Systems of Joint Access Application of DECT technology allows the development of the system of joint utilization, in particular, micro-cellular systems, which have got at present a wide distribution and are called **CTM (Cordless Terminal Mobility)** deploying in regions of significant concentration of subscribers (airports, trading centers, hotels, etc.). The DECT equipment allows effective utilization of the same phone in the domestic system while at transfer into CTM to provide the subscriber with communication without changing its number in the stationary network.

Multimode Wireless and Cellular Systems Cellular systems of the GSM 1800 standard and wireless systems DECT operating in the same frequency range (Table 1.2), as well as the ramified network of the GSM 900 standard, mutually supplemented each other. One of them provides the large covering zone and has the ramified infrastructure, while another is able to service the intensive subscriber traffic at small covering zone. Development of the subscriber terminal, which remains effective in any network including the inter-network roaming, is rather promising.

The Direct Communication Mode Systems of the DECT standard were developed for radial network system, when the communication arrangement between two portable terminals (**PT**) was provided through the central base station. However, portable terminals located in the same service zone have a possibility to implement the mode of the *direct duplex* communication, using them as the portative radio stations (walkie-talkie mode). In this case, one PT plays a role of the base station, and during the exit from the service zone, it informs the subscriber about it, transferring the control to the central base station.

At present, development of local computing networks (WLAN) is performed in the direction of wireless system utilization in configuration “point-point” and “point-many points” on the base of the dynamic spectrum spreading OFDM technology. In addition to the IEEE 802.11 standard, the working team developing this standard published the IEEE 802.11a standard, in which the approach is outlined to OFDM utilization in the frequency range of 5.8 GHz. The fundamental principle is based on division of the high-rate binary signal to transmit it on several sub-carriers with the lesser rate. Each low-rate digital flow is used for modulation of the separate sub-carrier of the single channel in the range of 5.8 GHz. 48 sub-carriers are formed for informational flow and 4 for pilot signals (totally 52 sub-carriers). Before transmission, data are encoded with application of convolution codes with the coefficient $R = \frac{1}{2}$ and symbols’ interleaving to achieve the required rate of the information flow. Then, the complicate number, which corresponds to the modulation type and to belongings to 48 information sub-carriers or 4 sub-carriers of pilot signals, is conferred to each message. Sub-carriers are unified using the inverse Fourier transform and then are transmitted. In the receiver, the sub-carrier is inversely converted into the low-rate digital flow on many sub-carriers with the help of FFT. After that, sub-carriers of low-rate flows are again grouped into a block of the high-rate digital sequence.

Domestic Systems Appearance of cordless telephony systems (CT) (**Cordless Telephony**) is caused by the subscriber needs to have a larger freedom for displacements in a space on small distances and with the low speed at operation in PSTN. The main idea at implementation of the cordless telephone (radio stretcher) consists in the replacement of the cord between the telephone apparatus and the receiver without degradation of communication quality and at saving of the telephone apparatus cost, which should be comparable with the cost of usual wired apparatus.

The role of the base station is fulfilled by the telephone apparatus connected in PSTN, while the role of the mobile station is fulfilled usually by the only mobile telephone receiver providing the long operation without recharging of the built-in rechargeable battery.

The feature of such systems consists in the large number of base stations per a unit of area and in the low number of subscribers per a unit of spectrum, as well as the low communication distance.

Business Systems Application of digital technologies allowed widening of the covering zone of base stations and exclusion of the problem of subscriber call locking by another base station. With the aim of wireless communication unification in the European Union, the unified digital standard optimized for using in premises was introduced. In this case, base stations are connected to controllers of office ATS (**OATS**). At variation of the subscriber position, the near-located base stations (at degradation of information exchange conditions) can provide the relay subscriber transfer to each other on the controller command. Potentially, the single transceiver of the base station can simultaneously support up to 12 duplex conversations on 12 telephone lines.

To organize the covering by the mobile communication in large offices or industrial enterprises, the multi-cellular systems of cordless DECT systems (large business telephone systems) are used. Utilization of the adaptive algorithm of channel choice in DECT provides effectiveness and reliability of such systems, which does not depend on the simultaneous application of several systems located inside or outside the premise.

Systems of the Subscriber Radio Access Such systems are called **WLL** (Wireless Local Loop) or **RLL** (Radio Local Loop) and provide the wireless radio communication in some territory-limited area. They are used for connection of the subscriber or the group of subscribers to PSTN.

Communication services are given to the subscriber through the wireless terminal adapter **CTA** (Cordless Terminal Adapter), which is connected to the standard telephone apparatus. Factually, CTA provides the stationary radio channel in the DECT network, which is connected to PSTN. The base station of the DECT network should be at the distance less than 5 km to provide the subscriber with the stable radio channel. Implementation of the stationary radio channel through the wireless relay station **WRS** (Wireless Relay Station) is provided by the longer length, and a possibility of a larger list of services is given, as well as elimination of breaks in territory covering.

2.4 Municipal Networks of Wireless Access (WMAN)

2.4.1 The Network of the Wideband Access (WiMAX)

Escalating needs in services of the wideband data and multimedia transmission to the mobile terminals at small distances between subscribers led to active development of municipal wireless networks (WMAN). Revolutionary changes in technology of integrated circuit manufacture and development of new algorithms of large data flows' processing led in 2004 to the new IEEE 802.16 standard for fixed and then for mobile communication systems. The IEEE 802.16-2004 WiMAX [84] standard describes principles of the network structure for regional scale in ranges up to 66 GHz (methods of the channel access, radio interfaces, modulation methods, systems for flow control, structures of transmitted data, interfacing mechanisms for protocols of data transmission of higher levels with protocols of the physical level). In Europe, the equivalent of IEEE 802.16 standard is adopted called **HIPERMAN** (**HI**gh **PER**formance **Met**ropolitan **A**rea **N**etwork), which anticipates an operation in the range 2...11 GHz in the WirelessMAN-OFDM mode.

By this time, for fixed networks of the WiMAX standard, the utilization of IP and the technology of the *asynchronous transfer mode* (ATM) of data of the upper level was anticipated (the technology of packet switching of fixed length) of different data types (a speech, a video, multimedia, the local area computing network traffic) for significant distance. The operation radius of such networks is up to 10 km, and connection between terminals is within the limits of direct visibility. Certified WiMAX networks of fixed access use the frequency band of 3.4...3.6 GHz with the time and frequency duplex applying the OFDM multiplexing at a number of sub-carriers in the channel equaled to 256. At radio signal generation, the quadrature phase shift keying (QPSK) is used and the 16-position quadrature amplitude shift keying (16-QAM) of the carrier at the bandwidth of the radio channel of 3.5...10 MHz. In the mobile WiMAX, the time duplexing (TDD) is only used with operating frequency bands of 2.3...2.4; 2.496...2.960; and 3.4...3.6 GHz with the sub-carrier number in the channel of 512 and 1024 and the bandwidth of the radio channel of 5...10 MHz. In the engineering channel, the QPSK modulation is used, while in the information channel 16-QAM or 64-QAM with OFDMA multiplexing is used.

The problem of data flow transmission between the subscriber station (terminal) and the base station is solved on the physical level of the IEEE 802.16 standard, while the structure formation of these data as well as the control of system operation is realized on the MAC level (the access control to the data transmission medium). MAC is the lower sub-level of the channel level in the seven-level ISO/OSI model of interaction of open systems, which specifies access methods to the medium, the frame format, and addressing; supports the multiple access to the communication channel; fulfills the reception and transmission of information and controlling frames; and detects the transmission errors. The problem of the transport medium formation for all applications of the MAC level protocol independently on the physical channel features was stated for developers of the IEEE 802.16 standard. In this, there is the main difference of the IEEE 802.16 standard developed for implementation in the municipal distributed networks of the wireless wideband access technology of the operator class from the non-licensed IEEE 802.11 standard.

On the physical level of the IEEE 802.16 standard, we can distinguish three methods of data transmission: the method of single carrier (SC) modulation; the technology **OFDM** (**O**rtogonal **F**requency **D**ivision **M**ultiple) by means of orthogonal carriers; and the method of multiple access (multiplexing) by means of orthogonal carriers **OFDMA** (**O**rtogonal **F**requency **D**ivision **M**ultiple **A**ccess).

The WirelessMAN-SC method assumes the *single carrier modulation* in each frequency channel. Three modulation types are used: 4-QPSK, 16-QAM, and 64-QAM (optionally). Depending on the channel width and the modulation method, the different rates of the physical data flow are realized (Table 2.2).

For WirelessMAN-SC networks operating in the range 10–66 GHz, the channel width is 20–25 MHz, at the transmission rate up to 120 Mbit/s.

The *OFDM mode* is the technology of data flow formation in the single frequency channel (with width not less than 1 MHz) with central frequency f_c . For data modulation by means of orthogonal carriers in each sub-channel (totally 16), we select N sub-carriers with the distance between them $\Delta f = 1/T$, where T is the duration of data transmission in the single OFDM symbol. At fragmentation of the single channel with variable parameters into N parallel Gaussian channels with different SNR, each sub-carrier is modulated independently by 4-QPSK or 16-QAM. The channel number is defined by the optimality condition of the direct and inverse Fourier transform ($N_{FFT} = 2^m$), which is $N_{FFT} = 256$, from which $N = 200$ are informative (others form the protective frequency interval).

The *channel resource with OFDM* can be given not only in the time domain but also in the channel (a group of channels), which was the base of mesh network creation. Such networks organized only in the IEEE 802.16 standard in the OFDM mode are characterized by the fact that its nodes provide the end-to-end traffic transmission (Fig. 2.11), which enables the direct interaction between subscriber stations.

Such an arrangement of the decentralized wideband network, when the traffic can be transmitted on the chain from several stations, allows the solution of the information transmission problem at direct vision absence between stations. Mechanisms of the decentralized distributed network control are anticipated by this standard, although the treelike architecture with the allocated base station and dominating flows BS-SS and their combinations is the base architecture. The typical difference of mesh network structures is the interaction of each node, which has the unique 48-bit MZC address, with adjacent nodes on the MAC level. The network control is performed on the base of lists of data about its own neighbors stored in each node, which organizes the direct connections with stations organized in the adjacent environment. The whole information exchange between nodes occurs by means of frames. Stations transmit information in the allocated time interval in the absence of their mutual subordination (ascending/descending channel).

2.4.2 Standards IEEE 802.16e; j; m

Active operation is performed for creation of the IEEE 802.16m standard, which has the more perfect physical level at compatibility keeping with the IEEE 802.16e standard and satisfies to requirements of the fourth generation (4G) standards.

Table 2.2 Different rates of the physical data flow

Channel width, MHz	Symbol rate, Mbod	Rate of physical data flow, Mbit/s		
		QPSK	16-QAM	64-QAM
20	16	32	64	96
25	20	40	80	120
28	22.4	44.8	89.6	134.4

In the IEEE 802.16e standard adopted in 2005, some points of the IEEE 802.16-2004 standard (the standard for fixed networks) are removed, and also additions in the physical and MAC levels are introduced for the joint operation with other standards in licensed frequency ranges [87]. This allows the transfer of IEEE 802.16 standard networks from the group of local computing networks into the family of mobile communication networks granting to them the mobile applications. In the new standard, all subscribers are divided into two categories: *nomad (migrant)* and *mobile*. Subscribers may move within the limits of network action, but in the moments of communication sessions, they are localized (are in the zone of the single segment of the base station). In these cases, the information exchange procedure does not differ from the procedure in any fixed network.

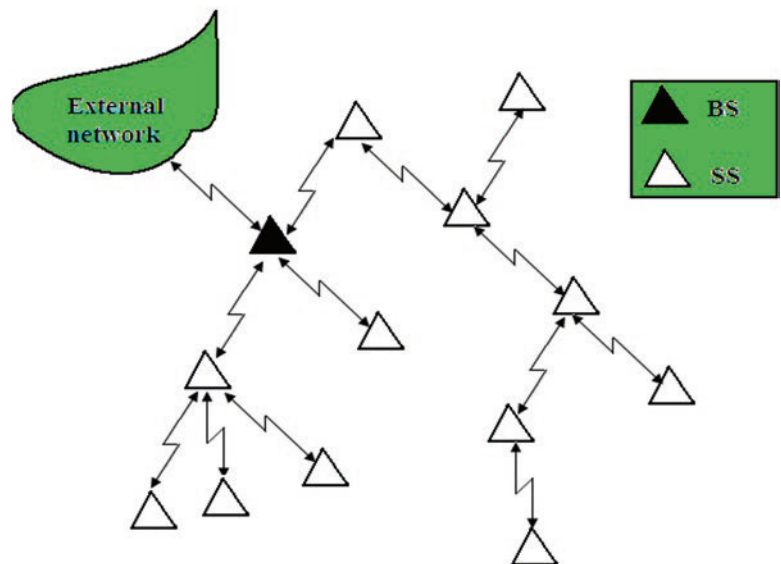
Mobile subscribers have an access to the network being in the motion (e.g., in the car). The order of the full subscriber mobility ensuring is defined by the standard in the form of the subscriber transfer procedure from one BS to another or between segments of the single BS without communication quality reduction: the relay transfer (handover). The transfer into the category of mobile networks intensifies the problem of effective utilization of allocated frequency resource (control of the frequency resource, urgent frequency redistribution, reuse). A possibility of proportional distribution of the frequency band and the technology of multi-channel antenna systems (MIMO) is anticipated by the standard. As in all mobile communication networks, issues of power consumption of the mobile devices become relevant. Simultaneously, the network architecture becomes more complicate: if the fixed SS is connected to the single BS station (see Fig. 2.11), then for the MS we consider as “service” station those BS, from which the last registration (at entering in the network at handover) was fulfilled. BS is called “adjacent,” when for MS there is a possibility to receive the descending flow. There are variations in the content of concepts (physical channel) and in sub-level of MAC access control to the channel on the network level.

The concepts of “mobility”, “mobility station”, which have arisen in the middle of the last century as result of emergency of land and maritime mobile communication system. The widespread of digital mobile systems and introduction of cellular networks and digital signal processing methods have led to the appearance of new indicators [84] such data delay, signal jitter, quality of service (QoS) criteria, etc. The “global service class” – the set of eight parameters always remaining constant – is the instrument of quality control. In this standard, the different types of request planning on resources’ granting are anticipated: dimensions of data blocks, the minimal assured rate of data blocks, a delay and an interval between transmission sessions, and tolerance to the jitter. There are different types of their delivery: without requirement (with agreed periodicity), delivery in the real time with variable speed, and delivery outside the real time with variable speed, as far as possible.

Operation modes of MS. With a purpose of energy saving, which is consumed by the MS, two energy-saving modes are used: the doze mode and the idle mode.

In the *doze mode*, MS remains inaccessible for reception and transmission remaining in the subscriber list of any BS and not occupying the channel resources. Transfer into the doze mode occurs according to request of MS and the command from BS in the form of the special message. The doze mode represents the sequence of sleeping and listening intervals, which are defined by the chosen class of energy-saving. In the inaccessible intervals, MS cannot connect with BS, and if BS receives data addressed to MS, then BS may either buffer them or to dispose them. MS exit from the doze mode is possible either according to the special command from BS or under action of definite events (e.g., the signal at the detector output exceeds the threshold level) or on the schedule.

Fig. 2.11 End-to-end traffic transmission (BS, basing station; SS, subscriber station)



Appearance in MS of the optimal *idle mode* provides the periodic availability of MS for the widely broadcasting descending traffic without registration in any BS. The idle mode in MS is factually the paging mode, when the station periodically listens to the widely broadcasting channel, which potentially includes the transmitted information. The paging interval is the period of listening, during which MS is waiting the paging message. In remaining time the station can be in switch-off state or fulfill operations, which do not require registration in the network. The transfer into the idle mode can be on MS initiative, which directs a request from BS about registration finishing and the transfer into the idle mode. After confirmation exchange, MS goes to the idle mode, but at that, BS, in which MS was registered, or BS performing the controller functions can save during some time its network adjustments for easy transfer into the active mode.

During motion of the MS, the signal reception conditions are varied, and at the value of the quality factor QoS below admissible level, MS chooses the new BS. The *handover procedure* is forestalled by scanning and associating operations. For BS choice providing the better conditions of signal reception, the service BS, according to the request of MS, allocates to it special time intervals, in which BS transmits information about adjacent BS including their profiles, operation modes, etc. MS independently or with the help of service BS, during these time intervals (scanning intervals), estimates signal parameters from the adjacent BS. The decision about MS connection to the new service BS, depending on the associating variant, is made by MS independently or together with the service BS on the base of estimation of necessary (for MS) channel resource and the QoS parameter. After connection with chosen BS, the initial registration procedure is fulfilled in the network.

The paging mode in the IEEE 802.16e standard is provided by modes: grouped (multicast) and widely broadcasting transmission. Taking into account the small difference between modes, there is a service in the standard: the grouped and widely broadcasting transmission **MBS (Multicast and Broadcast Service)**, which is supported by each MS.

Main variations of the IEEE 802.16 standard in discussing version are concerned to the physical level, and they allow the larger flexibility of system operation with different channel bandwidths. Main variations in the standard are related to the OFDM *technology*, the channel multiplexing technology, and the OFDMA *method of access to the grouped channel of channel multiplexing*. In the OFDM mode, the principle of sub-channel formation is specified, when 256 sub-carriers form 12...16 sub-channels. Channels are multiplexed on the base of time channel division (TDMA), and duplexing can be temporal (TDD) or frequency (FDD).

In the OFDMA mode of the basing standard IEEE 802.16-2004, it was anticipated the formation of 2048 sub-carriers, and in the considered version, the mode of the scaling OFDMA (S-OFDMA) appeared. In this variant, 1024, 512, or 128 sub-carriers are used, which allows variation of the frequency band occupied by a channel – the distance between nominal sub-carriers. It is known that the distance decrease between sub-carriers leads to channel bandwidth reduction and, relatively, to decrease of the permissible modulation speed. In this case, the OFDM technology advantages as the approach to fight with intersymbol interference are lost, especially at the large speed of subscriber motion, when an influence of the Doppler shift grows. This leads to the loss of the main advantage of such a modulation, which permits to operate under the condition of direct vision absence (large reflections).

Varying of sub-carrier number proportionally to the bandwidth, we can provide the invariance with respect to the bandwidth of such parameters as the OFDM symbol duration and the distance between sub-carriers. For instance, at transfer from operation in the 20 MHz band to the 5 MHz band, we can transfer from operation on 2048 carriers to 512 carriers with the capacity reduction by 4 times at constant other parameters. However, the S-OFDMA introduction requires that all mobile terminals at primary registration and the environment scanning would be able to determine the bandwidth of available BS and the number of used sub-carriers.

One of the methods to increase the QoS parameter is application of the automatic request for repeat transmission (ARQ) for utilization of the OFDMA method in the IEEE 802.16e standard; it has got new development in the form of the *hybrid automatic repeat request* (HARQ) mechanism. At error appearance in reception, the error confirmation is sent through the specially allocated channel (method of checkback transmission).

IEEE 802.16-2004j is the version of the base standard, which directs on development of mesh networks: increase of their performance at multi-stage transmission. Relays applied for this, as the subscriber terminals, may be mobile, while the relay can operate in the broadcasting mode. The relay station (RS) can operate in transparent (transmitted data only) and in non-transparent (transmitted data, preambles, and controlling messages) modes. In nontransparent mode, this allows the cell capacity growth of more than 40%.

The new **IEEE 802.16m** standard, which has better wireless interface, should have the full compatibility with other IMT Advanced technologies and IMT-2000 technology intended for developing system of wireless communication of fourth generation. At its development, the following conditions become basing: operation on frequencies less than 6 GHz in wide-band mobile and fixed networks and a possibility to support the modes of frequency and time duplexing. The equipment of this standard should provide the transmission rate more than 150 Mbit/s in the descending channel with width of 20 MHz with delay of IP-level data packet transmission no more than 10 ms. The peak normalized capacity is not less than 8.0 and 2.8 bit/s/Hz.

The system of the IEEE 802.16m standard provides the additional energy benefit by 3 dB compared to IEEE 802.16e. The covering zone is increased. With radius up to 5 m, the system must operate with specified characteristics. When the radius increases up to 30 km, the gradual characteristic degradation will occur, but at that, the system keeps the effectiveness for removal from the base station to 100 km distance (the sensitivity is limited by the level of thermal noises at the receiver input). The speed of the mobile station can be up to 300 km/hour.

The most important feature of the IEEE 802.16m standard using the OFDMA mode is variation of the frame structure (Table 2.3).

As we see from parameters in Table 2.2, the IEEE 802.16m standard is factually the standard of fourth generation and competitive to LTE Advanced.

2.4.3 The Interface of the WiMAX Network

The WiMAX network represents the variety of wireless and basing (reference) segments. The wireless segment based on the packet switching technology is described in the IEEE 802.16 standard, while the basing segment is defined by specifications of the WiMAX forum including the IP and Ethernet standards and specifies the base station interaction parameters with each other and with local and global networks (WiFi, iDSL, Internet), etc.

The WiMAX network should be easily scaled (according to subscriber number density, the categories of the network are district, municipal, suburb) and flexible (to operate in several frequency ranges, to use the various network configurations, to ensure services for fixed, nomad, mobile subscribers).

The network WiMAX model (base model – BM) is described by a set of logical functional modules and standard interfaces (the module interface points). At practical realization, each module may include several functional elements, and each function is distributed between different devices.

The base model contains three main elements: a variety of subscriber stations (**MS** – mobile stations, terminals), a variety of networks of access (**ASN** – **A**ccess **S**ervice **N**etwork), and a variety of connection networks (**CSN** – **C**ircuit **S**witched **N**etwork). As we see from Fig. 2.12, for the base model, the interfacing of functional modules occurs through the basing points (R1...R5).

The ASN belongs to the provider of the network access (NAP) – the institution providing an access to the radio network of service providers of WiMAX (NSP). The service provider WiMAX (institution) provides IP connections and services to WiMAX final subscribers (conclude agreements with Internet providers, operators of other networks, agreements on roaming, etc.). With regard to the subscriber, service providers can be domestic and guest each having its own CSN.

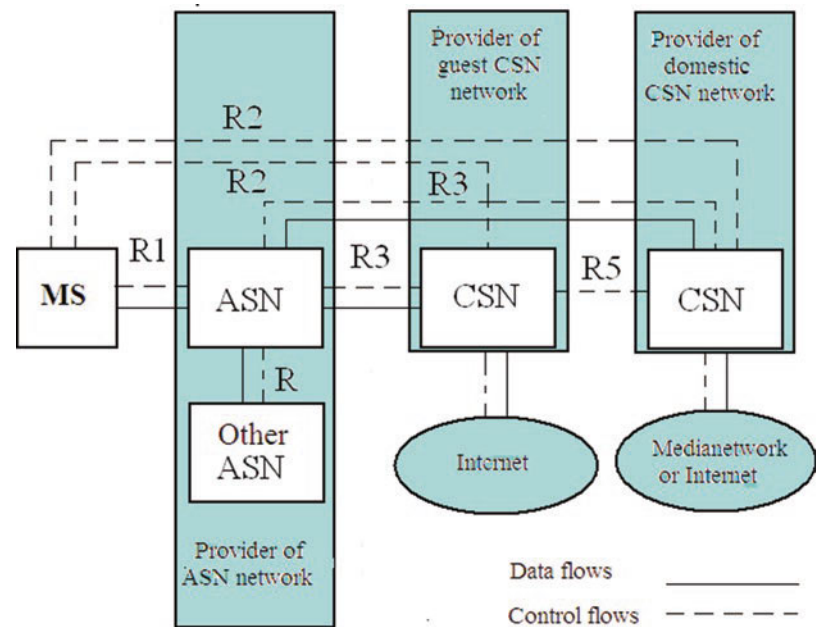
The access network ASN is a variety of base stations of wireless access according to the IEEE 802.16 standard and gateways for communication with the transport IP network. In minimal integration, ASN contains the single base station and the single gateway. At that, for the mobile station, the base station is the logic device supporting the set of IEEE 802.16 protocols and providing the external interfacing. The gateway is the logic device connecting the base station of the single ASN with other access networks and with the connection CSN.

The connecting network CSN is the network of the WiMAX operator, which realizes the control functions for authorization, authentication, and accounting (AAA), the subscriber connections to IP network (an access to PSTN, IP telephony, the

Table 2.3 Parameters of the frame structure of the IEEE 802.16m standard

Nominal channel bandwidth, MHz		5	10	20	7	8.75
Discreteness index (discreteness frequency/bandwidth)		28/25			8/7	
Discretization frequency, MHz		5.6	11.2	22.4	8	10
Formal sub-carrier number (for FFT)		512	1025	2048	1024	1024
Sub-carrier step (kHz)		10.938			7.8125	9.7656
OFDM symbol duration T , μs (without cyclic prefix)		91.429			128	102,4
Cyclic prefix with duration $T_{av} = 1/8 T$	Symbol duration, μs	102.86			144	115.2
	OFDM symbol number in frame	48			34	43
	Switching interval, μs	62.8			104	46.4
	Symbol duration, μs	97.143				
Cyclic prefix with duration $T_{av} = 1/8 T$	OFDM symbol number in frame (not more than)	51			Not used	
	Switching interval, μs	45/71				

Fig. 2.12 The structure of the base model



access to Internet, etc.). The connection network CSN can be connected to networks of access of different providers and give to mobile subscribers the IP addresses and provide the data transmission between access networks and connections (tunneling), billing of WiMAX subscribers and inter-operator calculations, and data tunneling between different CSN at roaming.

Basing points in the WiMAX network represents the communication channels between basing modules and are presented by the standard interfaces: R1 is the communication channel between MS and the ASN of access, and R2 is the communication channel between MS and CSN (represents the purely logic interface: authentication of ZMS, authorization, IP configuring). Functions of other basing points can be understood from the structure of the base model (Fig. 2.12).

The WiMAX network supports the TCP/IP protocol (TCP-Transmission Control Protocol), the main protocol of the transport and session levels, which allows you to identify mobile devices connecting to various nodes of the global network. The mobile IP supposes the existence of two addresses of each device: domestic (HoA) and additional (CoA). At mobile station transfer from the domestic network into external one, it can be conferred the global additional IP address, which is directed simultaneously to own domestic router, allowing the monitoring of all mobile station displacements. Another approach to provide the handover in the WiMAX network is possible: the router in the external network, which had registered the mobile station, confers it the IP address from its own pool of addresses, and during the message transmission for the mobile station, the domestic agent directs the encapsulated packets to the external router.

Considered functions of the base model in the WiMAX network were based on the **IPv4** (Internet Protocol version 4) protocol. The new version of the IP (IPv6), also used in LTE network of the fourth generation of the mobile communication networks, will be applied in networks of the wideband access. Its application is caused by the absence of embedded support of quality parameter QoS estimation and the lack of the address space restricted by the 32-bit IP address. Application of IPv6 allows utilization of 128-bit addresses, as well as application of “the alternative address,” which is referred to as the group of devices distributed over the network. In this case, the packet delivery is performed to the nearest node from the sender, and the optimization routing will exclude the obligatory packet transmission through the domestic agent (HoA).

Commercial exploitation of WiMAX networks in Moscow and Sankt-Peterburg (Russia) was begun in 2008 by Scartel Co., (trademark YOTA), and the communicator developed together with company NTS provided the simultaneous access to 3G networks WiMAX and WiFi. The “Expedience” network using the WiMAX technology provides the access in the Internet, the access in the client virtual private networks VPN (Virtual Private Network), the access to **DSL** (Digital Subscriber Line), and PSTN. The virtual private network, the sub-network of the corporate network, provides the safe entry into it of remote subscribers. Sub-networks are used for safe transmission of confident data through the Internet owing to encapsulation (tunneling) of IP packets inside the other packets, which then are routed by the wireless VPN.

At present, there is a large number of equipment for the computer periphery to connect to the WiMAX network, including special telephone receivers similar to the traditional smartphone. However, this standard has not got the wide distribution, since the majority of companies in the world have chosen the LTE standard, including “Scartel” in Russia, switching into this standard.

2.5 Systems of Cellular Communication (WWAN)

The historical course of coming into being and development of wireless (cellular) communication systems is presented in Fig. 2.13.

We describe briefly used designations in Fig. 2.13:

AMPS (Advanced Mobile Phone System) are improved mobile phone systems.

NMT (Nordic Mobile Telecommunication) the mobile communication systems applied in Scandinavian countries.

EDGE (Enhanced Data rates for GSM Evolution) are systems of high-rate data transmission for improved GSM system.

PDC (Personal Digital Cellular) is the mobile communication system used in Japan.

PHS (Personal Handyphone System) is the personal communication system similar to DECT systems used in Japan.

IS-95 standard of digital communication with code channel division.

CDMA is the system with the code division multiple access.

cdma2000 is the mobile communication system with code channel division of the IS-95 standard.

WCDMA (Wideband Code Division Multiple Access) is the wideband system with multi-station access and the code channel division.

WCDMA/UMTS (Universal Mobile Telecommunication Services) is the wideband system with code channel division/the universal system of mobile communication.

3G 1X EV-DO (Enhanced Version Data Only) is the perfected system of the 3G standard (cdma2000) intended for data transmission.

MIMO (Multi-Input and Multi-Output) is the system that uses several antennas for reception/transmission in the single frequency band. This allows increased data exchange rate by four times without front-end width growth.

LTE/LTE-A (Long-Term Evolution/Long-Term Evolution Advanced) is the system of mobile communication of new generation using the OFDMA technology of multi-station access with orthogonal frequency multiplexing.

Development of mobile communication systems began from the construction of the analogous agency-level networks with further integration ("Altai-3 M" system in Russia) with utilization of analog modulation methods. Such a system allowed communication provision in the duplex mode with the subscriber of PSTN or with another mobile subscriber through the central control center with manual or automatic connection. The active operation developed in this direction in different countries led to construction of analog cellular network of public use of the first generation (1G in Fig. 2.13). The cellular communication network developed in Scandinavian countries (NMT-450 and then NMT-900 – Nordic Mobile Telephone) was actively used in the northern regions of Russia at the end of the twentieth century and at the beginning of the twenty-first century. The cellular networks of the **AMPS** (Advanced Mobile Phone System), which were deployed in exploitation in the USA in 1983, had got the more restricted distribution.

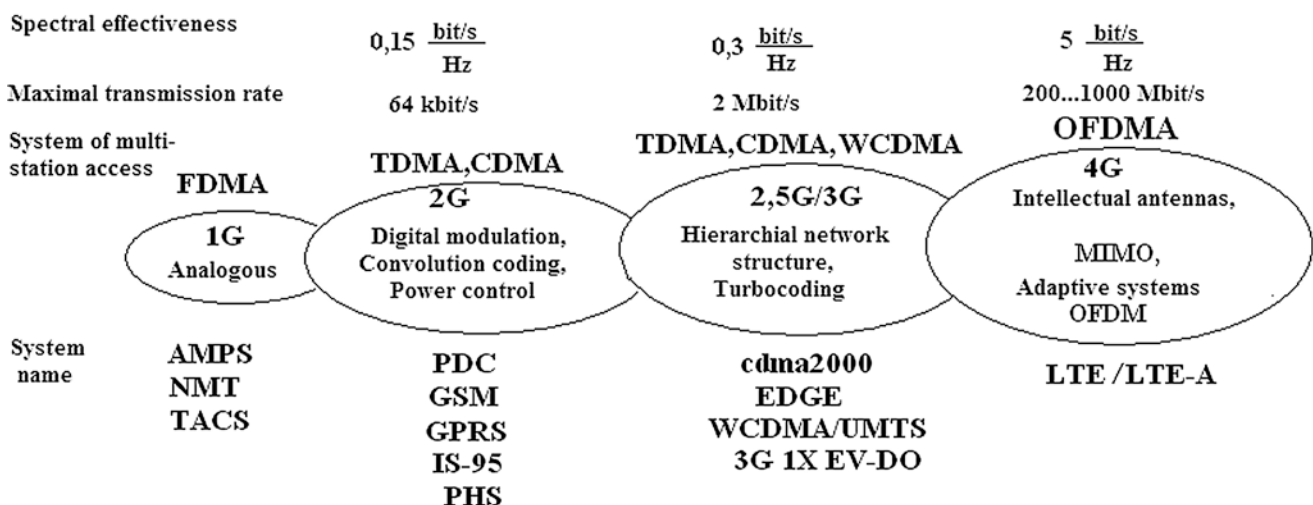


Fig. 2.13 The development of wireless (cellular) communication systems

The parallel work on development of devices of the personal radio call systems and on deployment of such networks was performed. Miniature systems consuming the small power and allowing message transmission to the particular subscriber or to a group of subscribers have got the wide distribution at the end of the twentieth in western countries, Japan, China, and Russia. The construction of such networks did not actually serious costs, since it allowed using the existing ground-based equipment of the radio broadcasting network for these purposes. At that, it is possible to create the local networks restricted by the plant or supermarket area and also the networks covering regions and even states. At present, such networks, which were functioned in the UHF range, are displaced practically in all countries due to active implementation of the trunking and cellular communication systems with the mobile objects. However, there are the development prospects of such network of special destinations in decameter range.

In spite of a variety of standards for the mobile (cellular) networks, the networks, which are adopted as standards in Russia, have the largest interest. As Russian federal standards, the following were chosen: the analog NMT and digital GSM. At construction of regional networks, we may use most often the AMPS standard and its digital variant **DAMPS** (**D**igital **A**dvanced **M**obile **P**hone **S**ystem). Implementation of wideband networks of the third generation with the code channel division is based on usage of the cdma2000 system and its improved 3G variant 1X EV-DO, as well as the wideband WCDMA system, which provides a wide spectrum of services (speech transmission, high-rate data transmission, multimedia).

The *spectral effectiveness*, which represents the index calculated as a ratio of the data transmission rate to the 1 Hz of the used frequency band (bit/s/Hz), is the main characteristic of digital systems of the mobile communication.

Historically, the first systems of cellular communication (1G) of the analog type with frequency channel division appeared in Northern Europe countries (1980, NMT, TACS) and in Northern America (1979, AMPS), which had a serious shortcoming – the incompatibility with other systems. Such communication systems had a large covering zone, when the cell radius achieved 120 km. However, the applied frequency range of 450 MHz with rather high (compared, for instance, with the frequency range of 900 MHz) level of external (industrial) interferences created the low noise immunity. The main reasons, which prevented analog system distribution, are not high enough communication quality, its dependence on the remoteness of the mobile station from the base station, the complexity of message ciphering, and limited possibilities of rendering services.

The Digital Networks of the Second Generation The new (2G) generation of cellular communication systems based on digital methods of information processing, which was developed by European countries and constructed on the principle of open system (OSI) interaction, provides the operation in the range of GSM900 (890...915 MHz), GSM1800 (1710...1785 MHz), and GSM1900 (1850...1910 MHz). In the GSM standard, the principle of channel switching is implemented. This assumes that after connection making between subscribers (necessary connection is fulfilled in the network, and necessary resources are allocated), the channel between subscribers is fixed on all levels of the OSI model, which keeps on all period of connection. To provide the simultaneous operation on reception and transmission, the frequency duplex (FDD) is used at offset of reception and transmission of 45 МГц.

The GSM architecture constructed on the principle of maximal independence between network elements, such as: the base station controller (**BSC**), the mobile switching center (**MSC**), etc., simplifies the system construction and implementation, gives a possibility to use blocks of various manufacturers, to ensure its development. The GSM standard, which has got the wide distribution in Russia, provides the speech signal transmission with the spectrum limited from top by a frequency of 4 kHz, when this speech signal is exposed by discretization with a frequency of 8 kHz and then the conversion in ADC into the digital sequence. Digital mobile systems provide not only the speech transmission but also the low-rate data transmission (up to 14.4 kbit/s).

The GPRS technology of packet transmission of speech and data with rate up to 50 kbit/s, instead of the channel switch technology, which was used earlier, was the important stage in the development of mobile systems.

The growing needs in data transmission and the necessity to increase the transmission rate led to networks improvement, to application of the new technology of high-rate data transmission in networks with channel switching **HSCSD** (**H**igh **S**peed **C**ircuit **S**witched **D**ata). Its implementation increases the data transmission rate from 9.6 kbit/s to 28.8 kbit/s, which causes the modification of software and exchange protocols at saving the existing infrastructure. The next stage of improvement of GSM standard systems is implementation of packet data transmission services **GPRS** (**G**eneral **P**acket **R**adio **S**ervice), which allows the provision of transparent data transmission (from subscriber to subscriber) over the IP with rate up to 115.2 kbit/s. Application of the high-rate types of modulation in already deployed systems permitted increase of data transmission rate up to 384 kbit/s without infrastructure transformation and growth of the occupied frequency band. The example of such technology application is **EDGE** (**E**nhanced **D**ata rates for **G**SM **E**volution), which is the improved technology of increase of data transmission rate in GSM networks using the modified eight-position phase shift keying 8 PSK. The adaptive channel

adjustment will permit to increase the rate in the radio channel up to 2 Mbit/s depending on subscriber demands and interference situation.

Further improvement of the packet data transmission technology, which provides the transmission rate increase in the direction from the base station **HSDPA** (**H**igh-**S**peed **D**ownlink **P**acket **A**ccess) and from the mobile terminal **HSUPA** (**H**igh-**S**peed **U**plink **P**acket **A**ccess), and the technology of wideband radio access resulted in the appearance of the third-generation (3G) network. In Europe, these networks, providing the high-rate transmission of multimedia information and the wireless access in Internet with succession saving with GSM networks, are known as universal mobile telecommunications systems (**UMTS**). Distinguishing traits of the 3G systems are:

- High equipment unification.
- Compatibility with services of stationary networks (within the limits of IMT-2000).
- High quality of subscriber service.
- Suitable portable subscriber equipment.
- Worldwide roaming.
- Rendering of multimedia services.
- High network flexibility providing the support of new types of services.

The Sim card contains digital data about the network which the subscriber belongs, individual user information, encryption programs, as well as list of foreign networks in which roaming is allowed. The subscriber's ability to work in a foreign network is limited only by the presence of multifunctional multi-band phone and tariff agreement. The subscriber having the **SIM** (**S**ubscriber **I**dentify **M**odule) of any network has a possibility to support the continuous communication with another subscriber at its motion within the own network or transferring in the adjacent network. The SIM card contains data in digital form about the network, in which this subscriber operated, its personal (phone) number, and other information allowing subscriber identification in any network, and a possibility of subscriber operation in another network is restricted only by the presence of multi-functional multi-range phone. GSM capabilities are not limited to telephone communication, since its improvement (GPRS) allows you to organize the transmission of packet data in standard channel at a speed of 22.3 kbit/s, 2.3 times higher than the speech transmission rate in the basic version of standard. At the same time, the packet organization of transmission/reception allows you to reduce the load on the channel, reducing its busy time to the intervals of receiving or sending a message. The PCS system is a version of the GSM1900 standard, which provides personal communication for not two mobile stations but two subscribers, but, at that, it does not require the cellular technology application.

Network of the Third Generation Construction of new personal communication networks (PCS) uses digital technologies of multi-station access with time (TDMA) and code (CDMA) channel division.

The technology of temporary multi-station access (TDMA) has been most widely used in second-generation systems in USA, the IS-136 system, in Japan, the PDC system and the European GSM 1800 personal communication system. Its characteristic feature is the use of a single carrier in radio channels, and a message from a separate source is transmitted in a time interval (slot) assigned to it. Systems based on the CDMA technology (IS-95) realize the principle of the direct spectrum spreading, when the whole radio channel band is simultaneously used for data transmission from each message source. In this case, the radio channel bandwidth is significantly wider than the band at the transmitter modulator output, which is occupied by the separate message. The great contribution in the theory of code channel division belongs to Russian scientist D.A. Ageev, who in 1935 published the main position of his theory in his famous book in Russian language. Implementation of the technology of orthogonal digital signals for construction of noise immunity systems belongs to famous American scientist C. Shannon [15].

The main requirement in the development of 3G system is providing of video-communication with resolution not less than 320×240 , which is possible only at information transmission rate of 384 kbit/s. The first step in this direction was made in systems of IS-95 standard using the wideband signal (1,25 MHz), which provided the increase of data transmission rate in downlink up to 144 kbit/s. Taking into consideration that the channel is not symmetrical at data transmission, it enables to decrease the transmission rate to 20 kbit/s in opposite direction. The information transmission from each network subscriber occurs at the constant rate in the total frequency range, and for extraction of data stream intended to the specific subscriber, on the receiving side, the pilot signal is used, i.e., the unique code adopted to each subscriber terminal.

The unified IMT-2000 (International Mobile Communication) standard is built on this principle, which includes the open systems cdma2000 and UMTS. The sub-system cdma2000 EV-DO provides the transmission rate in downlink up to 2 Mbit/s, and the UMTS system forms the radio signal with transmission rate of 3.84 Mchip/s at the radio channel width of 5 MHz. Integrated by the single technology of channel division, these systems have the serious difference at construction of networks: cdma2000 is the synchronous system, while UMTS is the asynchronous system. In the first case, all network components are unified by the mutual network time providing its synchronous interaction, which simplifies the network structure.

In the UMTS network, any terminal equipment must form the autonomous time, and the necessary interaction accuracy of separate nodes of the single network or other networks is provided by the transmission of the reference signals from the active base station to the subscriber terminal; these reference signals have the correlation function of the δ -function type. Application of the frequency synthesizer (spanned by the PLL system) in final devices together with matched filtering or the RAKE receiver provides synchronous interaction of network elements. The principle of group signal formation with application of the orthogonal spread sequence and correlation processing of received signals is posited in the basis of receiver and transmitter interaction in this standard.

The main tendency of 3G network construction assumes the minimal application of the existing infrastructure of the 2G communication network with implementation of new network elements during its modernization (assimilation of new frequency ranges, multi-sector antennas, spectrally effective methods of modulation). Networks of the third generation, besides the communication services, high rates, and large volumes of transmitted information, render the subscriber the following services: information reference, entertainment, education, mobile telemetry, mobile commerce, navigation, etc. independently on object location. The architecture of the UMTS (Universal Mobile Telecommunications System) network on the physical level consists of the subscriber equipment domain and the infrastructure domain, which interact with each other through the set of reference points (interfaces). In compliance with IMT-2000 requirements, the system architecture is divided into the network of radio access **UTRAN** (UMTS Terrestrial Radio Access Network) and the corner network (CNk), as well as the user equipment (UE).

Further developments of UMTS networks in the direction of data transmission rate resulted in development of **HSPA** (High-Speed Packet Access) technology using the multi-position types of modulation (16QAM, 64QAM). Appearance of more advanced technologies of the multi-frequency access, modulation, and signal generation with the multi-frequency orthogonal modulation (OFDM) resulted in development the multi-station access networks **E-UTRAN** (Enhanced UMTS Terrestrial Radio Access Network) and the basing network **SAE** (System Architecture Evolution).

Another direction of the 3G system evolution is perfection of the IS-95 standard; it resulted in the appearance of the mobile communication system cdma2000. Parameter optimization of such systems affecting on appearance of cdma2000 EV-DO (Evolution-Data Optimization) systems, and later revisions of EV-DO protocols, resulted in the creation of 4G networks of **UMB** (Ultra-Mobile Broadband) version, which provides the peak rate in downstream up to 300 Mbit/s and in the upstream – up to 100 Mbit/s.

In 3G and 4G systems, in contrast to the 2G system, the principle of data transmission with packet switching is laid in the basis, when on levels of OSI model upper than the network level (Fig. 1.10), the data exchange between adjacent networks (systems) is performed in form of packets only. Each packet, which enters in a device in different time intervals and from different sources, has an “address” indicating on its belonging to the specific frame and a queue of its location in this frame structure. This eliminated a necessity of the channel fixation for subscribers during connection time, making the algorithm of signal processing more complicate.

Another specific feature of these systems is application of phase types of modulation of various complexities, beginning from the simple BPSK to the quadratic (QPSK) and multi-level amplitude-phase modulation (nQAM). This predetermined the utilization of correlation methods of signal processing: non-coherent signals for the group signal formation in the transmitter and optimal demodulators in receivers.

Both in the 3G system and in 4G one, the radio channel is used with fixed bandwidth, in which data exchange occurs of all subscriber terminals of the separate cell and the interacting base stations. If in the UMTS or cdma2000 systems the radio front-end bandwidth is strictly limited, then in the LTE system, this bandwidth can be different and is defined by the cell dimension, the traffic intensity, etc.

The principal difference of these types of system consists in methods of multiplexing and approaches for separate channel extraction from the total frequency band of the radio front-end. So, in 3G systems, the code multiplexing and channel division based on application of orthogonal channel-forming sequences are used. The unique (etalon) sequence allocated to the specific subscriber terminal is orthogonal in time to all others used in this cell, which allows separation of the signal belonging to this subscriber terminal from others by means of its comparison with signals acting on its input. The channel-forming sequence is simultaneously the signal spectrum spreading sequence and the measure of struggle with the multipath property. The impact on the receiver input of the direct visibility signal from the base station and its echo copies leads to intersymbol interference, which is one of the most dangerous types of interference in systems of mobile communication, which decreases their sensitivity. The effective method of struggle with this phenomenon is the application of the matched filter in the radio front-end, which is invariant in time. Another variant is the parallel processing of several signals received by subscriber terminal in the RAKE receiver, which removes phase shifts in echo copies with regard to the direct visibility signal.

Cellular Networks of the Fourth Generation In 2009, the development of the up-to-date technology **LTE** (Long-Term Evolution) was finished, the last version of which LTE Advanced (4G version) will provide in the downstream the informa-

tion transmission rate up to 1 Gbit/s and in the upstream up to 500 Mbit/s when using radio channels with bandwidth of 100 MHz.

The complicate enough but highly efficient OFDM technology worked through in systems of IEEE 802.11 standards, and the wideband access WiMAX has got application in the 4G system of cellular communication. It is based on the formation of large number of orthogonal subcarriers in time, the modulation of which is also performed by slow orthogonal sequences obtained from the information sequence. Mutual orthogonality between modulated subcarriers in the frequency domain is ensured by the choice of the duration of the modulating symbols.

Application of slow modulating sequences provides the reliable protection from the intersymbol interference caused by the signal multipath in the radio channel, as the edge distortions of the long modulating pulse happen in its initial part. A loss of mutual orthogonality of the modulating symbol can be recovered by replacement of its initial part by the process segment at the end of this pulse called the protection interval and saving the orthogonality. Thus, the mutual orthogonality is saved in the whole length of modulating symbols. The group from the 12 sub-carriers forms the OFDM symbol, each of which is added by the cyclic prefix to save its mutual orthogonality in the group. Six (or seven) OFDM symbols form the radio block depending on the length of used cyclic prefix. For the specific frequency band allocated to an operator, the necessary number of radio blocks is chosen.

The important stage of formation of the large number of orthogonal quadrature components of sub-carriers modulated in phase (QPSK or nQAM) is solved either by circuitry approach or the software method – by means of execution of the Fourier transform. Effectiveness of the Fourier transform is provided by a choice of a number of conversion points multiple to 2. At that, obtained “redundant” sub-carriers remain non-modulated and are located near the boundary of the frequency band occupied by the radio channel.

2.5.1 The Cellular Network of the GSM Standard

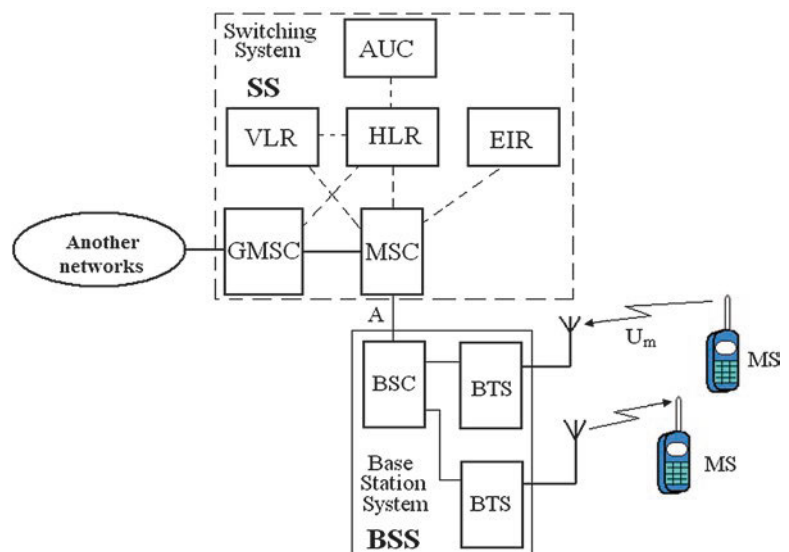
2.5.1.1 The Network Structure

The global system GSM of mobile communication appeared owing to efforts of a series of countries (mainly, European), which formulated requirements to the etalon model of the open systems (OSI). This allowed solution of interaction problem of developing digital systems on the base of requirements formed by consumers. The main requirements, which were arisen at development of the GSM systems, are the speech, data, and control signal transmission, which provides interaction of devices of own and adjacent networks. Restriction of used frequency range and digital signal processing methods required application of the highly effective modulation methods providing the noise immunity of radio signals.

The classical system of digital communication of the GSM standard was developed for speech signal transmission between the mobile subscribers. The system service zone is divided into fragments, one of which is presented in Fig. 2.14.

As additional services, the possibility to transmit short messages without usage of any additional external devices and the facsimile transmission was anticipated. The service of data transmission is possible using the narrowband channel for speech

Fig. 2.14 The fragment of system service zone



transmission, but to support this service, it is necessary to connect a computer or a server at both ends of this connection. In this case, the GSM system provides the data transfer only, not having gone into details of the type of rendered service. The system of the GSM standard can operate in two modes: with channel switching and with packet switching – GPRS.

The mobile station (**MS**) is connected through the radio interface (U_m in Fig. 2.14) to the radio access network **BSS** (**Base Station Sub-system**), which contains the base transceiver station **BTS** (**Base Transceiver**) and the base station controller (**BSC**). Together with other identical structures, local networks are connected through the A-interface to the switching center of mobile communication **MSC** (**Mobile Switching Center**) of the core network (**CN**). The core network is intended for connection of different elements of the radio access system with the fixed part of the system and other networks, for example, PSTN, iSDN, PSDN, and the Internet.

Each MSC is connected with appropriate visitor location register (**VLR**), which contains the necessary information about mobile stations temporarily located in the service area of the local MSC. Besides VLR, the core network is equipped by other registers:

- **HLR** (**Home Location Register**) having information about permanently registered mobile stations in the system of the specific operator.
- **AuC** (**Authentication Center**) having information about the list of system services admitted to utilization by the subscriber, who has the produced subscriber identity module (**SIM**).
- **EIR** (**Equipment Identification Register**) the database of serial numbers of base stations used in the system (including numbers of lost or stolen phones prohibited to utilization).
- **OMC** (**Operating and Maintenance Center**) provides operation of separate elements of the GSM system. It is connected to all elements of the mobile communication switching network and performs functions of administration (tariffing and traffic monitoring) and takes measures to eliminate failures of separate network elements. OMC is connected with other elements of the GSM system with the help of fixed communication networks.
- **GMSC** (**Gateway Mobile Switching Center**) is the connection of several MSC belonging to several network fragments, which fulfilled functions of gateways into external networks, such as PSTN, iSDN, and PSDN.

The HLR register represents the central database, in which information is stored about the permanent network subscribers and information about the current location. Data in HLR transmitted from VLR allow determination of current location of called subscriber, and according to the information from the HLR, we can provide the connection with those MSC, information about which is stored in VLR.

Functions of performing and supporting of subscriber connections, the dynamic resource control in the service zone, and connection rerouting at subscriber transfer in the cell, which is in service of another BSC, are entrusted with the switching center of the mobile network (**MSC**).

To organize the double-sided communication in the GSM system with frequency duplexing, which separate receiving and transmitted data streams, two frequency ranges are allocated, each with the width of 25 MHz. The frequency range 890...915 MHz is used for data transmission (in uplink communication) from the mobile stations (**SS** and **MS**) to **BS** (base stations). The range 935...960 MHz is allocated for data transmission from the base stations to the mobile ones (downlink). Both ranges are divided by the frequency interval of 45 MHz realizing the principle of the duplex data transmission – a possibility of simultaneous operation on reception and transmission (**FDD** – **Frequency Division Duplex**).

Both frequency ranges are divided into 124 frequency channels (bands) each with width of 200 kHz, and the carrier is located in the band center. Eight channels are arranged on each carrier using the principle of time channel division (**TDMA**). Thus, the separate carrier (or its sequence at frequency hopping during reception-transmission) is allocated to each connection and the specific time interval. At that, the message transmitted to the subscriber in the allocated time interval of the informational frame is transmitted in the next frame on the new carrier frequency. Application of slow frequency jumps allows decrease of group error probability in the transmitted signal due to fading reduction. Since Rayleigh fading is frequency-dependent, at transfer from one frequency, where fading occurred, to another frequency, fading will be absent with a large probability. At that, new frequency value can be rather near the previous value. Thus, fading probability decreases and hence the group errors.

During frequency jumps, the separation between carrying frequencies of the reception and transmission channels remains constant at 45 MHz. Such frequency offset allows maintaining its mutual orthogonality at simultaneous switching of all carriers in the single cell. The definite orthogonal formatting sequence is associated with each active subscriber, which excludes the mutual interference between subscribers in the cell. The sequence parameters are transmitted to the mobile station from base one during the channel definition. Adjacent cells use different orthogonal sequences.

The system of the GSM standard is considered as the system with frequency-time channel division (**TDMA/FDMA**). Such a form of the multi-station access organizes the physical communication channel as the sequence of time slots (desig-

nated by slot numbers), which are transmitted on the chosen carrier frequency. Physical channels exist in pairs – one in each direction, and they have the same number of the time slot, and their carrier frequencies are spaced by 45 MHz.

In accordance with the plan of frequency distribution for the cell or the sector, the time slots, which have the same number of carriers on the ascending line (from MS), are shifted in time by three positions with regard to the time slot on the descending line (from BTS). In consequence, MS cannot simultaneously transmit and receive the signal, what weakens interaction between the MS transmitter and receiver and reduces requirements to characteristics of the front-end and of the section of digital signal processing. This also excludes the possibility of simultaneous transceiver operation on the reception and transmission and simplifies implementation of the radio-frequency block and the block of digital signal processing.

2.5.1.2 Description of the Connection Settle Procedure

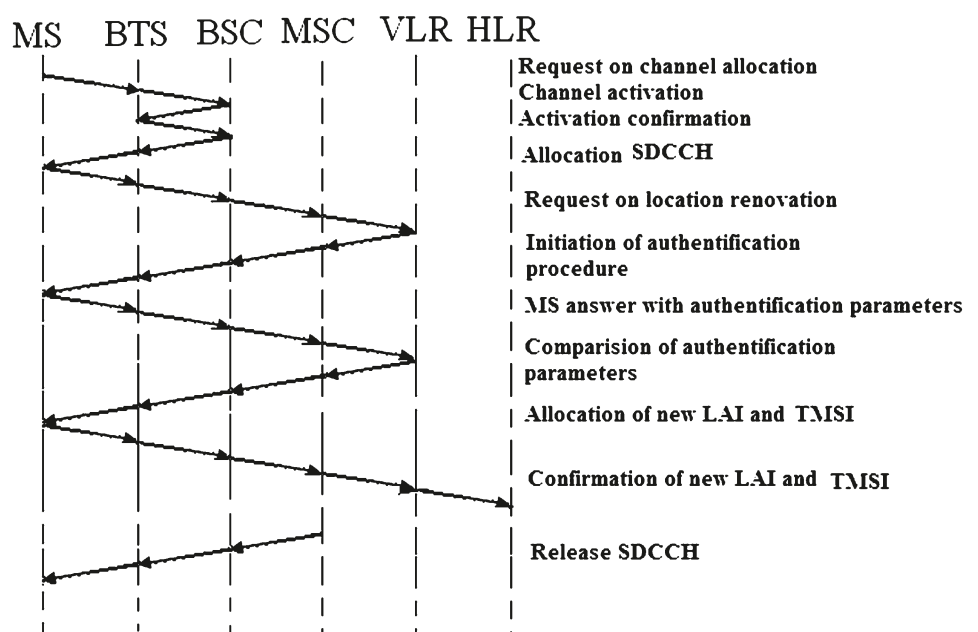
Preliminary Adjustment To settle the connection of the mobile station with some subscriber of the mobile or fixed network, it is supposed to fulfill a series of preliminary actions, which MS performs just after switching on the power supply.

At first, MS searches the carrier frequency, on which the broadcasting channel **BCCH** (Broadcast Control CHannel) is transmitted in the given cell. In such logical class, information for the control of the network, the specific cell, and adjacent cells is transmitted. For this, MS measures the power level of signals, which are transmitted by the base station on each from 124 carrier frequencies used in the GSM900 standard. The carrier frequency, which contains the broadcasting channel, has the highest power. On measurement results, MS forms the list of signals received on each carrier in the order of power decrease. Then, MS sequentially scans the carrier frequencies to find the frequency correction channel (**FCCH**). Finding of this channel allows synchronization of the reference MS oscillator frequency, which is equivalent to finding the zero time slot of the broadcasting channel carrier. In zero time slots of the following frame, MS finds the other important control channels and, first of all, the synchronization channel (**SCH**) and decodes information contained in this channel: the base station identity code (**BSIC**) and reduced frame number (**RFN**). Then, the information received in the broadcasting channel from the 16 strongest carriers of adjacent cells, on which the global cell identifier and measurement results of the six strongest carriers and other parameters are transmitted.

Registration of the Mobile Station To have a possibility to call MS, or to initiate from MS any connection, it is necessary to perform the registration procedure: MS with the help of the random access channel (**RACH**) sent to the BTS the request on registration fulfillment (Fig. 2.15).

The base station sends the request to the base station controller using the A-interface (see Fig. 2.15). The controller allocates BTS in allocated fixed control channel (**SDCCH**) intended for the preliminary information exchange and informs BTS about it. The base station confirms information about the channel allocation for BSC and informs MS via the AGCH chan-

Fig. 2.15 The request on registration fulfillment



nel – the channel of access permission, about the SDCCH channel allocation. Then MS sends via the SDCCH channel the request on the location renovation procedure, which contains the temporal identifier of the mobile **TMSI** (Temporary Mobile Subscriber Identity) subscriber and the identifier of the location area **LAI** (Location Area Identity). These parameters are stored in the MS memory. The MS request falls into the switching center of the mobile communication MSC and the visitor register of position **VLR** (рис. 2.14) through the base station BTS and the BSC base station controller. If there is TMSI in VLR, then information renovation about MS activation happens. If the temporal identifier of the mobile subscriber is absent in the VLR memory, then the location area indicator is encoded, which was sent by MS. According to this sequence, the VLR is indirectly determined, which is served by the MS during the last connection, and after information exchange between the previous VLR and the current one, the international identification number of the mobile subscriber **IMSI** (International Mobile Subscriber Identity) is conferred and other data. In the absence of data about the last location in VLR, MS receives the permission on the single sending of its IMSI. Sent IMSI falls through MSC and VLR in the home location register (**HLR**), where subscriber identification occurs, and this information is loaded into the current VLR.

Identification Procedure Positive results of the authenticity check of the mobile station define new MS parameters – the temporal identifier of the mobile station (TMSI) and the location identifier (LAI) – and information is updated in VLR and HLR. The new TMSI number is sent to MS in the ciphered form, which in future increases the protectability of transmitted information via the radio channel. The assigned number TMSI is true only on the time of MS being in the service zone of the current VLR. After performing of full MS synchronization with the network, the initiation of connect settle is possible. The mobile station scans the common control channel (**CCCH**) to find the paging channel (**PCH**) intended for this MS.

To perform the call, after finishing the registration and identification, the MS sends a random access packet to the base station BTS, thus realizing the logical RACH. The BTS sends three packets to the MS, constituting the logical AGCH access channel. The access grant channel (**AGCH**) contains a number of the fixed control channel, which is allocated for MS during settle time of connection, and conditions of signal transmission.

Then the mobile station sends the confirmation of MSC about channel obtaining. The base station, in turn, measures interferences in time slots, which are free in the current moment. The center of operation and maintenance (**OMC**) performs the monitoring of the current loading level. These values can be also used at connection transfer.

Formation of Information Sequence Structural transmitter diagram of the GSM900 mobile station (Fig. 2.16) reflects the main stages of input signal processing.

The speech signal in a microphone is converted into electric signal, the bandwidth of which is limited by the frequency $F_h = 3.4$ kHz.

Then the signal is digitized in ADC with the frequency $f_d = 8$ kHz, the value of which satisfies to the Kotelnikov theorem: $f_d \geq 2 \cdot F_h$. ADC should be linear, which is caused by the operation character of the speech encoder. Each quantization level is coded by the 13-bit binary sequence, which leads to digital flow formation at the ADC output with 104 kbit/s. At simultaneous operation of eight subscribers (TDMA/FDMA), the digital flow rate increases up to $8 \cdot 104$ kbit/s = 832 kbit/s, which cannot be transmitted in the bandwidth of the single physical channel equaled to 200 kHz.

To reduce the digital flow rate, one can perform the segmentation and speech encoding. The digital sequence is *divided into blocks* (segments) with the length of 20 ms, each of which is processed by the speech encoder with the method of long-term prediction with excitation of the regular pulse sequence (RPE-LTP). Each 20 ms block is represented by 260 bits. Division of the initial sequence into segments allows decrease in the digital flow rate. This is explained by the fact that in the time interval of 20 ms, we can neglect by the speech signal variation, which allows its' transmission by a single bit only

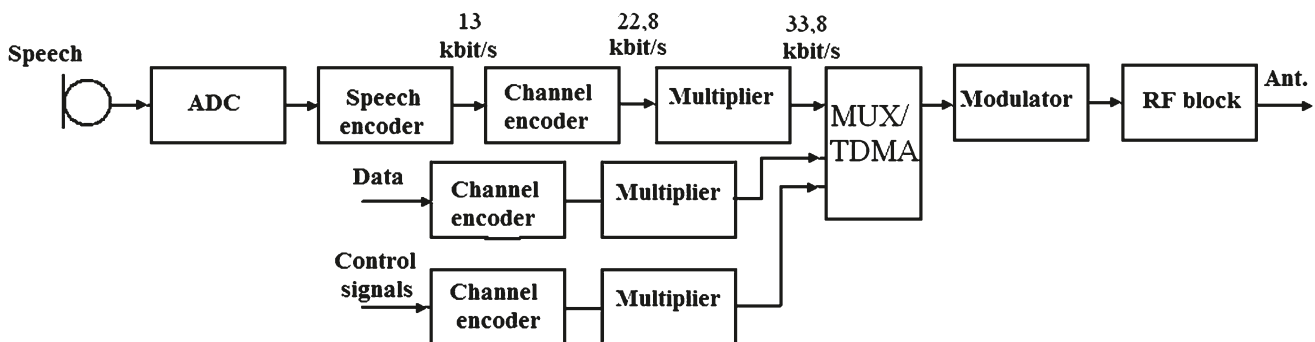


Fig. 2.16 The structural transmitter diagram of the GSM900 mobile station

instead of the pulse train in this interval. In other words, instead 8000 samples per one second, we can transmit 50 samples only. Thus, the procedure of digital sequence conversion from the PCM encoder output into the sequence forming on the speech encoder (trans-coding) reduces the digital flow rate up to $50 \cdot 260 = 13 \text{ kbit/s}$, which provides the speech signal transmission quality comparable to the quality in PSTN. At simultaneous operation of eight subscribers in the channel, the rate of resulting digital flow will be $8 \cdot 13 \text{ kbit/s} = 104 \text{ kbit/s}$.

For effective decoding and error dispersion, one can use the channel coding and interleaving. The channel encoder with the coding coefficient $R = 1/2$ allows error correction at message reception. In GSM, the block data interleaving is used, when the describing speech fragment with the length of 20 ms is recorded in the matrix of random access memory (RAM) containing 8 columns with 57 bits each. Then, these bits are read in columns. So, 8 segments with 57 bits each are formed. Segments are distributed in eight successive packets located in allocated time slots.

The digital sequence formed from sequences from different sources passes to the multiplexer, which forms the TDMA packet. Since the width of the single channel is only 200 kHz, this gives high requirements not only on the modulation type defining the radio signal spectrum width but also on its compactness at low level of side lobes. To achieve the high noise immunity of the radio signal, it is necessary that its envelope becomes constant, which is especially important for mobile stations. This allows utilization of the nonlinear mode, having high effectiveness, in the active elements of the transmitter power amplifiers.

Data transmission in the GSM standard can occur with rates 2400, 4800, and 9600 kbit/s. For each rate the appropriate convolution code is used with the coding coefficient $1/6 \dots 1/2$ and interleaving. At fulfillment of these procedures, the much larger time delay happens compared to the delay at speech transmission. However, if the delay is the important parameter at speech transmission, then at data transmission, the requirements to the error in frequency will be higher. This can be achieved by application of channel coding and the deeper interleaving. Much more rigid requirements are made to the control data transmission.

Coded data blocks in input of the interleaving system (Fig. 2.16) are formed in the TDMA packet providing the accuracy of pulse location of the sequence $1/4$ bits with the help of the synchronization system and are directed to the modulator. In the GSM standard, we can recommend to use the Gaussian minimal shift keying (GMSK), which is the specific case of the frequency shift keying (FSK) with the continuous phase. The difference is only in the modulation index value, which is $BT = 0.3$ for GMSK. The digital sequence after conversion to the form **NRZ** (Not **R**eturn to **Z**ero) is passed through the Gaussian filter and then to the modulator. The modulator represents the retuned VCO, initial frequency of which is given by the frequency synthesizer. In the operating frequency band of the GSM900 standard (Table 1.2), 124 channels with bandwidth of 200 kHz are formed and in the GSM1800 standard 374 channels with the same bandwidth. The central frequency for each channel can be chosen both as constant and as changing in jumps 217 times per second, taking into account the frequency distribution plan between the cluster cells. The signal from the modulator output passes to antenna through the radio-frequency module.

2.5.1.3 The EDGE Technology

In spite of the presence in the GSM systems of a large number of services connected with data transmission, the main shortcoming is the low transmission rate. The GSM standard defines the data transmission rate not more than 9600 kbit/s and in special cases 14.4 kbit/s. At that, the binary data rate at the channel encoder output is 22.8 kbit/s, independently on the type of transmitted signals, either speech or data. In order for data transmission rate corresponds to the rate 22.8 kbit/s adopted in the data packets, it is necessary to use the special correcting codes, the form of which depends on the rate of input data packet. Simultaneously, we need to adapt it with the data flow rate of different subscribers with channels joining the base transceiver station (BTS), the base station controller (BSC), and switching center of the mobile network (MSC), which have the data transmission rate of 16– or 64 kbit/s. In real conditions, the data transmission blocks form the structure parallel to the existing for speech transmission and ending by the terminal reproducing the obtained data (a computer, the measuring device). In this connection, the GSM network can be considered as the access network to other telecommunication networks, such as PSTN, the integrated digital enhanced network (iDEN), or the integrated packet switched digital network (iPSDN).

Data Transmission We can distinguish the service of data transmission such as the short message service (SMS) technology permitting to send and to receive short text messages with the help of the mobile phone and the data packet transmission (GPRS).

SMS Factually, the short message transmission corresponds to paging with confirmation of message reception. If the message recipient is situated outside the network covering area or his mobile phone is switched off, the system stores the message until the subscriber radio station will appear again in the network. The SMS message can be delivered to the subscriber

in parallel with the speech transmission. This message can have length up to 160 bytes, which is transmitted by 2140 bytes with the help of a specific alphabet via the allocated control channel.

HSCSD The first approach for data transmission rate growth, not requiring the essential variations in the structure and operation organization of the system, has got the widest distribution and is called high-speed circuit-switched data service (*HSCSD*). The aim of HSCSD introduction is integration of services for subscribers with different rates in the structure of the single physical channel. The available capacity of HSCSD connection is several times more than the capacity of the channel full-rate traffic, which leads to significantly widening of possibilities at data exchange. The claimed maximal rate of the main data flow is 14.4 kbit/s, and at using of the HSCSD technology, it achieves to rates available in the wired lines: 57.6 kbit/s by joining of several time slots (channel intervals) with rates 14.4 kbit/s. The growth of the data transmission rate from 9.6 kbit/s to 14.4 kbit/s leads to insignificant reduction of the limit value of the code correcting ability, which is defined by the protocol of the GSM system radio front-end and permitting the subscriber to increase the data transmission rate.

Theoretically, if all eight time slots will be allocated to the single connection, then using the channel full-rate traffic (9.6 kbit/s), we can obtain the transmission rate of subscriber data equaled to $8 \cdot 9.6 \text{ kbit/s} = 76.8 \text{ kbit/s}$. Logically, all traffic channels of the HSCSD radio interface belong to the single HSCSD configuration and are considered by the network as the single radio channel. Each channel from included channels is exposed to the same channel coding, interleaving, the procedure of the rate adaptation, and ciphering. The minimal number of traffic channels, acceptable channel coding, the type of the remote modem, and the rate value in the fixed network for realization of the chosen service are requested by the subscriber in the connection moment. Application of HSCSD does not require the channel switching, as in the standard GSM version, which causes traditional distortions typical for data transmission.

GPRS The next stage of the data transmission rate is an evolution in the service of the short message transmission (SMS), such as the exchange of fast message and the service of non-structured additional data, which led to the appearance of the GPRS service: the packet data transmission on the base of TCP/IP and X.25 protocols, which allow the rate to increase up to 115.2 kbit/s. There are several reasons for its appearance: data transmission rates in existing networks of the mobile communication are not enough, and connection time is rather large. The growth of the transmission rate up to values, which are necessary for Internet applications within the frames of the GSM standard, is possible by means of:

- Allocation to the single connection of several time slots in the single frame. At that, we ensure the communication saving with channel switching at growth of connection capacity.
- Packet switching with a possibility of allocation of several time slots of the frame for transmission of the data packet.

The GPRS system anticipates operation in multi-slot mode, when *up to eight time slots* can be allocated to the mobile station, if it is possible from the point of view of system resources. In addition, during time of the packet transmission or reception in the descending and ascending communication lines, the various resources are allocated, which leads to asymmetry of the packet traffic. In intervals between packet's transmission, this channel can be used by other mobile stations included in the GSM system. In this case, the same physical channels are used, and system resources are used more effectively owing to statistical multiplexing. In the simplest case (without application of redundant coding), the data transmission rate increases up to 171.2 kbit/s for allocation of eight time intervals to the single subscriber.

After connection with the switching network, the mobile station transfers into the ready mode, and in order to begin the data exchange process, it must get an address of the packet data protocol (PDP). At the presence of information about the subscriber (for instance, another mobile station connected to the Internet network of the IP), the connection occurs on the GPRS network with the required network through the gateway of the base IP network, and data transmission takes place. Since GPRS does not require allocated channel between subscribers, the network resource and the bandwidth are spent only in the interval of their active interaction. At that, the system effectiveness significantly increases, which factually equals to effectiveness of the Internet traffic, when short data packets of the active periods are dispersed in relatively long intervals of low activity of the physical channel.

The **EDGE** (Enhanced Data rate for Global Evolution) technology is the perfection (expansion) of GPRS and HSCSD systems simultaneously, which was the base for creation of the high-speed system IS-136 HS in the USA. At present, EDGE for TDMA system is realized in two variants: *EDGE Compact* and *EDGE Classic*. In the EDGE Compact system, the new structure of the control channels with 200 kHz width is used. Base stations operate in synchronous mode, and to realize the EDGE technology it is enough to have a spectrum bandwidth of 1 MHz. Transmission channels of packet data is organized with the frequency reuse coefficient 1/3. In the EDGE Classic system, the traditional structure of control channels with the width 200 kHz is used, which is applied in GSM. The coefficient of frequency reuse of the broadcasting control channel is 4/12 and for minimal integration requires the presence of 12 carriers or, in other words, 2.4 MHz together with protection bands.

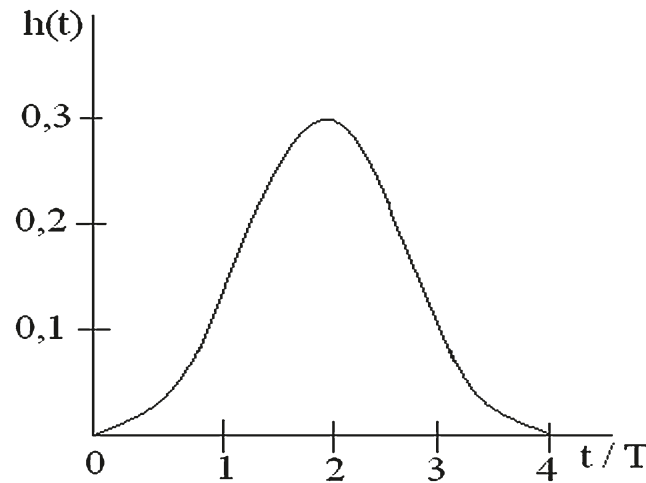


Fig. 2.17 The shape of the filter pulse characteristic

In the EDGE system, the superstructure over GSM, for high-speed transmission of packet data, the modulation of 8-PSK is applied, which allows increase of data transmission rate by three times compared to the standard GSM under condition of usage with the same symbol rate of 270, 833 kbit/s, and the frequency band of 200 kHz. This is explained by the fact that at the 8-PSK modulation, each information symbol contains three bits of information. At the same time, besides the phase shift defined by the information sequence, the phase is shifted additionally by $3/8\pi$ at each next information symbol. This allows avoiding of low levels of the envelope and saving the ratio of maximal and average powers not less than 3.2 dB. In the low-rate modes, the GMSK modulation is used as usual. In order to make the signal corresponding to the channel spectrum width and to save the GMSK-spectrum shape, the special pulse shape is used formed for the linearized system according to the Loran method [88]. The shape of the filter pulse characteristic $h(t)$ for pulses' formation in the 8-PSK modulator of the EDGE system is presented in Fig. 2.17. The shape of the filter pulse characteristic at the modulator input is similar to the Gaussian curve.

Another feature of the EDGE system is the slow jump-like frequency retuning, which allows reduction of the signal fading influence. Simultaneously, the co-channel interference level is essentially reduced, and hence, the system capacity increases. At usage of the jump-like frequency retuning, the encoding block is transmitted on four different carriers.

The next important property of the EDGE system is the quality control of the radio channel, when the subscriber radio stations transmit the information about the reception quality to the base station. This defines the choice of the modulation type (8-PSK or GMSK) of the base station, and depending on the capacity, it allows a choice of the one from nine coding rates. Switching between coding combinations and the modulation allows obtaining of the maximal data transmission rate per one carrier of 556.8 kbit/c, at utilization of the 8-PSK modulation, and 185 kbit/s for GMSK.

2.5.2 Wideband Mobile Communication Systems

2.5.2.1 Methods of Wideband Signal Generation for Systems with the CDMA Technology

CDMA (multi-station access with code channel division) is the modern technology of wireless digital communication. It provides the capacity 6..0.8 times more of this network compared to the analogous communication (e.g., AMPS) and more by 4 times than the used TDMA technology of the digital network. The quality of the telephone message is significantly higher compared to any other cellular system, especially in complex radio engineering conditions, such as the dense urban building or the mountain region. Both at the organization stage and in the case of the long-term operation, CDMA provides the most optimal (from the point of view of a cost) solution. The data transmission rate exceeds the speed of the telephone messages, and new speech codec provides the speech quality close to the quality of wired lines. This technology performs an access to the Internet and was the initial point to development of the mobile systems of third generation (3G).

It is known from the information theory that an information quantity, which can be transmitted via the channel with additive white noise, achieves its own maximal value, called the *channel capacity*, at the signal in the input, which has the Gaussian distribution. To provide such a character of the input impact, we can by either increasing a number of simultane-

ously acting sub-carriers (having the same power) or expanding the signal spectrum until the signal level would be below of the noise level. In the first case, the random process is formed at the receiver input as the result of adding of non-correlated narrowband signals. This can be achieved by application of complex technical procedures of signal processing on the transmission side: the trellis-coded modulation, equalizers application, etc. Another approach of wideband system construction has got the widest distribution, when spectrum spreading is realized by the method of the *direct sequence* CDMA-DS (Code Division Multiple Access Direct Spread) or usage of *frequency hopping* CDMA-FH (Code Division Multiple Access Frequency Hopping) or usage of *time hopping* CDMA-TH (Code Division Multiple Access Time Hopping).

The Direct Sequence Method The very widespread type of the *system with spectrum spreading* is **CDMA-DS** (Fig. 2.18), when (on the example of cdma2000) the spectrum of the digital information signal with the rate of 19.2 kbit/s is spread by means of replacement of each information bit by the spread pseudo-random sequence with the length of 64 chips, each of which has the duration of about $0.8138 \mu\text{s}$.

On the transmitting side, informational sequence is exposed to the channel coding, which equalizes the digital sequence rates from different sources by addition of the definite quantity of special informational bits. Then this sequence passes to the interleaving device, which performs the error dispersion arising at signal passage via radio channel. The informational sequence from the encoder output is recorded sequentially in lines of the two-dimension matrix, and after its filling, the bit reading is provided in columns. The informational sequence consisting of bits with duration of T_b has the spectrum width of W . In CDMA-DS systems, each bit of information signal is replaced by some fixed digital sequence of definite length: the signal base. Each element of the binary sequence is called the chip and has the duration $T_c = T_b/W$. The spectrum width of this pseudo-random sequence (B) by W is times wider than the spectrum of the information signal ($B \gg W$), and its time behavior is close to the noise properties. In such noise-like signal, a unit and a zero may be coded by the inverse sequences. For each channel, the definite sequence is specified. Spread sequences are usually chosen as orthogonal (the scalar product of any two sequences equals to zero). Then, the signal passes to a modulator, where the spectrum of pseudo-random sequence is transferred in the radio-frequency range. For this, some variety of the phase shift keying is usually applied.

The transmitted signal achieves a receiver passed through the radio channel, which has Rayleigh and frequency-selective damping. The signal arrives to the receiver input by different ways having various attenuation and influence of the environment properties. Since all transmitters of mobile stations form the sequence of partially overlapped pulses, the application of the same carrier frequency leads to the appearance of a large number of simultaneously acting signals from different sources in the receiver input. To extract the signal from wanted station on the background of their mutual interference, which is caused by multiple access (**MAI**, **Multi-Access Interference**), the receiver (after demodulation of received signals and further filtering) performs the operation of de-spreading (inversed spreading) and forms the narrowband signal.

To recognize the noise-like sequence intended to the specific receiver from the similar received sequences, the receiver must know the pseudo-random sequence, which is used in the transmitter, and to be synchronized with it. All other sequences from other mobile stations have nonzero time shifts. Any sequences formed by mobile stations are orthogonal between them; therefore, at correlation processing of received sequences, the mutual correlation function between received and reference sequences is close to zero. The autocorrelation function achieves the maximal value, if the sequence under processing is synchronous and in-phase with known transmitted one. Existence of any time shift between correlated sequences of the same length leads to zero value of the autocorrelation function.

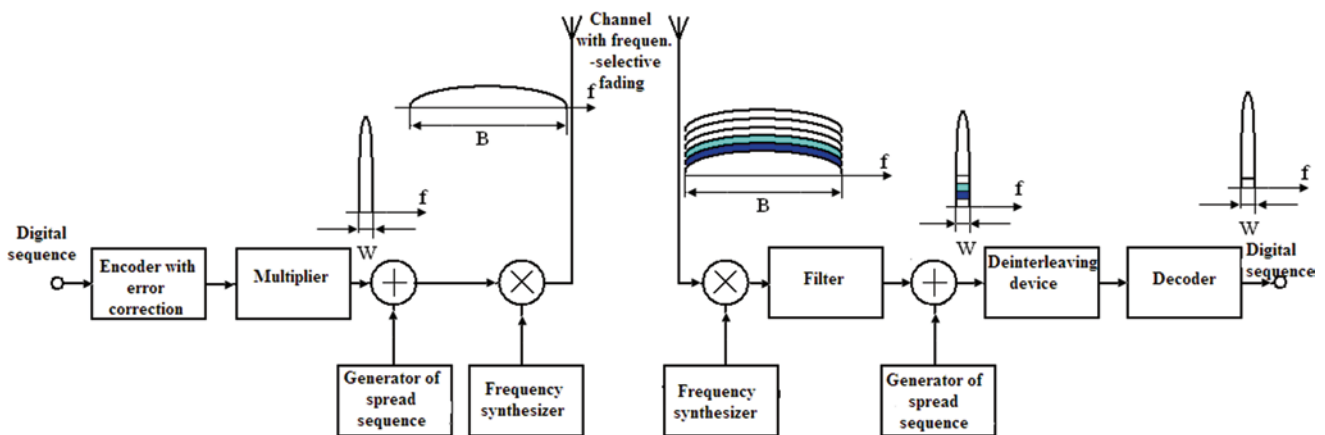


Fig. 2.18 The system with spectrum spreading CDMA-DS

Such a feature of signal processing, which uses the multi-station access with code channel division, provides the high system stability to the multipath signal propagation. The increased stability arises at chip duration, which is shorter than the minimal difference between delays at signal propagation by various ways. Owing to the fact that receiver correlation devices receive the one sequence only, several subscribers can operate in the same frequency band and use the different pseudo-random sequences. Due to broadbandness of the received signal, its input power can be equal (and even below) to the level of the white noise power.

The pulse sequence formed at the output of de-interleaving device is exposed to decoding and error elimination, which are arisen at signal transmission.

The Method of Fast Frequency Hopping Another method for system construction with spectrum spreading is based on the usage of *frequency hopping* **CDMA-FH**. Development of this variant of spectrum spreading of the transmitted signal is caused by the fact that it is rather difficult to provide the condition of synchronism and in-phase, which is necessary in CDMA-DS systems, at time-varying properties of the channel.

Information sequence acting at the encoder input of the system with the frequency hopping (Fig. 2.19) is added by the specially selected additional properties, which permit to correct (or, at least, to detect) errors in information blocks.

Using the encoder of **FEC (Forward Error Correction)** type with the direct error correction (without requirement of transmission repeat), it is possible to increase essentially the transmission rate in such systems. Pulses at the encoder output form the modulating sequence at the output of FSK modulator. The generator of the pseudo-random sequence controls by the frequency synthesizer. Frequency at the synthesized output may take $2^m - 1$ different values, and its current value is defined by m -sequential bits of the pseudo-random sequence (PRS) generator. Multiplication of two signals from modulator (FSK) output and the frequency synthesizer occurs in the mixer, which is equivalent to the additional shift in frequency by the value defined by the PRS generator. During the period of the FSK modulation, which is defined by the duration of the information transmission T_b , the variety of current frequency values of frequencies, in which values are defined by the chip combination (elementary signals) from the output of the PRS generator, is created on the mixer output, when the duration is defined by the hopping time T_h . The time interval T_h defining the current value of the current oscillation frequency is significantly less than the duration of the information transmission bit ($T_h \ll T_b$); therefore, we can tell about the *fast* frequency hopping. In GSM system, the frequency hopping is also applied, which occurs once per the time of one data packet transmission (148 bit), which concerns to the slow frequency hopping. Formation of short radio pulses with frequency shift keying created difficulties in detection and suppression of the signal source, which predetermined the use of these systems for military purposes. Reduction of the hopping time T_h widens the radio signal spectrum, providing their mutual orthogonality and, accordingly, the simultaneous operation of many subscribers.

Due to the wide range of generated frequencies, it is very difficult to provide the in-phase character of transmitted signals; therefore, the non-coherent FSK demodulator is used in the receiver. The received signal is multiplied in the mixer by the reference signal, which is generated by the frequency synthesizer, the instantaneous frequency of which is specified by the PRS generator synchronized by the current received frequency.

The Method of Time Hopping The system of spectrum spreading with the pseudo-random retuning in time code division multiple access time hopping (**CDMA-TH**) (Fig. 2.20) divides the transmission duration of the one information bit T_b into time segments.

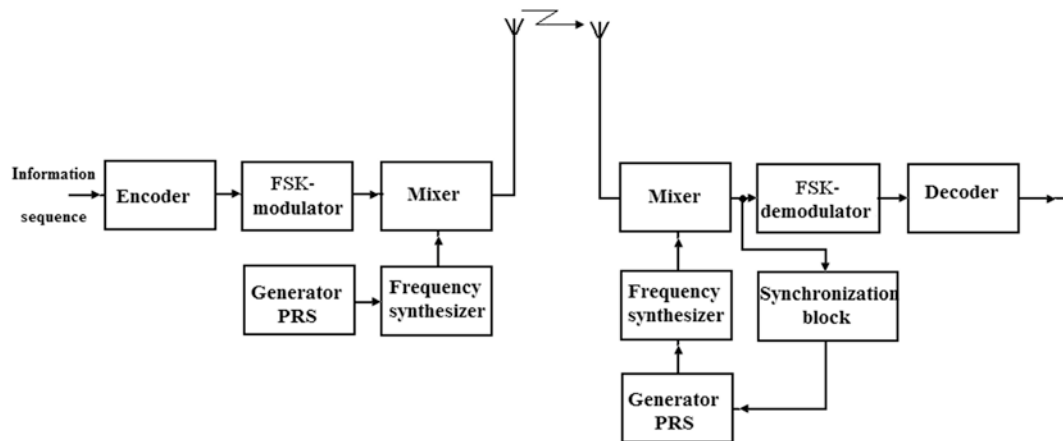


Fig. 2.19 The system construction with spectrum spreading based on the usage of *frequency hopping*

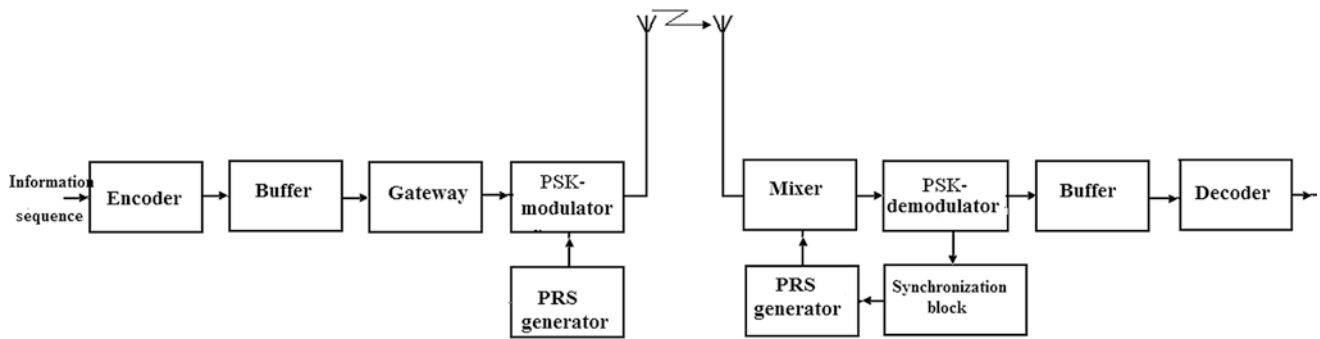


Fig. 2.20 The system of spectrum spreading with the *pseudo-random retuning in time*

The period of transmission is divided into M -time steps (M_{ts}) – time slots, which are usually about 1000. The signal is transmitted during the one time slot ($1/M_{ts}$). The PRS generator defines the time slot number for transmission of the informational signal. A choice of the large number of time slots for transmission of the single pulse and the random character of the allocated time interval for its transmission provide independence of radio pulses modulated in phase. Simultaneously, this leads to the appearance of serious problems with synchronization; the application of the packet information transmission is one of the approaches of this problem solution. Technically, this problem is solved by introduction in the information section structure of the transmitter and the receiver of the buffer, which provides the packet formation and its transmission in the strictly definite time moment and, relatively, its unpacking. Technically, the synchronism of receiver and transmitter blocks is the more difficult task than in the CDMA-DS system, but on the structure, this synchronism is close to its realization in the system applying the frequency hopping for spectrum spreading.

2.5.2.2 Systems and Networks of Third Generation

Evolution of Broadband Systems The existing variety of mobile communication systems (Fig. 2.1) and the cellular communication systems (Fig. 2.13) of the second generation were the consequence of the need of fulfilling specific tasks, which led to the creation of the mobile communication networks using various equipment. They operate only in those environments, for which they are developed. The rapid development of the Internet led to necessity while technologies to the need of manufacture of multi-functional ICs, which provide a possibility of creation of the universal system operating in any environments with usage of universal equipment.

Development of virtual home environment (VHE) is the global conception of the UMTS network. The fundamental property of VHE is possibility of the subscriber, who is at possibly larger distance, of condition realization for rendering of maximal list of services with indices, which are provided by the stable interface and the effect of presence. The VHE standard should not only to support an access to traditional services but also to provide a formation of new services.

Besides the data transmission rate increase in the developing system, the main task is realization of the matching condition of existing networks' standards and the developing network standard. This was the consequence of interest collision of the equipment manufacturers and the owners of the most distributed standards of existing networks IS-95 (Northern America) and GSM (Europe). At first stage, creation of the cdma2000 1x system and high-speed circuit switch data HSCSD and packet GPRS transmission systems (Fig. 2.21) was the development of these two standards.

In the cdma2000 1x system, which is the civil variant of the secret IS-95B standard, the data transmission rate provides 614 kbit/s, and the system capacity is twice increased. In the cdma2000 1xEV version, the data transmission rate is increased up to 2.4 Mbit/s in the bandwidth of 1.25 MHz, and in the cdma2000 3x system, the data transmission rate is 4.9 Mbit/s in the bandwidth of 3.75 MHz.

The growth of the data transmission rate due to improvement of the GSM standard by implementation of the high-rate transmission technology with the channel switching (HSCSD) is achieved by allocation to one connection of several time slots in the single frame with saving of communication with channel switching. The importance of GSM-EDGE systems, which is called also the 2.5G system, for modern operators consists in the fact that it allows the growth of the data transmission rate up to 384 kbit/s and potentially even higher at high indices of electromagnetic compatibility, which increases the spectrum utilization effectiveness and the carrier structure effectiveness. Such a technology (GSM-EDGE) using the base network GSM is an alternative to the WCDMA system. Another possibility to increase the resource effectiveness is the implementation of packets' switching (GPRS) by means of allocation of several time slots to the single data packet. In this

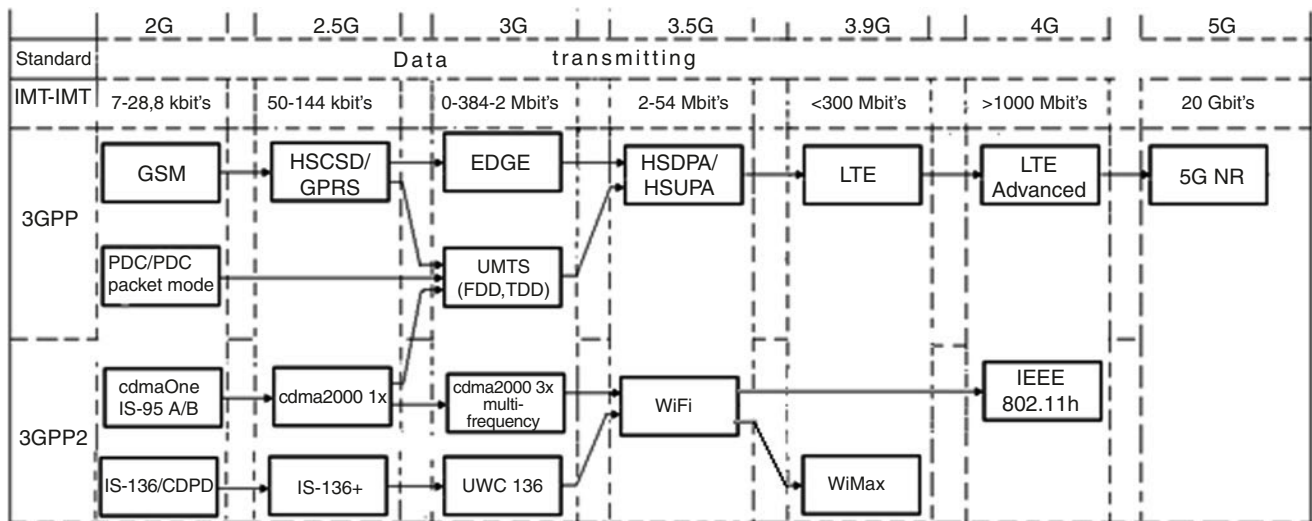


Fig. 2.21 Evolution of digital mobile communication systems

case, subscribers use the same physical channels, and the system resources are distributed more effectively due to application of statistical multiplexing.

The final solution of the 3G system structure appeared in expedience of development of two types of technology: W-CDMA and cdma2000. The W-CDMA technology has got the wide distribution, which was laid in the base of the ground-based mobile segment of European universal telecommunication structure **UMTS** (Universal Mobile Telecommunications System). The W-CDMA technology is based on the spectrum spreading by the direct sequence method (DSSS) in base of 5 MHz per channel and provides the data transmission rate in downlink of 2 Mbit/s and in uplink of 384 kbit/s. In the UMTS and EDGE (evolvable from GSM) systems, the GSM network is used as the reference network, while in the cdma2000 and EDGE (evolvable from IS-136) systems, the reference network is IS-41. Inter-network interaction allows rendering service of roaming and other services.

To achieve a high spectral efficiency of radio engineering system, it is necessary to carefully select the units included in the RF front-end of the receiver/transmitter that implements the system of multiple access to channel and satisfy the interface used. The comparable estimation of several different schemes is performed within the limits of the program of research in advanced communications equipment (**RACE**). Utilization of the hybrid CDMA/TDMA/FDMA technology with usage of advantages of each of them is one of the possible solutions, taking into account the varying requirements on the channel capacity, the informational loading, and the transmission quality at different combinations of cellular systems and PCS. The shortcomings of such systems of multi-station access consists in serious difficulties at transceiver realization, which has a low consuming power and a low cost at saving of high effectiveness of control in multi-level networks.

The main difference of 3G networks from 2G networks is the hierarchical cells' structure, which provides support to the wide series of multimedia services in network of various levels and uses the prospecting approaches of information transmission and new technological protocols. In 2G systems, as a rule, cells of the specific type are used, and frequency reuse is applied in the adjacent cells in such a manner that each separate cell controls its own covering zone and provides an adjustment within the mobile network including the traffic control and the handoff procedure control. The traffic supporting in each cell is fixed because there are frequency restrictions and possibilities of the transmitter control, in which parameters are optimized for the speech signal transmission and the low-rate data. The traffic growth leads to expensive reconfiguration of cells, when cells are divided and its sectorization is used.

The presence of breaks in coverings is eliminated by utilization of the multi-layer cell structure in 3G systems, by creation of pico- and micro-cells inside the macro-cell with large covering area. Global/satellite cells can be used in this sense to provide covering, when the macro-cell configuration is economically unprofitable and/or we need to provide the traffic along a large distance.

At low subscriber mobility or at small dispersion of delay time values, the high rate of transmission and the high traffic density can be provided at insignificant complication in contrast to the low rate in the radio channel and the low traffic in the macro-cells, which is required to provide communication at high mobility. The freedom of location and the access approach are provided with the help of smart cards allowing subscribers to register different terminals with different possibilities (speech, multimedia, digital flow, SMS).

Perfection of the GSM system by implementation of EDGE technology into the technology of radio access network (GERAN) allowed approaching of the GSM network parameters to requirements of the 3G network of the UMTS standard.

The EDGE system affects the unification of GSM, DAMPS, and WCDMA systems owing to the utilization of terminal operating in two standards. Operators, who have a license for operation in the 2 GHz range, have an opportunity to introduce in exploitation the wideband UMTS system in regions where there was the need, for instance, to broaden the service zone. Double-standard mobile terminals (EDGE/UMTS) allow provision of full roaming and handoff from one system to another, in accordance with the schedule of interaction between systems. Although GPRS and EDGE require new functional possibilities of the GSM network for operation with new types of connected networks, which operate with the data packets, they essentially broaden the network possibilities of the GSM standard. Similarly, application of the GSM/UMTS network using two standards as the base network will lead in the future to its broadening.

In spite of the fact that the aim of 3G system creation corresponds to the principle “everywhere, in any time,” cellular networks can cover rather restricted regions, which have a high cost of infrastructure. For this reason, the satellite systems can be the component part of 3G networks. The satellite will provide the widening of cover by wireless communication of the remote regions and air and naval objects.

The cdma2000 System The variant of 3G system construction was suggested in the 3GPP2 project on the base of American radio interface of the IS-95 standard. In this project, developers took as the basis the radio access system IMT-TC. The technology of the digital advanced mobile phone system (DAMPS) based on the UWS-136 standard together with cdmaOne was the base for creation of open networks of wideband access (Fig. 2.22).

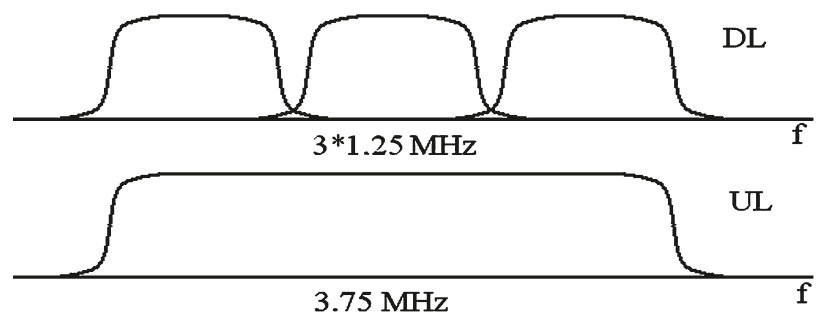
The first stage of development of the IS-95 standard was the creation of cdma2000 1x system, which permitted to double the system capacity and to increase the data transmission rate up to 614 kbit/s. The cdma2000 1x-EV system using the high-rate technology HDR provides the data transmission rate of 2.4 Mbit/s. The rate growth in the last system version is provided by application of data packet transmission owing to allocated carrier for this purpose. The last version of modernized cdma2000 system called cdma2000 3x, anticipates utilization of three independent and non-overlapping CDMA channels at compatibility saving with CDMA2000 and IS-95 systems. Three times increase of the bandwidth (Fig. 2.22) and elimination of some restrictions allow increase of the rendering services and the transmission rate of up to 2 Mbit/s.

The base station provides the data transfer in downlink in three attended spectrum bands, while in uplink transmission is possible in all allocated band. The cdma2000 system allows creation of any dimension cells – from megacells (more than 35 km in radius) to pico-cells (the radius less than 50 m) – for operation in wireless access networks. Procedures of digital sequence formation and interfaces of different levels do not practically differ from that used in the IS-95 systems.

The cdma2000 system uses the air interface (A-interface) basing on the broadened IS-95 standard, which provides a quality of wired communication systems for speech signals at the transmission rate of 144 kbit/s for mobile subscribers, 384 kbit/s for pedestrians, and 2 Mbit/c for stationary subscribers. Data transmission with the rate of 64 kbit/s in the CDMA IS-95B system provides the high-rate access in Internet in the mobile mode excluding the parallel application of other narrowband digital technology.

Application of the feedback for fast estimation of the channel parameters in the 3G 1X EV-DO system of the cdma2000 standard, diversity reception, and the algorithm of urgent control of receiver operation at diversity technique, as well as adaptive phase modulation of 8-PSK type and the 16-position quadratic amplitude modulation (16-QAM), permit to transmit the packet data with rate higher than 2.4 Mbit/s. The system variety of the cdma2000 standard is the synchronous system with the time channel division (TD-SCDMA). Application of adaptive quadrature modulation with the phase shift (QPSK) or 8-PSK allows (together with turbo-coding) providing the rate 2 Mbit/s in the information downlink flow. In the TD-SCDMA system, the time channel duplexing (TDD) is realized at the operating bandwidth of 1.6 MHz, while the cdma2000 system realizes the frequency duplexing (FDD) with occupied frequency band of 2.5 MHz (1.25 MHz for each uplink and downlink channels). The TD-SCDMA system allows usage of the least frequency band among all 3G technologies.

Fig. 2.22 The technology of the digital advanced mobile phone system



Radio Access Networks of the 3GPP Standard In the basis of 3G technology of wideband CDMA (WCDMA), which is the successor of the GSM standard, the requirement to achieve the digital flow rate of 2 Mbit/s was laid in networks with the low-mobile subscribers and 384 kbit/s for mobile subscribers. This technology is the base of the mobile segment of European universal mobile system UMTS with the radio interface of ground-based access UMTS terrestrial radio access **UTRA**, which uses frequency (UTRA FDD) or time (UTRA TDD) duplexing. Wideband systems with the code channel division and the frequency duplexing are similar in many respects with the cdmaOne, but in contrast to the last, they can operate both in the synchronous and in asynchronous modes.

The technologies for speech and data transmission of the GSM standard (Fig. 2.21), which exist at the time of the beginning of the initial work under the project, had formed the basis of the UMTS standard. The most claimed from them were HSCSD, GPRS, and EDGE.

The main UMTS difference from GSM/GPRS networks consists in the interface of wireless transmission. To organize the new access method to the radio interface, the new network of wireless access UMTS terrestrial radio access network UTRAN was required, which, in turn, led to controller replacement in the radio access network RNC and mounting of new base stations (Node B).

The W-CDMA/UMTS network was developed as the replacement of GSM networks; therefore, its network infrastructure is compatible with MAP/GSM. Besides, such a network oriented on creation of global networks should provide modes with packet switching (protocols IP, X.25). This permitted to place UMTS into operation in regions with the dense traffic, gradually widening networks; therefore, terminals of W-CDMA standard in Europe support GSM. The radio signal is formed using the procedure of spectrum spreading by the method of direct sequence providing the chip rate of 3.84 Mchip/s in the channel with width of 5 MHz.

Implementation of the GERAN (GSM/EDGE Radio Access Network) technology in GSM and UMTS networks basing on the UTRAN technology provided their interaction and became determining in development of 3G network (Fig. 2.23).

Why are the high-rate GPRS and EDGE modes with packet switching used in the GSM system in the base network core network (CN)? In the base network of UMTS system, the technology of asynchronous data transmission is used for the creation of the transport network between elements of the radio access UTRAN. At that, the transmission rate for the single speech channel of the base network domain with channel switching (Chanel Switch domain) in UMTS was 64 kbit/s, and the domain with packet switching (Packet Switch domain) was constructed on technology of IP tunneling. The technology based on the IP was used not only for data transmission but also for speech, video, etc. Implementation of novel technologies of speech transmission over the IP – voice over IP (VoIP) – led to transformation of the UMTS system into the system, which is fully integrated into the IP network and permits co-existence of two different networks constructed on the base of unified base network with channel and packet switching.

The architecture of UMTS system (Release 7) reflects the interaction with the mobile station (MS), the radio interface, and the interface of the radio access network and the core network, as well as rendering services of the mobile communication network.

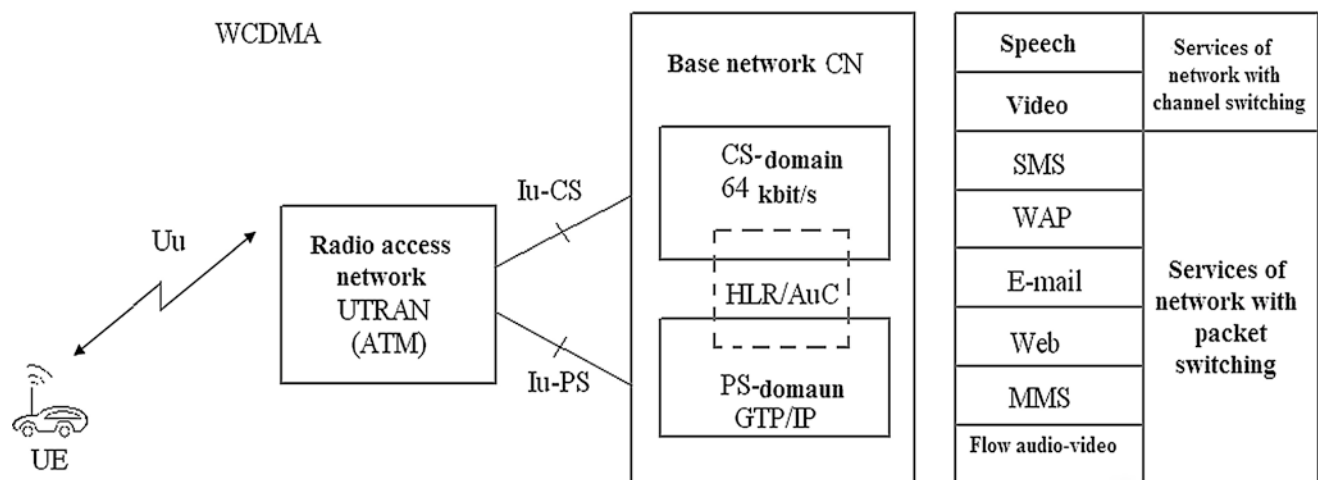


Fig. 2.23 Development of 3G networks

HLR – **H**ome **L**ocation **R**egister is the data base of home subscriber location,
AuC – **A**uthentication **C**enter is the authentication system and the access control.
GTP – **G**PRS **T**unneling **P**rotocol is the technology of IP tunneling.

Development on the base of the WCDMA technology of high-speed access called (**HSDPA**) for information transmission in downlink allowed increase of the data transmission rate up to 12 Mbit/s. The mode of high-speed data uplink transmission (**HSUPA**) and introduction of IP multimedia sub-system (**IMS**) domain defined new technological possibilities for the UMTS system:

- The high-rate data transmission mode on the base of TDD (7.68 Mbit/s).
- Application of multi-beam adaptive antennas based on the **MIMO** technology, which increases the capacity of the radio sub-system.
- Different multimedia applications.
- Providing of the access to the IMS sub-system of fixed wideband networks for data transmission, etc.

The mobile communication networks of third-generation international mobile telecommunication/universal mobile telecommunications system (**IMT/UMTS**) of the European standard have got the widest distribution in the Russian territory. The main difference of such networks consists in utilization of dynamic control of the energetic rate relations, which realize an optimality of the system energy resource in the *ground-based radio access networks* UTRAN of the UMTS system. UMTS networks (up to version 5 Release) provide the peak rate of data transmission of 2.048 Mbit/s. The signal base increase due to introduction of frequency redundancy gives to the radio signal a high noise immunity and weakens the influence of the multipath propagation.

The refusal from the asynchronous data transmission technology (ATM) for channel switching, and the possibility of multimedia service rendering in real-time mode on the base of IP technologies owing to one additional PS domain, allowed changing of the core network.

Necessity in growth of the data transmission rate led to development and implementation of the high-rate packet data exchange in the UMTS network on the base of multi-position amplitude modulation (16QAM, 64QAM), which, in turn, increases requirements to improvement of the access protocol to the physical media MAC.

The radio access system of the UMTS network in two duplex modes (FDD and TDD) provides the information transmission between subscribers in two (in pairs) or in single (non-paired) frequency ranges. In the TDD mode, UTRA operates in three rate modes: 7.68 Mchip/s, 3.84 Mchip/s, and 1.28 Mchip/s. In TDD mode, we can simultaneously use the mode of sequence spreading with the code channel division and the mode of multiple access with the time channel division (TDMA/CDMA).

The Universal System of Mobile Communication UMTS (WCDMA) Code division multiplexing direct spread spectrum (DS-SS) mobile systems are allocated “up/down” bandwidth for UMTS at 5 MHz Frequency Duplex Division (FDD), and for Time Duplex Division (TDD), the radio signal rate is 3.84 Mchip/s. For narrowband CDMA 2000 system and FDD, radio channel width is 1.6 MHz, and for TDD 1.28 Mchip/s. It is possible to apply the band of 10 MHz for the mode 7.68 Mchip/s. The overwhelming majority of WCDMA networks use FDD with the band of 5 MHz.

The access method to the transmission medium in WCDMA represents a combination of methods TDMA and CDMA. Simultaneously, several physical channels, concerning channels of the transport level, can be transmitted. Each physical channel has its own channel code and coefficient of spreading. Signal transmission is performed by radio frames with duration of 10 ms. The *radio frame* is the separate time interval, in which there is the codes’ (time slots) conferment to the physical channel, and the method of encoding/decoding is specified, and parameters of spread sequences are defined, etc. Depending on the chosen duplex approach and the multi-station access method, the WCDMA system can be characterized by a series of constant values: the rate of chip transmission and the number of channel intervals (time slots) during one radio frame, the number of chips in the time slot, and the value of the spread coefficient.

The procedure of signal processing in the WCDMA system can be represented by the structural scheme in Fig. 2.24.

The speech analog signal passes on the source encoder – the adaptive multi-rate codec AMR – which forms the digital stream (bits) with eight different rates, in which a specific value is chosen (Fig. 2.23) by the radio access network (RAN), depending on the network loading and quality indices of the signal source. The video signal, the signal of audio accompaniment, data, etc. are processed by the codec H 324 in the block of channel switching (CS domain).

The physical channel is defined by the carrier frequency used by the spread sequence and by signal component (in the FDD mode) or by the frame time slot (in the TDD mode). Two types of allocated channels are defined on the physical level: the control channel and the data channel. The physical control channel and several data channels are allocated by each connection (downlink). In the uplink direction, the data transmission is organized through the transmitter quadratic input and the

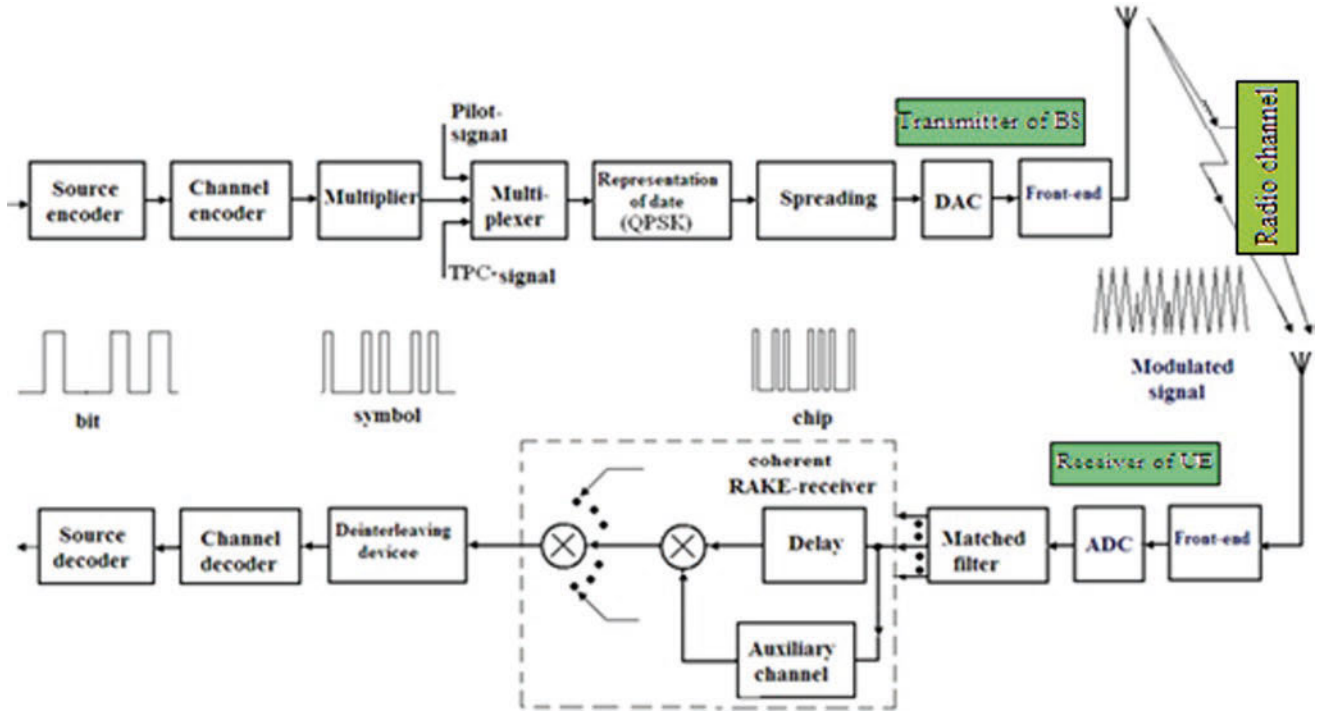


Fig. 2.24 The procedure of signal processing in the WCDMA system

control channel through the in-phase input. Both digital streams (data and control) are dispersed by two mutually orthogonal codes of channel formation (**Channelization Codes**).

Cyclic Redundancy Check (CRC) channel coding increases the correlation between symbols, allowing signal recovery after interference. However, the signal correction at reception leads to the signal delay growth. For speech signals, the convolution codes with the coding coefficient 1/2, 1/3 are used, when, relatively, 16 or 24 bits are added to the input sequence. Besides, the turbo-coding with the coefficient 1/3 with addition of 12 control bits is applied. The increase by two or three times of the sequence length on the channel encoder output compared to the input length decreases the time interval, defined for the single transmission (symbol), and increases the probability of correct reception of the informational bit.

The interleaving procedure reduces the error arising probability in the following bits (one after another). It is possible to use three intervals of interleaving: 20, 40, and 80 ms. After coding and interleaving in the downlink direction, each frame of the digital sequence is converted in the symbol QPSK sequence, and then (each 0.667 ms) multiplexing is fulfilled with the pilot signal and the signal controlling the transmitter power. The QPSK symbol sequence is formed from the digital sequence without return to zero by means of its division into synchronous and quadratic components with further scrambling of each component.

Coding and interleaving of the informational sequence allows obtaining 4...256 times bandwidth increase, which is occupied by the transformed signal spectrum depending on the spreading factor (SF).

Quantitatively, SF (the signal base) is defined as a ratio of signal spectrum bandwidth Δf_s and the message spectrum bandwidth ΔF_s (a width of the initial signal spectrum):

$$B = \Delta f_s / \Delta F_s$$

If transmission of the one information bit is performed by the signal with the duration $T = 3.69 \mu s$ in the network of the GSM standard, then $B = 1$. For the signal with the GMSK modulation, $\Delta f_s \approx 1/T$, and the message spectrum bandwidth $\Delta F_s = 200 \text{ kHz}$ is defined by the transmission rate of informational bits $R = \Delta F_s = 1/T_b$, where T_b is the transmission duration of the informational bit. Since the duration of the one informational bit coincides with the signal duration, which transmit the bit $T = T_b = T_o$, then $B = \Delta f_s / \Delta F_s = \Delta f_s T_b$.

In wideband networks, the signal base is significantly more than 1, since for signal spreading we apply the video-pulse fragmentation of the T duration into L elements with duration $\tau_{el} = T/L$, which increases the signal base by L times

$$B = \Delta f_s / \Delta F_s = T / \tau_{el} = \nu / R = L$$

where ν is the chip rate and R is the data transmission rate. Reception of the signal with the bandwidth Δf_s on the background of wideband (for instance, the white Gaussian noise) interference ($\Delta f_{int} \gg \Delta f_s$) has noise immunity, which is defined by a ratio of the radio signal energy E_s and the noise (interference) spectral density N .

At action of the interference with the spectrum width, which is comparable to the useful signal spectrum ($\Delta f_{int} = \Delta f_s$) and has the limited power P_{int} , the spectral density of the interference power is defined as $N/2 = P_{int}/\Delta f_{int} = P_{int}/\Delta f_s$, and SNR at correlation receiver input is

$$SNR = \frac{2E_s}{N} = \frac{P_s T}{P_{int} / \Delta f_s} = \frac{P_s \Delta f_s}{P_{int} \Delta f_s} = \frac{P_s}{P_{int}} B.$$

From this it follows that SNR in the wideband network with the radio access increases compared to narrowband networks by the signal base times.

For noise immunity increase of mobile communication systems with the WCDMA technology, following the speech coding, the double spectrum spreading is used: at first, the channel coding by the orthogonal code and then the coding by the pseudo-random code. The channel coding is intended for subscriber signal division depending on the chosen services and is based on the subscriber data multiplication by the one from the possible orthogonal combination of the Walsh code of different length.

PRS coding is intended for division simplification of the useful and spurious signals of different sources, which operate in the mutual frequency range (subscriber terminals, base stations). Useful signal extraction is based on comparison of received marker sequence with the etalon code combination known of the reception side. Reception of outside PRS leads to appearance at receiver output of the noise signal caused mainly by interference noise. To simplify the decision-making procedure on the base of the maximal likelihood algorithm, we can form the introducing spreading sequence in such a manner that provides the equal probability of different elementary symbol appearance (the transmitted sequence is exposed by scrambling). In the downlink, orthogonal codes are used for separation of subscriber data, and PRS is used for separation of cells (transmitters of base stations). At the receiving side, the sequence of fulfilled transformations is inverse: at first, descrambling is performed and then multiplication of the digital sequence by the orthogonal code (Fig. 2.24).

The spreading procedure is performed in two stages: the operation of the band division into separate channels, at which the pulse sequence is transformed into the sequence of chips, and scrambling. The division of the frequency band into separate channels is performed with utilization of the orthogonal OVSF code with the variable spreading coefficient, which permits in the mutual front-end band to form the signal having the distinguishing signs belonging to the specific subscriber. This is implemented by application of pseudo-random spread orthogonal digital sequences (for instance, Walsh functions), when each bit of the data stream is represented by the full period of the Walsh sequence with the length of 64 symbols. The example of orthogonal sequence formation with different spreading factor is presented in Fig. 2.25.

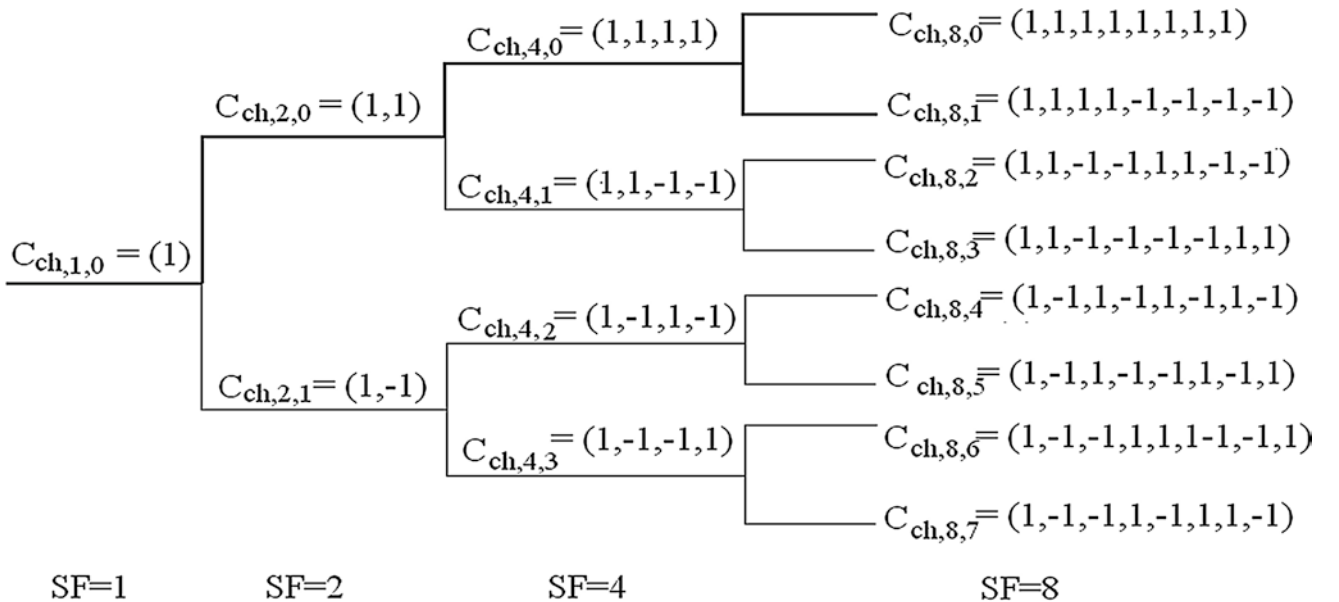


Fig. 2.25 The example of orthogonal sequence formation with different spreading factor

This short spreading code has the length of $4 \dots 512$ chips, the combination length of which is defined by the spreading coefficient. New coding words on the code tree are created by means of previous word addition by itself (in the upper branch outgoing from the node) or by its own negation (in lower branch). To provide the sequence orthogonality, which are chosen for dispersion of specific data stream, for example, $C_{ch,8,5}$ with the spreading coefficient $SF=8$, we can take any word besides $C_{ch,4,2}$ for another data stream, which requires $SF=4$. In the downlink, the channelizing code (OVSF code) is used for identification of physical channels inside of the cell, and in uplink direction, the physical channels separation of the one mobile station is used.

The goal of scrambling in the downlink direction is to identify different cells, while in the uplink direction it is to identify subscribers using the same carrier frequency at keeping unchanged allocated frequency bandwidth. Since the unique scrambling code is conferred to each cell (for downlink direction), WCDMA permits the asynchronous network operation, which facilitates its development. In turn, application of different scrambling codes in the different base stations increases the time for cell search.

Transmission organization in downlink direction should provide the synchronization of the mobile station with the base station, which is ensured by the scrambling sequence, which is transmitted by the base station. All 512 scrambling sequences divided into 16 groups with 32 sequences in each are distributed among cells at network planning. After activation, the mobile station receives (via the synchronization channel) the signals from several base stations with the help of a filter, which is matched with the known sequence, and chooses the nearest station according to the local maximum. On the second stage of the synchronism entering, the mobile station performs the processing of 17 possible Gold codes with the length of 256 chips each with the help of 17 correlators and records 32 possible PRS combinations in the frame and 16 cyclic shifts into the memory. After that, the mobile station determines the pair “Gold code/cyclic shift” having the largest correlation coefficient. At the final stage of establishing synchronization, one of the 16 sequences of main group is identified, which indicates the set time offset and allows the mobile station to determine its scrambling sequence, what makes it possible to connect to the control channel.

Noise immunity in the UMTS systems is provided by application of the multi-subscriber detection and specific systems suppressing the intra-system interference. In the first case, the signal processing of all active subscribers is performed. This allows utilization of all disposed statistics for decision-making concerning the received informational transmission. This is especially important for the asynchronous system (WCDMA), when due to random shifts between transmitted informational sequences, it is necessary to detect all received sequences from all subscribers. In this case, the sequence that has the largest a posteriori probability is considered as received after processing of the grouped WCDMA multi-subscriber signal. To make a decision about the receiving sequence, we can use the Viterbi algorithm, the complexity of which is estimated at transmission of the binary signal by the value of 2^K , where K is the number of active subscribers. Evidently, this algorithm for the real participator number requires the large volume of computing operations or simplification of the processing procedure, which accordingly leads to increase of the error probability.

Utilization of the method based on the interference estimation, which is introduced by the active subscriber, and its subtraction from the grouped multi-subscriber signal is the alternative. There are two variants of this method realization: in the form of parallel and in series circuits. Suppression of the intra-system interference occurs after correlation signal processing by the receiver from the current subscriber and after elimination of received signal from the received multi-subscriber signal. After signal processing from the current subscriber, the number of components in the grouped signal decreases, which increases the reliability of the final result.

In practice, to increase the struggle effectiveness with system interference, at first, the component ranking of received grouped signal in power is performed with the further suppression of the most powerful from them, which has the highest SNR. It is known that the pilot signal has the largest input power; therefore, we can achieve the significant increase of the true reception probability of the useful signal by its even partial suppression.

Application of circuits for the struggle with intra-system interferences requires more complicate algorithms of received signal processing compared to the conventional correlation receiver.

2.5.3 Wireless Systems of Fourth Generation

2.5.3.1 General LTE Network Structure

Technologies that allow data transmission in the cellular communication systems with rate above 100 Mbit/s are attributed to the 4G family. For the most worldwide widespread GSM/EDGE standard of the cellular communication, the data transmission rate is 240 kbit/s. In networks of the third generation (3G) deployed in Europe, the USA, and Asian countries (Taiwan, Japan, Singapore), they provide the transmission of the speech traffic and data with the rate of $7 \dots 14$ kbit/s.

4G systems are based on protocols of packet data transmission, which distinguishes it from the earlier generations. The example of the 4G technologies is the WiFi and WiMAX systems, which have the theoretical rate limit of 1 Gbit/s. Creation of the competitive technology of the mobile communication construction on the base of the mobile communication network WiMAX (IEEE 802.16e standard) activated efforts of 3GPP project participants on development on the base OFDM technology of the evolution variant of the UMTS network called **LTE (Long-Term Evolution)**.

The International Telecommunication Union defines 4G technology as wireless communication technology, which permits to achieve data transmission rate of up to 1 Gbit/s under conditions of source of receiver motion and up to 100 Mbit/s under conditions of data exchange between two mobile devices. 4G networks can be attributed to wireless networks of the arbitrary structure with the episodic access between independent nodes (ad hoc peer-to-peer networking) with high readiness to exploitation and global roaming, distributed computer structure, personalization, and multimedia support. 4G networks use the distributed architecture and the interface close to the Internet Protocol (IP). Each device will simultaneously be the transceiver, but for another device the switcher-router, which excludes the blocking concentrator (the shortcoming of the 3G cellular system architecture). The network capacity and covering will be dynamically varied in accordance with subscriber characteristics. Subscribers will be automatically excluded from the overloaded route permitting automatic network self-balancing.

At the time of 2010, LTE data rates were 326 Mbit/s (downlink) and 86 Mbit/s (uplink) [33] in the 20 MHz band. For data transmission, **IPv6 (Internet Protocol version 6)** is used, which greatly facilitates the operation of the network, especially when interacting with various types of systems.

The wideband access system LTE is adopted for operation within the limits of application of prospecting radio technologies by the Russian Government order №57-p dated 27 January 2011. Services rendering in the range 791–960 MHz are planned to provide in 2015, in the range 2300–2400 MHz, as in range 2570–2620 MHz and in 2500–2570/2620–2690 MHz in 2014. In the frequency ranges 2300–2400 MHz, 2570–2620 MHz, and 2500–2570/26020–2690 MHz is possible to deploy WiMAX network in the same years. At present, in Sankt-Peterburg 3.5G network of UMTS/HSPDA standard provides data transmission with rate up to 14 Mbit/s.

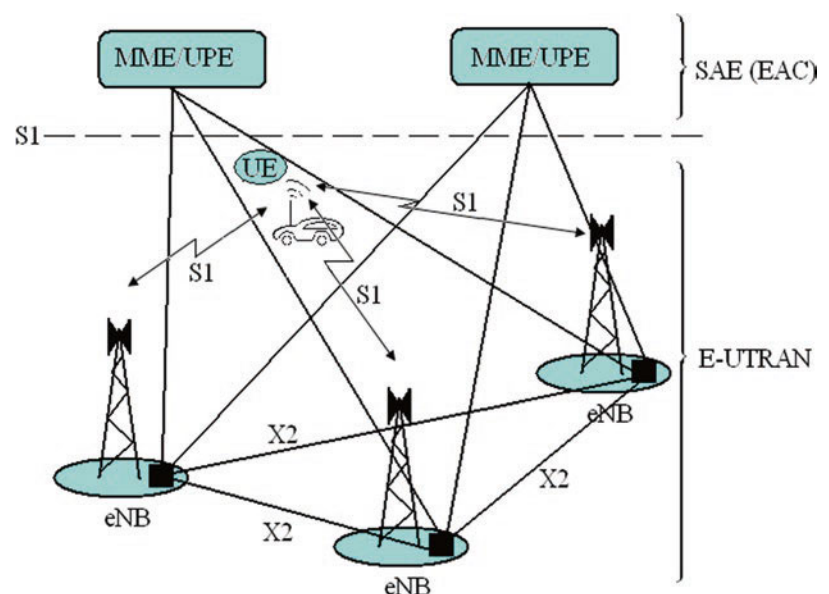
Improved technologies of the multi-station radio access, application of effective types of modulation, and signal formation on the technology with orthogonal frequency shift keying OFDM led to development of the new version of the UMTS network. This network (Fig. 2.26) included the radio access network E-UTRAN, and the base network system architecture evolution (**SAE**) has got the name (Release 8) of perfected system of multi-station access LTE.

Effective application of new OFDMA technology in networks of wireless access WiFi/WiMAX allowed development of the new radio interface for cellular 4G systems using the 3G interface family.

This interface called IMT Advanced (with OFDM technology) was created within the frames of investigations in the partner project on 3G systems (3GPP). The International Telecommunication Union developed the *standard family* of radio access **IMT-2000 (International Mobile Telephone System)** for both segments of 3G system – satellite and ground-based – operating in the 2GHz range.

LTE is based on application of three actively developed technologies: multiplexing of orthogonal carriers (OFDM), multi-beam antenna systems (MIMO, multi-input multi-output), and evolutionary architecture of the base network system architec-

Fig. 2.26 The structure of the radio access network E-UTRAN and its interactions with the base SAE networks



ture evolution (SAE). This chapter is limited to considering the network structure of this system and issues that reflect the features of the OFDM technology and determines the structure of the radio channel terminal devices. The structure of the LTE network (Fig. 2.26) limited by the physical level includes the radio access network E-UTRAN, which consists of interacting base stations (eNB) through the interface X2 and the base network SAE и абонентские устройства (UE) through interface S1.

The structure of the radio access network E-UTRAN is organized on the principle “each to each” and consists only of base stations (eNB) and uses the S1 interface with the base station SAE built on the principle of packet switching.

The base stations eNB fulfill the functions of radio resource control RRM (radio resource management). This increases the effectiveness of the frequency spectrum utilization and minimizes the mutual influence of subscriber terminals (UE) and provides the UE relay transfer (roaming).

The function of data transmission control in the radio channel RBC (radio bearer control) is laid on the radio access network E-UTRAN, which provides setting, supporting, and releasing of radio channels for data transmission according to requirements of parameters of quality of service (QoS) control.

Received testing signals contain information on decoding of uplink signals and serves for estimation of the uplink signals on the base station. The reference signals, which are transmitted through five sub-carriers with the offset defined by the cell indicator and by pseudo-random frequency retune, allow the radio access network to choose modulation types and speed and the frequency-time resource for each user. The probing signal (sounding reference signal), which does not carry information, is transmitted in the last symbol of the sub-frame in different parts of the channel band and serves for estimation of quality of uplink channels on eNB.

Signals transmitted through the physical channels between eNB and UE can be divided into two groups: housekeeping (control) and for data transmission. On the physical level (Fig. 1.10), in downlink, we must form the housekeeping channels: (1) wideband PBCH (Physical Broadcast Channel), which transmits information (cell-specific) about UE number (registered in the network); (2) the channel, which informs about the resource allocation, i.e., PDCCH (Physical Downlink Control Channel), which transmits information about numbers of OFDMA symbols; and (3) the channel transmitted useful data, PCFICH (Physical Control Format Indicator Channel). The indicative channel serves to confirm data delivering in the uplink channel PHICH (Physical Hybrid ARQ Indicator Channel).

To data transmission the following channels are allocated: the common for data transmission PDSCH (Physical Downlink Shared Channel), the group channel for transmission PMCH (Physical Multicast Channel).

Channels of higher level, transport channels (Fig. 1.10), are intended for housekeeping information transmission, which is transformed (represents) into the physical level and is transmitted between eNB and UE. For example, in downlink, the state indicating channel for the transmission repeat **HI** (HARQ indicator) forms the request to the transmission repeat hybrid automatic repeat request (**HARQ**) depending on its indications. The transport channel downlink control information (**DCI**) forms the sequence about channel state in downlink. Other transport channels perform similar functions (check of physical channel state in the system, an analysis of higher-level states, and formation of commands in accordance with obtained results).

The E-UTRAN with radio access consists of base stations **eNB** (evolved **Node B**) connected according the principle “each to each” with the help of X2 interface, which support the handover of the subscriber terminal. Each base station has the S1 interface with the base station of the SAE network constructed on the packet switching principle. The basing network contains MME/UPE nodes consisting of logic element mobility management entity (**MME**) and user plane entity (**UPE**). The MME node is responsible for mobility control of the subscriber terminal and its interaction with base stations with the protocols of the control plane (S1-C interface). The logic element UPE is responsible for subscriber data transmission according to protocols of the subscriber plane (U-plane) and interaction with eNB through the S1-U interface. The conception of “all through IP” is laid in the basis of the base network construction as well as the fact that the access to the base network SAE can be fulfilled both through the radio access 2G and 3G networks (UTRAN, GERAN) and through networks of non-European radio access (networks, which are not standardized by the 3GPP projects), for example, networks of the IEEE standard – WiFi and WiMAX – and also networks using wired IP technologies (ADSL+, FTTH etc.). This procedure of matching with networks of another standard is performed by the sub-system IP multimedia sub-system (**IMS**), which together with the gateway node of affixment to 3GPP Anchor and SAE Anchor form the unified affixment node **IASA** (inter-access system anchor) for connection of IP networks.

Technical Indicators of the LTE/LTE-A and WiMAX Systems Application the promising OFDMA technology of multi-station access with orthogonal frequency shift keying instead of usual CDMA technology in UMTS network, and also multi-beam adaptive antennas, which realizing the MIMO principle, allows increase of the downlink capacity up to 100 Mbit/s at bandwidth of the single frequency channel up to 20 MHz (with a possibility of scaling to 1.4; 3; 5; 10; 15 MHz) and up to 50 Mbit/s in uplink.

Requirements to designers caused by the necessity of functional possibility improvement, introduction of new services, and widening of exploitation possibilities are included in modernized variant of the LTE Advanced (LTE-A) system 4G standard taking into consideration new scenarios of LTE network development:

- The maximal rate of data transmission in downlink is up to 1 Gbit/s, while in uplink, it is up to 500 Mbit/s.
- The bandwidth in downlink is 100 MHz, while in uplink it is 60 MHz.
- The maximal effectiveness of spectrum utilization in downlink is 30 bit/s/Hz, while in uplink it is 15 bit/s/Hz.
- Complete compatibility with other LTE systems and 3GPP standards.

Development of the element base on the basis of new technologies and materials and also sufficiently a wide circle of manufacturers, who produce multi-functional integrated circuits, led to a variety of producing equipment, which corresponds to standards different from 3GPP/ESTI (European Telecommunications Standards Institute), for instance, 3GPP2 and IEEE (Fig. 2.21).

This supposes the radio channel presence with bandwidth up to 100 MHz, asymmetric separation of bandwidths between downlinks and uplinks in the case of frequency duplexing, the more perfected coding systems and error correction, the hybrid OFDMA and SC-FDMA technologies for uplink, and MIMO technology for LTE antenna systems. This will permit to essentially increase the network capacity in downlink up to 100 Mbit/s at frequency bandwidth of 20 MHz and to increase its effectiveness up to 5 (bit/s)/Hz. For the inverse channel (uplink) for the same bandwidth and capacity of 5 MHz, the channel effectiveness increases to 2.5 (bit/s)/Hz. Organization of the high-rate data transmission using the HSDPA technology of high-rate packet access in downlink can give the increase the spectral effectiveness of the LTE system by almost three times compared to the WCDMA/UMTS system.

Implementation in Russia of the cellular communication system of the LTE/LTE Advanced standard of 4G is defined by possibilities of frequency utilization allocated by ITU and existence of their occupation by different agencies. From the offered list of radio-frequency ranges – 694–915 MHz, 925–960 MHz, 1710–1880 MHz, 1900–1980 MHz, 2010–2025 MHz, 2110–2170 MHz, and 2500–2700 MHz – the winners of the competition for the obtaining of licenses are the following Russian Companies:

“Rostelecom” – frequencies in the range 791–798.5/832–839.5 MHz.

“MTS” – frequencies in the range 798.5–806/839.5–847 MHz.

“Megaphone” – frequencies in the range 806–813.5/847–854.5 MHz.

“VimpelCom” – frequencies in the range 813.5–821/832–839.5 MHz.

Development of the element base on the basis of new technologies and materials and also sufficiently a wide circle of manufacturers, who produce multi-functional integrated circuits, led to a variety of producing equipment, which corresponds to standards different from 3GPP/ESTI (European Telecommunications Standards Institute), for instance, 3GPP2 and IEEE.

Evolution transfer from 3G systems with the CDMA technology to the OFDMA technology in the downlink channel and to SC-OFDMA in the uplink in the LTE Advanced 4G system facilitates the indices' approaching of LTE and WiMAX (IEEE 802.16m standard) systems. The main difference of the LTE system compared, for example, with WCDMA system consists in application of principally another radio interface with the technology of the orthogonal multi-station access with frequency division (OFDMA) for the direct channel and the multi-station access with the frequency channel division on the single carrier frequency **SC-FDMA** for inverse channel. The global trend of mastering the LTE/LTE-Advanced in Russia has led to the fact that the company “Skartel” (trade mark “YOTA”) is transferring all its subscribers from the WiMAX standard to the LTE standard. This standard, which was adopted by the 3GPP partnership, includes the set of services of the wideband radio access, which, in the near-term outlook, will provide subscribers with the necessary volume and the character of requested information and the required data transmission rate.

The OFDM and DS-CDMA combination is already used in MC-DS-CDMA systems with many carriers, which combines the best properties of OFDM and DS-CDMA and, hence, ensures the good operation indices under conditions of multipath reception. At application of the MC-DS-CDMA architecture, at first, the digital stream is divided into sub-streams. Each separate data stream is spread on action durations of the spread code sequence and then transmitted by one of the orthogonal sub-carriers. This allows increasing of the spectral effectiveness and the length of the spreading code; nevertheless, different formed streams interacted with each other. The transmitter of MC-DS-CDMA system distributes the origin digital sequence on the orthogonal sub-carriers (chosen earlier) using the spreading code for the specific frequency range. Systems, which apply MC-DS-CDMA, can provide a very high average capacity. In order to further increase the system spectral effectiveness, we should use some forms of adaptive modulation.

4G systems of wireless communication should have a possibility to provide transmission on sub-carrier of different signal types such as a speech, digital data, an image, and video signals as well as to have the compatibility with systems of other standards. The new generation is the development of the 3G standard using the technology of direct spectrum spreading DS-CDMA, which shows a high effectiveness for speech message transmission. Application of the new OFDM technology in the 4G standard allows transmission rate increase and effectively fights with the multipath propagation and simultaneously to be the first step to equalization of output signal parameters in the frequency domain with utilization of the technology of dynamic spectrum spreading. In aspects, their combination can significantly simplify the system integration, which have different technologies, and realize the idea of the global communication network. Three variants of configuration are known from literature, which are called MC-CDMA, MC-DS-CDMA, and the multi-tone MT-CDMA. In the CDMA compression system with many carriers (MC-CDMA), the spreading code consequently confers a number to the orthogonal sub-carrier in the frequency band of the operating range.

Wireless 4G systems will use the multi-level combination of different technologies of the multi-station access:

- Cellular systems (for instance, existing 2G and 3G systems having a large covering zone).
- Systems developed for application inside premises (WLAN such as IEEE 802.11a, 802.11b, 802.11 g).
- Global compatible systems with UHF access (WiMAX, IEEE 802.16) for intraurban mobile communications (WMAN).
- Systems arranged for operation with low-mobile subscribers, which are located at small removal (WPAN such as Bluetooth, UWB).

As we see from the description of 4G systems of mobile communication WiMAX (IEEE 802.16 m) and LTE 4G systems of mobile communications, there is a variety of coinciding characteristics both in the network structure and in methods of used modulation and used protocols (Table 2.4).

The OFDM technology is a sort of frequency channel division, but in contrast to that used in GSM, the orthogonality must be ensured between sub-carriers. On each of sub-carrier, the slow digital sequences are transmitted, which are formed from the initial sequence. Although that sub-carriers are chosen with offset of 15 kHz only and the spectrum of modulated sub-carriers can be partially overlapped, their orthogonality is provided on the essential part of information symbol duration (T_{ort}) by introduction of the guard interval T_g , and interference touches the small part of information symbols only.

Requirement of orthogonality between signals on each from chosen sub-carrier, which act at the receiver input, defines a choice of the frequency offset between sub-carriers. So, for the mobile communication network LTE, a step between sub-carriers is 15 kHz, which corresponds to the clock frequency of 30.72 MHz, which is multiple to the standard processing frequency of 3.4 Mbit/s ($8 \cdot 3.84 = 30.72$) in the WCDMA system of the third generation with channel bandwidth of 5 MHz.

2.5.3.2 OFDM Technology

Formation of orthogonal sub-carriers. At that, utilization of the packet method of data transfer and a large number of sub-carriers allows detection of the informational symbols of each sub-carrier owing to the correlation processing method of the received signal. Necessary for this a synchronism and in-phase voltages of LO and received signal are achieved by introduc-

Table 2.4 Characteristics of mobile communication 4G systems

System parameters	LTE advanced	WiMAX
Duplexing	FDD and TDD	FDD and TDD
Frequency range for analysis	2000 MHz	2500 MHz
Channel width	Up to 100 MHz	Up to 20 MHz
Access method (downlink)	OFDMA	OFDMA
Access method (uplink)	SC-FDMA	OFDMA
Spectral effectiveness, bit/s/Hz		
Downlink channel, MIMO (2×2)	1.57	1.59
Uplink channel, MIMO (1×2)		
Maximal rate MC, km/hour	350	120
Frame duration, ms	1	5
Antenna systems		
Downlink channel	2×2, 2×4, 4×2, 4×4	2×2, 2×4, 4×2, 4×4
Uplink channel	1×1, 1×2, 2×2, 2×4	1×1, 1×2, 3×2, 2×4

tion into the transmitted spectrum of spectrum components of harmonic signals with the fixed frequency (pilot signals), which are modulated by the pseudo-random sequence.

At that, utilization of the packet method of data transfer and a large number of sub-carriers allows detection of the informational symbols of each sub-carrier owing to the correlation processing method of the received signal. Necessary for this a synchronism and in-phase voltages of LO and received signal are achieved by introduction into the transmitted spectrum of spectrum components of harmonic signals with the fixed frequency (pilot signals), which are modulated by the pseudo-random sequence.

The principle of orthogonal sub-carrier formation is clear from the representation of any periodic function $s(t)$ by the expansion into the Fourier series over trigonometric functions [1]. As the orthogonal system we can take:

$$1, \cos \omega_1 t, \sin \omega_1 t, \cos 2\omega_1 t, \sin 2\omega_1 t, \dots, \cos n\omega_1 t, \sin n\omega_1 t, \dots, \quad (2.7)$$

or

$$\dots, e^{-i2\omega_1 t}, e^{-i\omega_1 t}, 1, e^{i\omega_1 t}, e^{i2\omega_1 t}, \dots, \quad (2.8)$$

where $T_{ort} = 2\pi/\omega_1$ is the interval of orthogonality, which coincides with the repetition period of the $s(t)$ function. When we use the orthogonal system (2.8), the Fourier series takes the form:

$$s(t) = \dots + c_{-2}e^{-i2\omega_1 t} + c_{-1}e^{-i\omega_1 t} + c_0 + c_1e^{i\omega_1 t} + c_2e^{i2\omega_1 t} + \dots + \sum_{n=-\infty}^{\infty} c_n e^{in\omega_1 t} \quad (2.9)$$

where

$$c_n = \frac{1}{T} \int_{-T/2}^{T/2} s(t) e^{-in\omega_1 t} dt. \quad (2.10)$$

Formula (2.10) takes into consideration the existence of the complex-conjugated function $e^{-in\omega_1 t}$ for the expression $e^{in\omega_1 t}$.

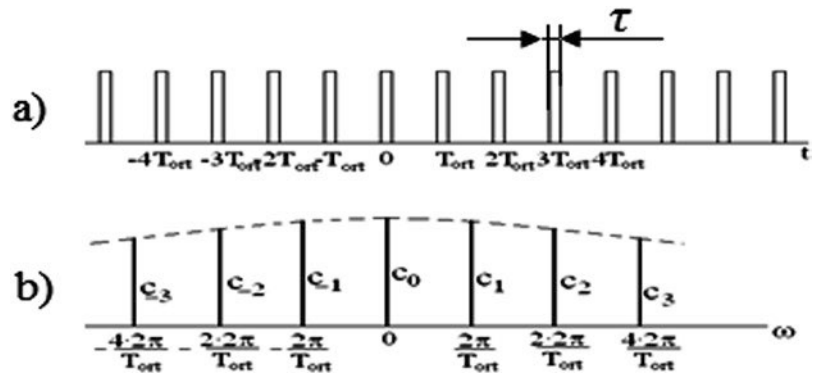
Chosen as the periodic signal $s(t)$, the sequence of the rectangular pulses, which *repetition period* is equal to the *orthogonality duration of the pulse* (T_{ort}), is presented in Fig. 2.27.

It is easy to calculate that at sub-carrier offset in $\Delta f = 15$ kHz, the pulse duration T_{ort} providing their orthogonality is

$$T_{ort} = 1 / \Delta f = 66.7 \mu s. \quad (2.11)$$

The edge slope of the hatch in this figure, which restricts from the top values of spectral components (c_n , expansion coefficients into the Fourier series), is defined by the pulse duration of the formed sequence. The closer the rectangular pulse form to the δ -function, the lesser the difference of harmonic amplitude. Thus, to obtain the sub-carrier frequency grid with required offsets, it is necessary to form the sequence of rectangular pulses with the required repetition period and then to calculate the Fourier transform. Obtained expansion coefficients (c_k) are the signal amplitude with the frequency $\kappa \cdot \Delta f$ of k -th sub-carrier. Multiplying the k -th sub-carrier by the segment of the slow digital stream formed from the initial fast data stream, we have the compact spectrum occupying the bandwidth not more than 15 kHz with the central frequency $f_{0\kappa} = \kappa \cdot \Delta f$. At total number N of harmonic components (the number of sub-carriers), the discretization interval duration is $T = T_{ort}/N$, which

Fig. 2.27 The sequence of the rectangular pulses (a), which *repetition period* is equal to the *orthogonality duration of the pulse* (T_{ort}), sequence spectrum (b)



defines the duration of the OFDM symbol. In real conditions, the special processor is used for generation of discrete fragments of signal, and this processor realizes the fast Fourier transform (FFT).

Modulated Orthogonal Sub-carriers A number of signal fragments N of the signal accumulated during the orthogonality period T_{orth} allows generation of signal fragments with several carriers with the help on the inverse discrete Fourier transform (IDFT). The IDFT algorithm has got the largest application, which, however, is possible only at N number multiple to the order of 2, i.e., $N = 2^m$. This means that at the receiving side we can easily recover the signal using N fragments of the OFDM symbol with known spectrum parameters on each sub-carrier. The IDFT algorithm has received the greatest application, which is significantly simplified when the number of decomposition harmonics N (number of subcarriers) is multiple of 2, i.e., $N = 2^m$. This means that on the receiving side, we can easily reconstruct the signal using N fragments (Fourier transform points) of OFDM symbol with known spectrum parameters on each subcarrier (Fig. 2.27). Thus, signal recovering with the multi-channel modulation is possible with utilization of the signal processor even at using several hundreds of sub-carriers.

The described procedure for obtaining orthogonal sub-carriers (spectral components of the Fourier series expansion) can be easily replaced by an algorithmic procedure in the DSP. The simplest way to obtain any fundamental frequency harmonic (orthogonal sub-carriers) follows from expression (2.7) – multiplying the argument of the first harmonic of the Fourier series expansion by n (the required number in the expansion).

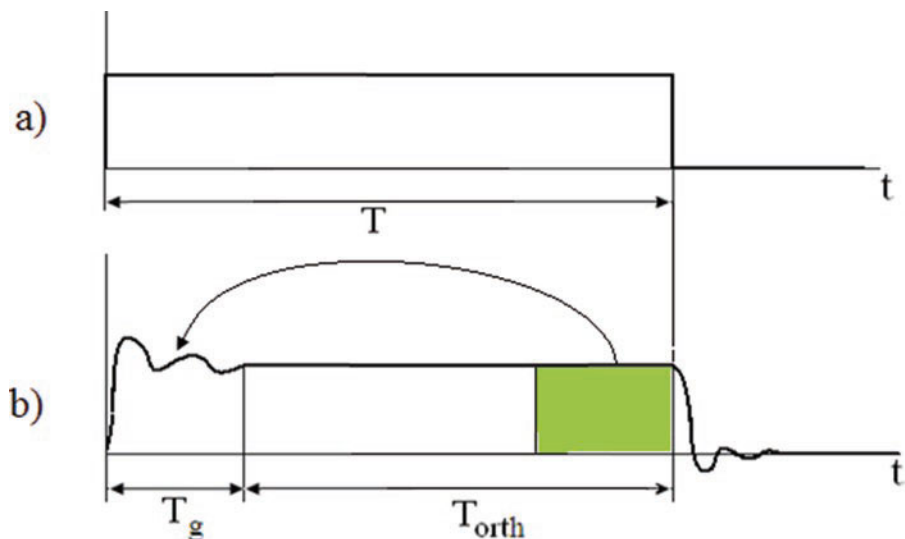
Expressions (2.9) and (2.10) show that the constant amplitudes of c_n sub-carriers and their orthogonality (Fig. 2.27b) are possible only if the sequence is a set of δ -functions ($\tau \rightarrow 0$, Fig. 2.27a). In reality, digital systems use sequences (data, traffic) in the form of rectangular pulses, the duration of which is equal to the following period (without returning to zero, [19]). In the frequency domain, they will be described by a set of discrete spectrum components localized near the frequencies c_n and the envelope of the spectrum of the form $\sin x/x$. When choosing the length of the modulating sequence symbol, the orthogonality of the sub-carriers must be preserved, which is the case when they are spaced in the LTE system at 15 kHz. In accordance with the condition (2.11), the length of the modulating symbol will be $T = 15 \cdot 10^3 = 66.7 \mu\text{s}$, which also provides orthogonality of modulated sub-carriers. Independence (orthogonality) of the component spectra of modulated sub-carriers is ensured by the fact that the envelopes of each of them have zero values at the frequency of any sub-carrier, except for one, on which information is transmitted.

However, modulating the symbol (Fig. 2.28a) in the process of signal transmission by the radio transmitter/ receiver in the radio channel, as well as the correction amplitude frequency features in the high frequency stages of the transmitter, a modulated sub-carrier orthogonality may be lost. This will lead to intersymbol interference that changes the shape of the symbol in the accepted information sequence (Fig. 2.28b).

To preserve the orthogonality of modulated sub-carriers along the entire length of the symbol, the initial part where distortions occur is replaced with the final part of the proper symbol, creating a protective interval T_g . In this case, $T = T_{\text{orth}}$ preserves the orthogonality of envelopes in the time and frequency domains for all modulated sub-carriers of the group signal.

This makes it possible to effectively deal with multipath caused by the action at the receiver input in addition to the line-of-sight signal of its echo copies, which have passed a longer path due to reflections and scattering of the radiated signal. If it is not more than 1.4 km in comparison with the path taken by the main signal, then its delay is no more than $T_g = 4.8 \mu\text{s}$ for the urban LTE network.

Fig. 2.28 The form of the modulating symbol (a), the envelope of the received signal with intersymbol interference (b)



Twelve modulated sub-carriers, each occupying the frequency band $\Delta f = 15$ kHz, are combined on the transmitting side into a radio block, with a common frequency band $\Delta f_{RB} = 15 \cdot 10^3 \cdot 12 = 180$ kHz. In the time domain, this structure forms an OFDM symbol with a duration of $T_{OFDM} = 66.7$ μ s. Each of them is separated from the other OFDM characters by a protective interval – a cyclic prefix that is $T_{CP} = 4.8$ μ s. Distortion of the front caused by fading of the input signal will fall on the cyclic prefix, ensuring that when it is discarded in the receiver, the mutual orthogonality of subcarriers is preserved over the remaining OFDM symbol time.

With the length of the cyclic prefix $T_{CP} = 16.7$ μ s, the number of OFDM symbols in a single radio block is reduced to six, and the coverage area in suburban networks is increased to 120 km.

2.5.3.3 Architecture of the LTE Base Station Transmitter

Formation of a radio signal. The described features of forming a group signal are reflected not only in the LTE network interface but also in the architecture of the transmitter, which provides multi-station access with orthogonal multiplexing and frequency division of channels (OFDMA) (Fig. 2.29).

Independently on the message source type, in the system of the LTE standard, the signal in the transmitter section converted in the digital form is structured in the form of data packets. This permits to effectively distribute the system resources and to increase the data transmission rate.

Data from different types of sources is redundantly encoded using forward error correction (FEC) codes or error detection. Such codes are an effective way to detect or correct single (in one symbols) errors. Digital information sequences with different bit rates are fed to interleavers. In the interleaver, a 10 ms fragment of a digital stream is written into the rows of the RAM matrix and read out in columns. Such a modified sequence in the process of transmission in the radio channel is exposed to interference, distorting single symbols and generating error bursts, which is typical for mobile communication systems. The inclusion on the receiving side of the RAM matrix complementary to the transmitter matrix, where the received symbols are written column by column and read by row, turns error packets from several symbols into single ones. Such random errors in the recovered digital sequence are easily detected in the decoder using the maximum likelihood algorithm, which significantly reduces the probability of erroneously received bits (BER, bit error rate).

Then, in the MUX multiplexer, a baseband signal with a sequential structure is formed, when the digital sequences from the sources follow one after another – the general information digital sequence.

Obtained digital symbol sequence is divided into two orthogonal components, which are modulating sequences for I/Q modulators. The duration of modulating pulse is defined by the sub-carrier spacing, which is $\Delta f = 15$ kHz in the LTE system. To achieve their orthogonality on the whole length of the modulating pulse, its final part called the guard interval T_g is transferred in the pulse beginning (Fig. 2.28). This provides the modulating pulse orthogonality in the whole length $T = T_{\text{ort}}/N$, where N is the nominal number of sub-carriers (a number of point of the Fourier transform).

In a serial/parallel converter (S / P converter), each bit of the information sequence is converted by some algorithm into two slow orthogonal sequences with the symbol length $T_s = 66.7$ μ s, which is much larger than the information ones. Each of them is multiplied in the modulator with its own orthogonal sub-carrier.

The modulation is performed with the help of the inverse fast Fourier transform (IFFT). Twelve sub-carriers integrated on some algorithm form the OFDM symbol.

To form a group modulated signal with a parallel structure, a circuitry or algorithmic method can be used. The number of sub-carriers used is determined by the bandwidth allocated to the system and can reach several thousands, which will require an appropriate number of mixers and oscillators. The algorithmic method of forming an amplitude/phase shift keyed signal

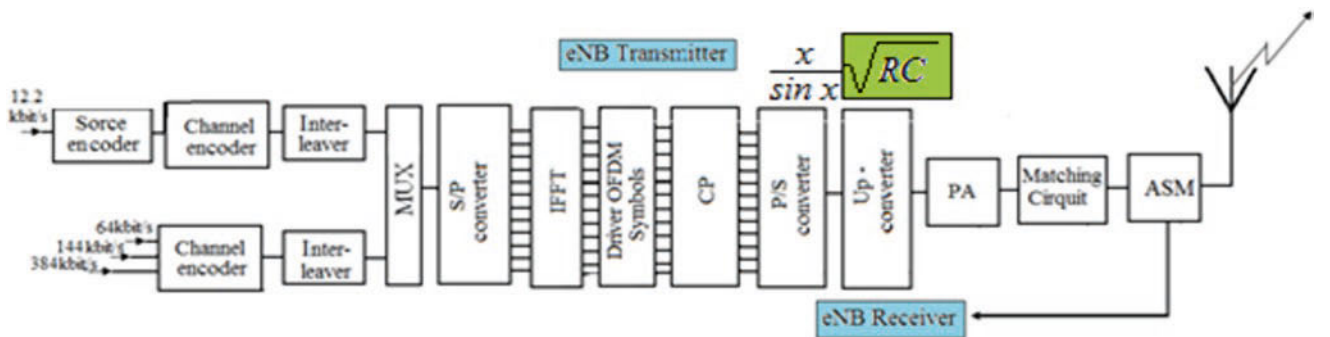


Fig. 2.29 Architecture of a BS transmitter that provides multi-station access with orthogonal multiplexing and frequency division of channels

allows for this to use the inverse discrete Fourier transform (IFFT). A signal with a parallel structure at the output of such a converter, a digital signal processor (DSP), has the number of sub-carriers determined by the number of Fourier transform points (2^n , $n \in \mathbb{Z}$). The resulting redundant (extra) sub-carriers are placed at the edges of the operating range, creating guard gaps.

The next block of the structural diagram Driver OFDM symbols changes the structure of the output signal, forming radio blocks and basic time-frequency structures, including 12 sub-carriers and 6 (or 7) OFDM symbols. The length of each OFDM symbol ($T_s = 66.7 \mu\text{s}$) is equal to the symbol length of the sequence modulating the orthogonal sub-carrier.

A cyclic prefix is introduced into the OFDM symbol structure, equal in length to the guard interval $T_{CP} = T_g$ and including the guard intervals of the radio block sub-carriers. Its inclusion will allow to preserve the mutual orthogonality of modulated sub-carriers and eliminate the influence of fading in the radio channel on the quality of the received signal. Since the distortion of the leading edge of the modulating symbol falls on the initial section (Fig. 2.28), discarding this part (CP) on the receiving side and replacing part of it with the tail part will eliminate the distortion of the envelope of the received symbol.

After the inclusion of the CP in the symbol, the parallel structure of amplitude/phase shift keyed signals on orthogonal sub-carriers is transformed by a P/S converter to a serial one.

All signal transformations in the transmitter were performed with signals having unlimited, but mutually orthogonal spectra. To transmit them over a radio channel with a limited bandwidth, it is necessary to reduce the bandwidth occupied by the radio signal, while maintaining the mutual orthogonality of its spectral components.

To fulfill these conditions, a Nyquist filter with a frequency response of the square root of the raised cosine ($\sqrt{RC}$) type is included in the transmitter path. A feature of such a filter is the linearity of the phase response – the constancy of the delay time of the frequency components of the spectrum, in contrast, for example, to the band pass filter, with a Π -shaped frequency response. This ensures the constancy of the frequency response of the radio channel in the operating frequency band.

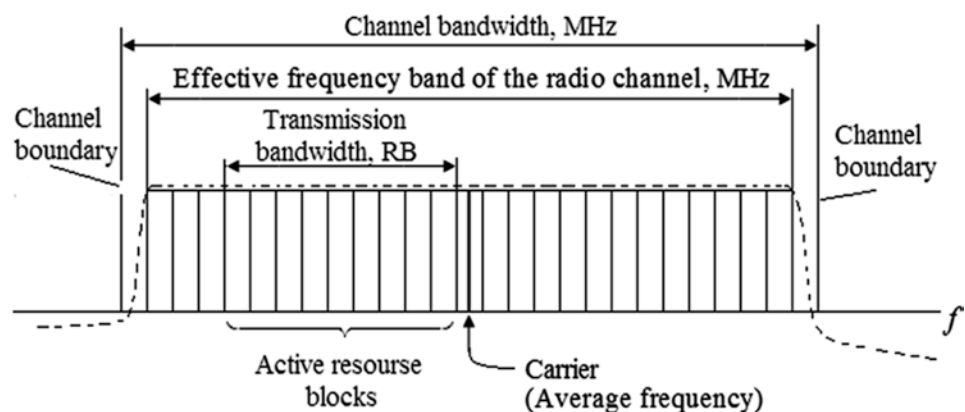
Spectrum of Radio Signal After the performed limiting and corrective measures, the signal is sent to up converter, where its spectrum from intermediate frequencies is transferred to the operating range and, after amplification in the power amplifier of the transmitter, is radiated into free space.

In contrast to CDMA technology, the variation of the transmission bandwidth is easily realized in OFDM. In CDMA systems, the operating bandwidth is fixed and defined as the quantity inversed to the chip transmission rate of elementary pulse ($1/T_{cl}$). In OFDM systems, the value of sub-carrier offset is defined as the inversed quantity to the integration time of IFFT. However, a number of sub-carrier and hence the transmission bandwidth are defined independently. This allows creation in LTE of different band configurations from 1.4 MHz to 20 MHz presenting more possibilities for subscribers. Determination of the channel bandwidth and the configuration of the transmission bandwidth for the radio access technology OFDM [46] for the modernized network of radio access E-UTRA (Evolved UMTS Terrestrial Radio Access) are shown in Fig. 2.30.

The channel bandwidth (MHz) specifies the nominal channel bandwidth defining the sub-carrier offset. The transmission bandwidth is defined through the resource block (RB), which can be transmitted through the channel of any configuration chosen from the permissible list of operating frequencies, regions, and the traffic. On the physical level, the resource block occupies 0.5 ms (1 slot) in the time domain or 180 kHz in the frequency domain. The number of sub-carriers in the resource block and a number of symbols in the resource block transmitted in downlink direction or uplink alternatively and is the function of the length of the cyclic prefix and sub-carrier offset. The sub-carrier in downlink is not transmitted and in frequency corresponds to the DC component of the signal in the informational section.

Time-Frequency Domains Information in downlink and uplink channels is transmitted in parallel manner but with time shift defined in advance in the standard. A number of resource blocks NRB in resource grid is defined by the bandwidth of

Fig. 2.30 Determination of the channel bandwidth and the configuration of the transmission bandwidth for the radio access technology OFDM for the modernized network of radio access E-UTRA



radio channel and is from 6 (the bandwidth is 1.4 MHz) to 110 (which corresponds to the bandwidth of 20 MHz) for uplink or downlink channels in LTE. Each resource block is the minimal information unit allocated to each mobile terminal by the planning personnel of the base station. The resource block occupies 12 sub-carriers (180 kHz) and 7 or 6 OFDM symbols (depending on the type of cyclic prefix) under condition of slot duration keeping of 0.5 ms.

The base station informs the mobile terminal about the variant of resource distribution in each time slot on the allocated control channel.

To achieve the compatibility of LTE and UMTS systems, all time parameters of LTE specification are attached to the minimal time quant (a chip). For the standard step between sub-carriers $\Delta f = 15$ kHz, the symbol duration is $T_{el} = 32.55$ ns, which corresponds the clock frequency 30.72 MHz, which is multiple to the processing frequency 3.84 MHz ($8 \cdot 3.84 \text{ MHz} = 30.72 \text{ MHz}$). At radio frame duration (Fig. 2.31) of $T_f = 10$ ms, 307,200 symbols are packed up in this interval.

The information exchange between the base station and the mobile station is built according to the principle of cyclically repeated frames (radio frames). When using the frequency duplexing, such a frame in the full duplex mode consists of 20 slots each with duration of 0.5 ms with numeration from 0 to 19.

To struggle with multipath property, the cyclic prefix is switched on in the interval of each OFDM symbol. At prefix duration of $4.7 \mu\text{s}$, we provide the elimination of reflected signal influence, which passed the way by 1.4 km more than signal arrived on the channel of direct visibility. For cellular communication system under urban conditions, this is enough for elimination of the intersymbol interference. For its elimination when using of cells with radius up to 120 km (applied for digital mobile TV and radio broadcasting), we can use the expanded prefix with the length of $33.4 \mu\text{s}$ and apply the time slot structure with sub-carrier offset by $\Delta f = 7.5$ kHz.

The definite range of channel resources (Fig. 2.32) in the frequency-time domain called the resource grid is allocated to the separate subscriber terminal in each time slot.

Each element of the resource grid (a cell) corresponds to the single sub-carrier in the frequency domain and the single OFDM symbol in the time domain. In the presented variant, the resource block occupies 12 sub-carriers (the frequency band $12 \times 15 \text{ kHz} = 180 \text{ kHz}$) and 7 (or 6) OFDM symbols (depending on the type of the cyclic prefix) providing the total slot length of 0.5 ms.

The LTE system provides the operation in different frequency ranges (0.9; 2.1; 2.6; and 3.5 GHz) of various bandwidths occupied by a channel (1.4; 3; 5; 10; 15; and 20 MHz). This indicates on the high flexibility of such systems compared to the previous generations of cellular systems (GSM, UMTS, cdma2000, etc.) and facilitates on its active implementation. The improved version of LTE-A (Long-Term Evolution Advanced) provides the transmission rate of 1 Gbit/s; it downlink operation to the mobile subscriber. These wideband channels assume the most rational utilization of the frequency resource, more advanced technologies, which increases all system indices. Potential holding of the wideband channel is perceived by subscribers as giving the possibility to apply of all bands allocated by the radio channel, whereas, in actual fact, this channel is one of the many operating in the total frequency band. The perception for separate subscriber of connection to its “own” channel comes from the fact that the channel loading is varied, and it quickly disappears at achievement of the limit data exchange rate in hours of the largest loading. This perception is called the “trunking effect,” i.e., the expanding of bandwidth and increasing of amplification.

2.5.4 Reconfigurable Radio Systems (RRS)

2.5.4.1 General Statements

The main problem in the construction of radio engineering systems for the transmission of information is the optimal organization of the physical layer channels in the structure of the reference model of the interaction of open systems (Fig. 1.10), which determines the efficiency of using the radio-frequency spectrum allocated to the operator. Such static scheduling of

Fig. 2.31 The structure of the single radio frame

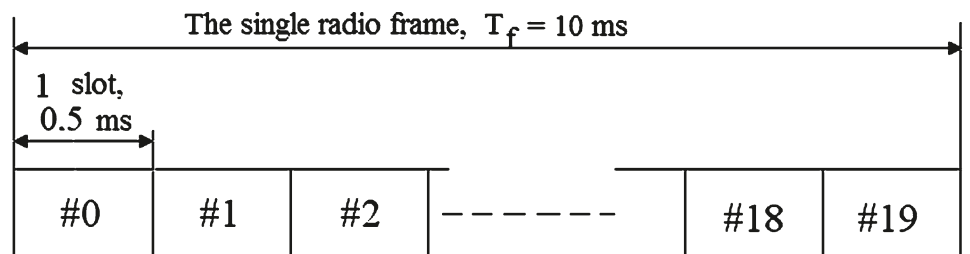
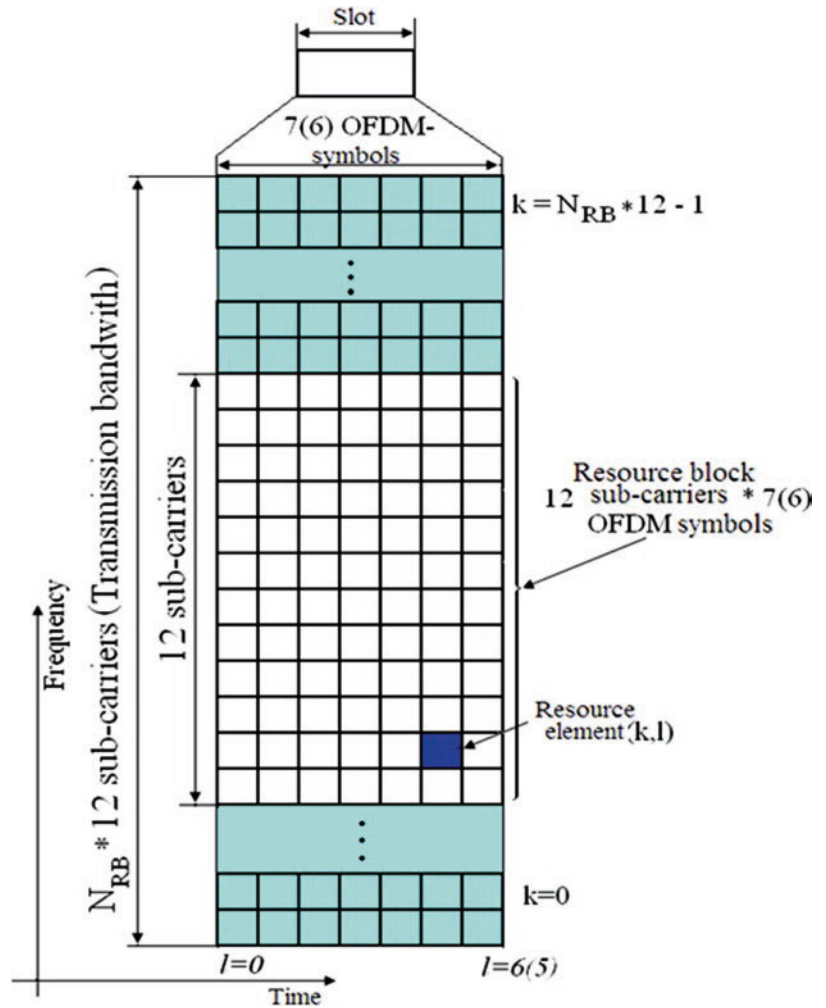


Fig. 2.32 The definite range of channel resources (Fig. 2.32) in the frequency-time domain called the resource grid is allocated to the separate subscriber terminal in each time slot



the allocated operating frequency range increases the efficiency of its use with fairly simple changes in the structure of the RF front-end of the receivers.

An analysis of the efficiency of using the spectrum allocated to a wireless system (network) showed that a significant part of the radio spectrum is used sporadically, forming its deficit in some time intervals, which is caused by the formation of a coverage area by one operator, without taking into account the possibility of using the congestion of networks of adjacent operators. Organizational measures in the distribution of the radio spectrum with the transition from analog to digital television and radio broadcasting have freed only a small part of it. The complication of radio interfaces by optimization and, at the same time, the complication of modulation types are limited by a number of technological and software factors. The use of separate parts of the spectrum allocated to specific systems and operators can be considered promising. This version of the development of radio technologies has led to the emergence of SDR (Software-Defined Radio), adapted to the surrounding network environment, which is reflected in the standards of LTE Advanced and IEEE802.11h systems (Fig. 2.21).

In addition to the previously discussed spectrum aggregation performed by the scheduler during the network design phase, this can be solved by monitoring the spectrum usage of base stations interacting with the UE in real time. The transition to the dynamic regulation of the radio resource will require the convergence of radio technologies and the creation of reconfigurable systems using the methods of dynamic reuse of the free part of the radio spectrum. This will require:

- Implementation of methods for processing dynamically allocated radio resource.
- Reconfigurable network management methods.
- Dynamic network scheduling, taking into account real-time performance of the UE.
- Ensure the openness of the network resource, enhancing the functionality of the UE and the entire network.
- Dynamic management of the allocated radio-frequency spectrum RRM (Radio Resource Management).

This concept of dynamic spectrum allocation DSA (Dynamic Spectrum Allocation) is implemented in the cognitive radio CR (Cognitive Radio), adapted to the surrounding radio-frequency spectrum [38]. Cognitive radio is a technology that

allows, based on the analysis of the electromagnetic environment, using software tools, to vary the parameters of the transmitter/receiver without changing its hardware structure. These features of DSA are reflected in the architecture of the reconfigured cognitive radio-technical system CRS (Cognitive Radio System), which allows to optimize the spectral characteristics of interacting systems and provides access to both the radio resource of its own network and shared access to the spectrum of other operators at the current time, without creating interference with them. Advanced cognitive radio technology optimizes the dynamic spectral characteristics of interacting systems by dynamically and autonomously adjusting operating parameters and transceiver protocols.

2.5.4.2 Static Methods for Increasing the Use of Radio Resources

Spectrum Aggregation The evolutionary development of the LTE system is the LTE Advanced system, which has an operating channel bandwidth of 100 MHz (Table 2.4). This is achieved by aggregating (combining) free spectral bands of neighboring operators and using space-time coding technology based on the use of multiple transmit antennas at the eNB and at the mobile terminal UE (MIMO technology). If the bandwidth extension can increase the peak rate to 3 Gbps in the downlink channel, then the use of a multi-antenna MIMO system will increase the efficiency up to 15 bit/s/Hz in the downward direction, when the eNB uses 4 transmit and 4 receiver antennas (4×4) and up to 6.75 bps/Hz in the uplink channel when using a combination of two antennas for transmitting and four antennas for receiving (2×4).

Among the possible options for aggregating spectra, the simplest is the combination of adjacent frequency bands that have the same transmission and reception width (Fig. 2.33).

The aggregation option is determined at the planning stage when drawing up a map of coverage of the territory and distribution of frequencies within its own network and neighboring operators. When aggregating bands, it is preferable to select components that have a wider occupied bandwidth and belong to the same class of aggregated frequency bands CA (Carrier Aggregation), determined by the number of resource blocks of the combined spectra.

Based on the compatibility requirements of LTE Advanced and LTE systems, the maximum size of one frequency component is 100 RB, that is, 5 frequency components with a 20 MHz bandwidth (CC, Component Carrier) can be combined during aggregation. The UE in the uplink cannot have more components than in the downlink direction with frequency duplexing, and with time duplexing, the number of opposing aggregated components must be the same.

An important factor helping to maintain a low error probability in the LTE Advanced system with aggregated spectrum is its symmetry about the center frequency of the radio channel (Fig. 2.34).

Radio blocks located near the center of an asymmetric aggregated channel (Figure 2.34a) are “destroyed” during signal processing in a receiver with OFDM technology, leading to the loss of information they contain.

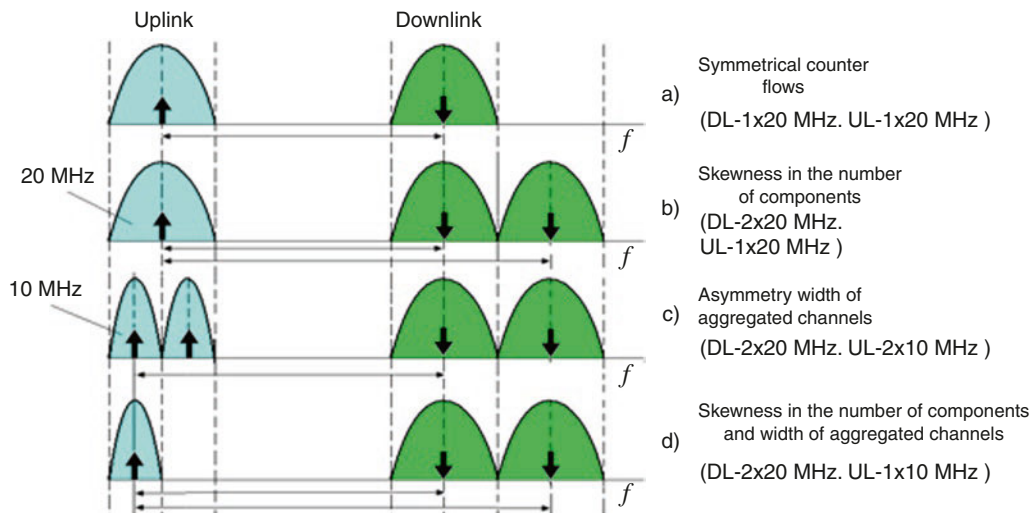


Fig. 2.33 An example of the arrangement of spectra “up” and “down” without aggregation, (a) different number of aggregated bands and their bandwidth occupied (b), (c), (d)

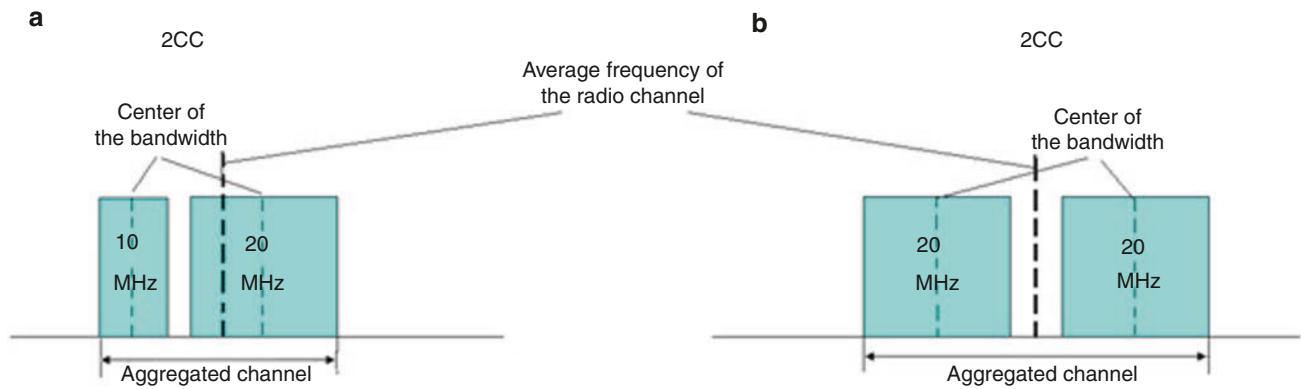


Fig. 2.34 Variants of placing the central frequency of the aggregated spectrum in the operating frequency band (a) in the spectrum of one of the combined frequency bands, (b) outside the bands of the combined spectra

2.5.4.3 Software-Defined Radio (SDR)

Software-defined radio is a collection of software and hardware technologies that redesign the interface for wireless networks and user terminals. The most difficult task in the construction of subscriber terminals is their receiving part, which ensures the implementation of the listed requirements of the structural diagrams of receivers. The most promising option seems to be the use of receivers with direct conversion or superheterodyne-type architecture and a digital intermediate frequency. The main advantage of such structures is the relative simplicity of non-tunable preselectors. The use of a single conversion of the input signal simplifies the intermediate frequency path of the superheterodyne-type receiver, and the transition from the analog signal to the digital form simplifies or excludes the demodulator from its structure. The use of high-speed ADCs will allow converting an analog signal into digital form at the frequency of the received signal, which will eliminate the intermediate frequency path from the receiver architecture. In this case, the main characteristics of the receiver will be provided by software, and not by reconfiguring its structure, that is, by circuitry methods.

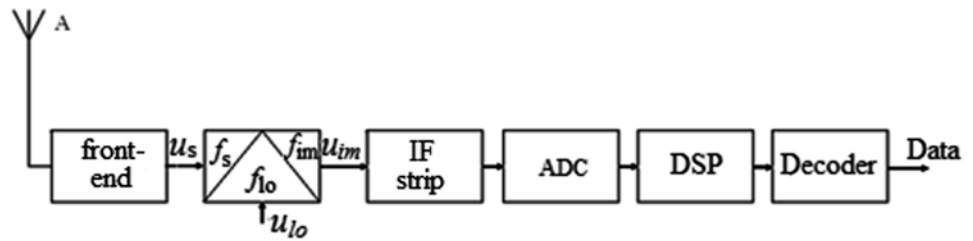
The transition to the digital format of the received signal in the first stages of the receiving device of mobile systems will save only the input radio-frequency module that separates the counter streams of information signals (reception/transmission) and one or more stages of low-noise amplifiers (front-end). This allows manufacturers to create designs that implement the functions of different types of devices that have scalability and flexibility of decisions. The implementation of SDR creates opportunities for users to get new services and services when connecting to networks with different standards while providing intra- and inter-network roaming. The enlarged architecture of the receiver with a digital intermediate frequency is shown in Fig. 2.35.

A broadband antenna (or several antennas when using MIMO technology) tuned to the middle of the received frequency band of the system forms at the input of a receiver with a superheterodyne-type architecture, an EMF containing a voltage at the frequency f_s of the useful signal. After limiting the band of received frequencies in the HF path and converting its part with the help of a frequency converter “down,” the channel bandwidth of the active base station is allocated in the intermediate frequency path. The signal spectrum at the input of the UE can be aggregated and include the frequency bands of the own operator’s BS and the signals of the BS of the neighboring cell transmitted to the active BS via transport channels. The transfer of the received signal to a single intermediate frequency will be provided using a tunable local oscillator, the frequency f_{LO} of which is changed by a command from the UE digital path.

An analog signal in an analog-to-digital converter (ADC) is converted into digital form, and in a digital signal processor (DSP) procedures are performed over it, eliminating intersymbol interference, fading, which are based on correlation signal processing. Reconstruction of the information sequence is performed in the decoder, which uses the “shortest” path search method on the trellis diagram to generate the bit sequence based on the analysis of the entire received sequence of one frame length (Viterbi decoder).

The considered superheterodyne structure of the receiver’s radio path has a number of advantages: transferring the signal spectrum to an intermediate frequency simplifies the struggle with adjacent channels, increasing the signal-to-noise ratio at the detector (decoder) input, and reducing the requirements for the ADC (relatively low sampling frequency, high input signal level). At the same time, an increase in the data transmission rate by expanding the bandwidth of the operating frequencies by aggregation with the frequency bands allocated to other systems requires the use of a higher intermediate frequency, which reduces the SNR at the detector input and increases the requirements for filtering devices of the intermediate frequency path.

Fig. 2.35 Simplified block diagram of a digital intermediate frequency receiver



The main requirement that determines where the ADC is switched on is the condition that the sampling frequency must be at least twice the carrier frequency of the received signal (Kotelnikov theorem). Modern advances in integrated technology make it possible to design high-speed ADC/DAC converters, which simplify the structure of the receiver front-end. For example, using Xilinx RFSoc, you can sample the received signal in a 12-bit ADC at up to 4 GHz and generate an analog signal with a 14-bit DAC at 6.4 GHz. This indicates the possibility of excluding the mixer and demodulator from the structure of the receiver of modern wireless LTE systems and digital television DVB-T2 (Fig. 2.35) and switching on the sampler directly at the output of the high-frequency path.

2.5.4.4 Cognitive Radio

Dynamic Spectrum Allocation The concept of cognitive radio is based on the dynamic spectrum allocation technology based on IP for mobile equipment **DRIVE** (Dynamic Radio for IP Services in Vehicular Environments). The architecture of this project integrates various access technologies, increasing the efficiency of using the radio spectrum by choosing the optimal technology when providing a specific type of service. It includes a number of necessary functions and procedures, among which the main one is the function of traffic control for the duration of the entire connection session. Optimization of its indicators is based on knowledge of the parameters of the user terminal and the required quality of service parameters, as well as its priority and properties of the selected network. The procedure for the user to select a radio access network is performed every time the AT is connected or the state of the radio channel changes.

The concept of dynamic spectrum allocation DSA is based on constant monitoring and searching for the region of potentially available radio-frequency spectrum, measuring its parameters and the possibility of using the free part of the spectrum at the current time.

There are two ways to reallocate the spectrum: the distribution of adjacent frequency bands and the distribution of spectrum fragments in different bands and frequency ranges. The algorithm for dynamic redistribution of the frequency spectrum of interacting radio access systems is based on knowledge of the statistics of changes in their traffic. Kalman filters, as well as artificial neural networks and genetic algorithms, are considered as the basis for the synthesis of such an algorithm. At present, the analysis of traffic and network parameters is used in 3G and 4G systems for intranet and inter-network roaming and for measuring the power level acting at the input of the AT receiver from the active and neighboring BS. With a fixed allocation of the radio-frequency resource **FSA** (Fixed Spectrum Allocation), this allows you to select the “main” BS, which creates the maximum signal level at the AT input in comparison with the BS from neighboring cells.

The transition to dynamic DSA spectrum allocation presupposes the mandatory use of the dynamic frequency selection (**DFS**) protocol for detecting devices using the same spectrum (channel). After measuring the power distribution in the operating frequency band and its compliance with the transmit power control (**TPC**) protocol, a decision is made on the possibility of using part of the operating band in another system. It is obvious that the transition to dynamic spectrum allocation complicates the algorithm of interaction between terminal devices during an information transfer session, which presupposes a preliminary scan of all potentially usable spectrum areas. At the same time, the demands on the amount of RAM memory and the computing power of devices operating in real time are increasing in cognitive radio. Proposed for calculating the power of the total interference acting at the input of the device, the DFS standard allows you to calculate the degree of their impact when moving to another frequency region of the licensed range. When an excess of the interference level determined by the DFS protocol is detected, the FH (Frequency Hopping) can be used, determined by some algorithm.

Cognitive Network Architecture As follows from the description of the principles of interaction between software-configurable SDR and cognitive devices CR, such can be subscriber terminals UE and BS of traditional systems, as well as multi-user equipment (**MUE**) of cognitive networks. The right side of Fig. 2.36 shows a composite wireless network (**CWN**), which includes a conventional radio access network and cognitive network elements.

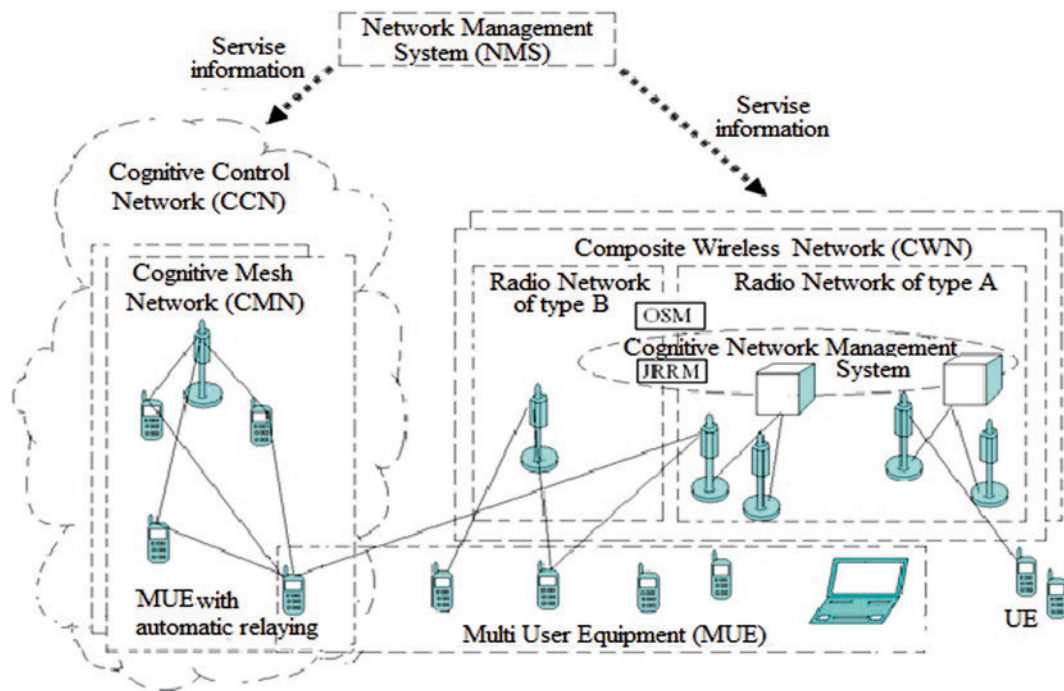


Fig. 2.36 Architecture option for cognitive radio

The network structure includes a set of radio access networks, the centralized management of which is performed by a common system network management (NMS) from one operator. The NMS structure includes base stations BS and subscriber units MUE, which can be reconfigurable and, accordingly, change their characteristics by software, relying on the multiradio cognitive radio technology. The structure of the CWN wireless network can include both traditional radio access networks and a cognitive network management system (C-NMS) using dynamic DSA spectrum allocation.

Elements of centralized management of the C-NMS cognitive system (Fig. 2.36), based on information about the network load and spectrum structure, distribute traffic between different **JRRM** (Joint Radio Resource Management) radio networks and manage the available joint radio spectrum operator spectrum management (OSM), distributing it among cognitive devices based on dynamic spectrum indicators. The cognitive network element manages (JRRM) shared resources and the choice of radio access technologies, increasing the spectral efficiency of the system by choosing the optimal modulation types.

In a Type A radio network (Fig. 2.36), the E-UTRAN terrestrial radio access network may include BSs of traditional and cognitive networks. In wireless complex CWN, it is possible to select a core network focused on working only in a radio network (type B), the terrestrial radio access network of which serves only cognitive subscriber devices MUE.

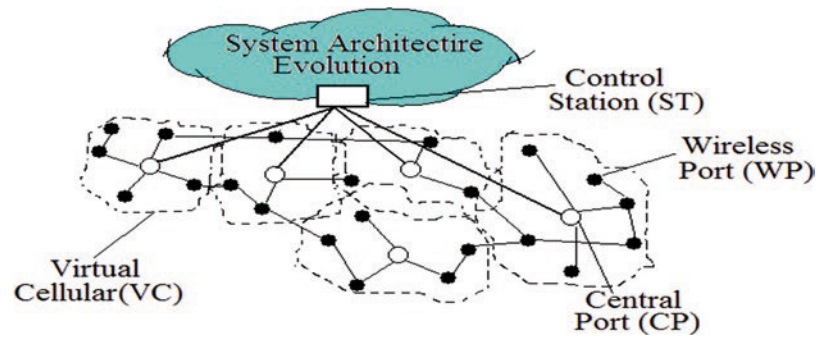
In the left part of Fig. 2.36, the structure of the cognitive mesh network CMN is highlighted in the cognitive control network (CCN). The CMN terrestrial network elements managed by the regulatory information system (CNN) contain only cognitive subscriber devices.

The design of the MUE should ensure the operation of the terminal with different standards and a dynamically controlled operating frequency range but also provide an active automatic retransmission mode. In such a mesh network, the AT can connect to the BS both traditionally – directly – and through MUEs, which perform the function of repeaters that transmit information along the chain. It is clear that the architecture of such a network, called a multi-hop network, is formed dynamically in the process of establishing connections.

The use of such architecture will significantly increase the data transfer rate. However, since this problem is solved when using OFDM technology by increasing the order of quadrature modulation (nQAM), this entails the need to increase the output power of the AT transmitter to maintain the same error probability. If you reduce the coverage area, then this increases the number of transmitted overhead messages due to the increase in the number of handovers between cells.

One of the possible ways to solve this problem is the organization of virtual multi-hop networks, multi-hop VCN (Virtual Cellular Network) in existing cellular networks (Fig. 2.37).

Fig. 2.37 Multi-hop virtual network architecture



The network consists of several fixed central ports (CP), which are gateways connecting the core network and many mobile (wireless) ports (WP). Wireless ports distributed in the virtual cell (VC) structure are connected directly to mobile MUE terminals, and then they are called end and in others remote.

While the CP (Fig. 2.37) occupies a fixed position in space and has known characteristics, the distributed wireless ports function as virtual base stations. It is understood that in this case, the transmitter power of the MUE in the virtual cell may be less than with standard network planning.

The described VCN structure differs from the cognitive mesh network (CMN) in that signals from the subscriber device are received by the WP end port, and through the central port, which provides a multi-hop connection to the control station, and the MUE is connected to the core network.

The functions of the control station also include the management of the mobility of MUEs, which requires constant monitoring of their location and, accordingly, a significant increase in the amount of service information in the network. An alternative is to use a distributed control structure, where mobility control in a virtual cell is transferred to a central port and between cells to a control station.

2.5.5 Wireless Systems of Fifth Generation

Implementation of 4G systems of mobile communication is naturally accompanied by new ideas and technologies development, which will be realized in the new following fifth generation of cellular communication systems [80, 81]. Currently, there is no single center for developing recommendations for the implementation of 5G networks, but it is known that a number of countries are actively involved in this problem. The greatest success in developing standards for hardware and software was achieved by Huawei (China), with the support of IMT 2020. Two centers Stanford CIS (UC SNARN) and NYU Wireless (WIN LAB) in the USA are also actively developing this area. Under the auspices of Ericsson company, the METIS consortium (Mobile and Wireless Communications for the Information Society Twenty-twenty) was formed. Within the framework of 5G Forum (Republic of Korea). 8 working groups have been formed, the main task of which is to develop a technological agreement, which makes it possible to prepare for the consideration of engineering issues of building a system.

On the base of the University of Surrey (UK), another project is developed called 5GIC, the purpose of which is the development of new capacity improvement conceptions of communication networks and the radio-frequency spectrum. In addition, there is the ISRA project, which is coordinated by Intel Company.

Wireless systems have appeared as a means of communication for transmitting a speech signal from one subscriber to another, when one of them may be on the move, in recent years they have begun to actively replace any traditional methods of obtaining information. To them we can attribute the mobile radio and TV, and also a possibility to render the services, which did not exist earlier, for instance, the mobile Internet, interactive TV, etc. The main aim of the information stream formation is the traffic creation between terminals (cars), but not persons.

2.5.5.1 Prerequisites for the Emergence of the 5G System and Its Main Characteristics

In the development and implementation of 5G sellular systems, there is not only the desire of some developed countries, which is expressed in the decisions of the World Mobile Congress in Barcelona in 2014, but also serious investments in this direction. The ideology of new evolutionally developed generation of mobile communication systems is not only the increase of the access rate to information resources but the new base for construction of the network platform. This will permit to

solve the integration problem of telecommunication systems and TV, which requires serious investments in the transport infrastructure. This will be able to activate the country transfer in the stage of “digital economics” and the Internet transformation into the “Internet of Everything.”

It is clear that the huge number of different applications and services currently in force has led to the emergence of a large number of standards and regulations that describe the requirements for the systems being developed and that allow them to be built into existing structures. The fifth-generation networks are based on the requirements for the provision of services based on existing land communication networks and radio access systems, which, in terms of IMT 2020, include, in addition to new interfaces, an evolving 4G network and a WiFiG radio access network (WiFi network designed to operate at frequencies in the 60 GHz).

There are several factors influencing the prospects for the development of mobile traffic:

- The growing demand for services for streaming radio and video content of ultra-high definition **UHD** (Ultra-High Definition).
- Active use of smartphones.
- Use of smartphone applications.

Currently, there are 300 billion app downloads. Purchasing mobile apps and updating them regularly will drive mobile traffic growth. These trends will become the fundamental directions for the growth of communication networks, the Internet, and new services, among which the following will be massive:

- *Services based on holograms and multimedia with a full effect of presence.*
- Currently, 4G networks provide HD (high definition) video calling and VoLTE (voice of LTE) sound quality. However, such systems do not provide an immersive experience for remote users. 5G network will allow transmitting ultra-high definition UHD 4 k-UHD, 8 k-UHD video signal. However, to obtain 3D visualization, a terabit-level network bandwidth is required, which is not provided even in 5G networks, and therefore, instead of a holographic one, a projection stereoscopic image is proposed.
- *Full-scale services of virtual and added reality VR/AR (virtual reality/added reality).*
- The bandwidth of LTE systems does not provide an immersive experience in mobile mode, but is limited to presenting content from a HD video camera, GPS, and other sources. The developed 5G network will allow not to be limited to information received from a smartphone, but, using additional means (sensors), to receive comprehensive, for example, about the surrounding space or closely spaced objects on additional devices (augmented reality glasses, car windshield, etc.).
- *Ultra-low signal delay in time.* The LTE-A network provides data rates as well as system delays at the wired network level. The implementation of 5G networks should reduce time delays by introducing new components into cloud-based transport networks and, using intelligent analysis, optimize the network structure, providing signal latency of several milliseconds. This will enable remote control of, for example, a rescue robot. Another example of the use of such networks is communication between D2D (device to device) devices. In this case, direct communication between the devices is organized, under the control of the network. A new direction for 5G development is the construction of low-latency networks, the active use of which should be expected in telemedicine and self-driving cars.
- *Internet-of-Things services.* If at present the use of the IoT (Internet of Things) is limited to radio identification systems RFID (radio identification), then the introduction of 5G will allow, based on new technologies, to remotely monitor an object or the state of the environment and change the location of an object. An increase in the number of connected devices (sensors) will make it possible to use remote control of a car and to create structures such as “smart home,” “smart plant,” and “smart city,”
- *Intelligent service based on large amounts of data.* The development of new intelligent services based on the use of unstructured data in a multidimensional space will make it possible to analyze them not only in the direction of increasing business efficiency but also for the operation and management of the network. The convergence of IoT and big data technology, coupled with the 5G network, could evolve into “service intelligence” technology, which could become the platform for the future lifestyles of people.

2.5.5.2 Radio Interface and Multiple Access in 5G System

Key requirements for the implementation of a 5G system are formulated in the recommendations of 3GPP TR 38.913 [76]:

- Data transmission downstream should be 20 Gbit/s with a spectral efficiency of 20 bit/s/Hz.
- The peak uplink data rate should be 10 Gbit/s with a spectral efficiency of 15 bit/s/Hz.
- The maximum density of devices from the world of the “Internet of Things” connected to the city network is 1,000,000 devices per km².
- Minimum delay in radio access sub-systems for URLLC services is 0.5 ms and for eMBB services 4 ms.
- Autonomous operation of devices from the world of the “Internet of Things” without recharging the battery for 10 years.
- Support for mobility at a maximum speed of movement of objects of 500 km/h.

Their implementation defines the basic technologies that will be used in the architecture of 5G networks; these are network function virtualization (NFV) and software-defined networking (SDN). They allow providing *ultra-broadband mobile access* **eMBB** (enhanced mobile broadband), *ultra-reliable communication* with low latency **URLLC** (ultra-reliable and low-latency communications), and *mass connection of various sensors* and actuators from the world of the “Internet of Things” **mMTC** (massive machine-type communications).

In some countries, the stage of scientific developments is already finishing, and the transfer is planning into commercial exploitation of 5G networks: in China in 2016, in Republic of Korea in 2018, and in the USA in 2020. In Russia, from 8 January 2020, the operator MTS (Mobile Telecommunication Systems) is putting into commercial operation a 5G network in the frequency range of 24.25–27.5 GHz and 37–43.5 GHz at large industries, enterprises, and logistics centers. With low power values of base stations, amounting to several tens of mW and signal propagation over distances limited by line of sight, as well as strong signal absorption in building structures, the use of such networks is safe and very promising. With a small coverage radius, several tens (hundreds) of meters, this can transfer data rates in the downward direction up to 20 Gbps. Such speeds are required, for example, for 3D video, working and playing in the cloud, unmanned vehicles, and other applications. The organization of such pico- and femto-networks, “embedded” in the macro-networks, allows to reduce the power of their BS and make the power distribution more even in the structure of the terrestrial network.

The tendency outlined in Russia is the decrease of the on-air TV broadcasting and its active replacement by the wideband access to TV and by the satellite TV. There is a noticeable interest in Russia and in the world to assimilation of the frequency range of 470–740 MHz for the wideband access systems. A possibility appears to transmit and to receive information in networks of mobile communication in the TV spectrum. Appearance of terminals, which have the possibility of dynamic choice of the operating frequency, allows arrangement of almost 400 usual digital TV channels in the operating bandwidth of 470–862 MHz, including the high-definition channels. The frequency range below 1 GHz, where good coverage is provided, can be organized networks that, at low cost, can be used to implement a number of 5G-based applications that do not require high speeds, for example, services in the IoT segment.

The development of networks in the range below 6 GHz (presumably in the range of 3.4–3.6 GHz) will allow obtaining wide frequency bands compared to 3G/LTE networks, providing high capacity and coverage of territories, as well as mass availability of speeds up to 100 Mbit/s. It is expected that continuous frequency bands of 300–400 MHz will be allocated for 5G systems and one operator up to 100 MHz in large cities with a population of more than one million inhabitants.

The main technological directions in network formation become the decrease of cell dimensions, the transfer from the concept of the cell to the concept of the wideband “cloud,” and joint operation with networks of other technologies (mainly WiFi). The network architecture will change: accelerated movement from hardware solutions to software and virtualization of its control.

The mankind is actively moving to the information society, and new generations of cellular systems must correspond to growing requirements.

2.5.5.3 5G Network Architecture

The Choice of the Channel Access Technology As it follows from the requirements for a 5G system, it must distinguish broadband, the ability to transmit signals with significantly different spectral and temporal characteristics, to provide a high data rate and signals with a low system delay.

Considering the LTE-A system and reconfigurable systems as the base for 5G, we will evaluate the applicability of the principles laid down in the architecture of these networks, methods of accessing the network, and the possibility of hardware implementation of the components. We will restrict ourselves to considering the processes at the physical level as the most important for the construction of radio receivers.

Code division multiplexing digital wireless communication systems of the third generation (cdma2000, IS-95, UMTS) use the principle of time synchronization of signal conversion processes in the transceiver and coherent detection on the receiving side as an effective way to combat intersymbol interference and fading.

4G systems use the most popular technology (OFDM) with higher efficiency and orthogonality in both frequency and time. The demodulation process on the receiving side is based on the principle of algorithmic signal processing using the Fourier transform. If the orthogonality of sub-carriers is violated, for example, due to interference in the radio channel or mutual influence of cells, the distortions increase indefinitely, which is caused by the reproduction of the Dirichlet kernel $\sin[\frac{f_0}{N}](Nx)/\sin x$, which rapidly degrades to $\sin[\frac{f_0}{N}](x)/x$ kernel, for large N , where N is the sub-carrier number. These phenomena are exacerbated by the frequency shift caused by the instability of AT frequency generators and base stations, the sampling rate, and the Doppler effect and can increase according to the logarithmic law: $\log N$. These reasons require, in order to maintain the orthogonality of the spectral components of the received signal, to monitor the channel state in real time and adjust the characteristics of the receiving path, which is a technically difficult task that causes an increase in the signal delay. An alternative is to relax the requirement for strict orthogonality of sub-carriers while reducing their number.

There is no problem of achieving synchronous operation of signal sources for sporadic MTC traffic, due to the low speed of transmitted messages. However, the transition to an active state from the hibernation state, known as “fast dormancy,” used to increase battery life, is a challenge to the radio access network. With the standard approach, UE for the transmission of some short message must be provided with complete synchronization (authorization of the subscriber, assignment of a current number to him, registration in the network, etc.).

In real conditions in the IoT network, this can happen several hundred times a day, which dramatically increases the amount of service information and can lead to network failure.

After the source is in the active mode, it is sufficient to use “coarse synchronization,” for example, reducing the number of resource elements in the resource block of the LTE-A system and transmitting them at a lower rate. However, simplified network requirements will require significant improvements in operational capabilities and features such as the reliability and longevity of autonomous MTC nodes. The cardinal solution is to abandon the use of orthogonal signals and to use the random access channel RACH for asynchronous process control.

One of the important disadvantages of the LTE-A system is the inclusion of a cyclic prefix in the OFDM symbol, which ensures orthogonality of sub-carriers at the output of the radio channel. On the length of the frame, it takes 7–10% of the time, which leads to a decrease in the signal transmission rate. The large length of the cyclic prefix prevents the use of the existing system for data transmission, the latency of which is about 25 ms with frame duration of 10 ms, while in a 5G system the network latency time should not exceed 0.5 ms for URLLC services.

The use of non-orthogonal signals in a 5G system can lead to an increase in the data transfer rate and at the same time solve the problem characteristic of LTE-A – the elimination of the protective frequency regions that separate the spectra of different operators or aggregated frequency bands. These stringent restrictions make spectrum aggregation difficult, especially with a sufficiently narrow occupying band, and reduce index utilization of bandwidth.

The achievement of global coverage and the transmission of converged signals in the 5G system being created led to the emergence of cells with a small coverage radius, only a few meters; these are femto-cells and pico-cells. Obviously, with such cell sizes, it is practically impossible to avoid the overlap of coverage areas, which actually excludes the use of orthogonal types of modulation in data transmission, but it is possible to maintain synchronism and orthogonality in the control network for new types of modulation.

The Physical Layer of the Network In 5G networks, spectrum availability is one of the key challenges of supporting the enormous mobile traffic demand. Nowadays, the current spectrum is crowded already. Especially in very dense deployments, it will be necessary to go higher in frequency and use larger portions of free spectrum bands.

Another important feature of fifth-generation networks is the variety of users with very different indicators: in terms of the occupied frequency band, information transmission speed, the form of transmitted signals, and properties of the transmission medium. Fixed and mobile networks at the upper layers of the OSI open systems reference model (Fig. 1.10) transmission use cables as guiding media: multicore and coaxial. This is due to their high throughput and parameter stability.

In networks providing information and data transmission, when at least one of the subscribers is on the move, their connection to the fixed network “last mile problem” is solved using wireless access networks. Such systems include devices that provide communication between a telephone headset and a cell phone, a medical probe and an external monitoring device, and radio microphones for speakers and performers, as well as local radio communication systems that provide communication between monitoring objects and actuators, intra-office communication, and a home electronics control network. This should also include GSM radio communication networks, third- and fourth-generation broadband access networks, and, of course, the 5G network under development, the architecture of which is shown in Fig. 2.38.

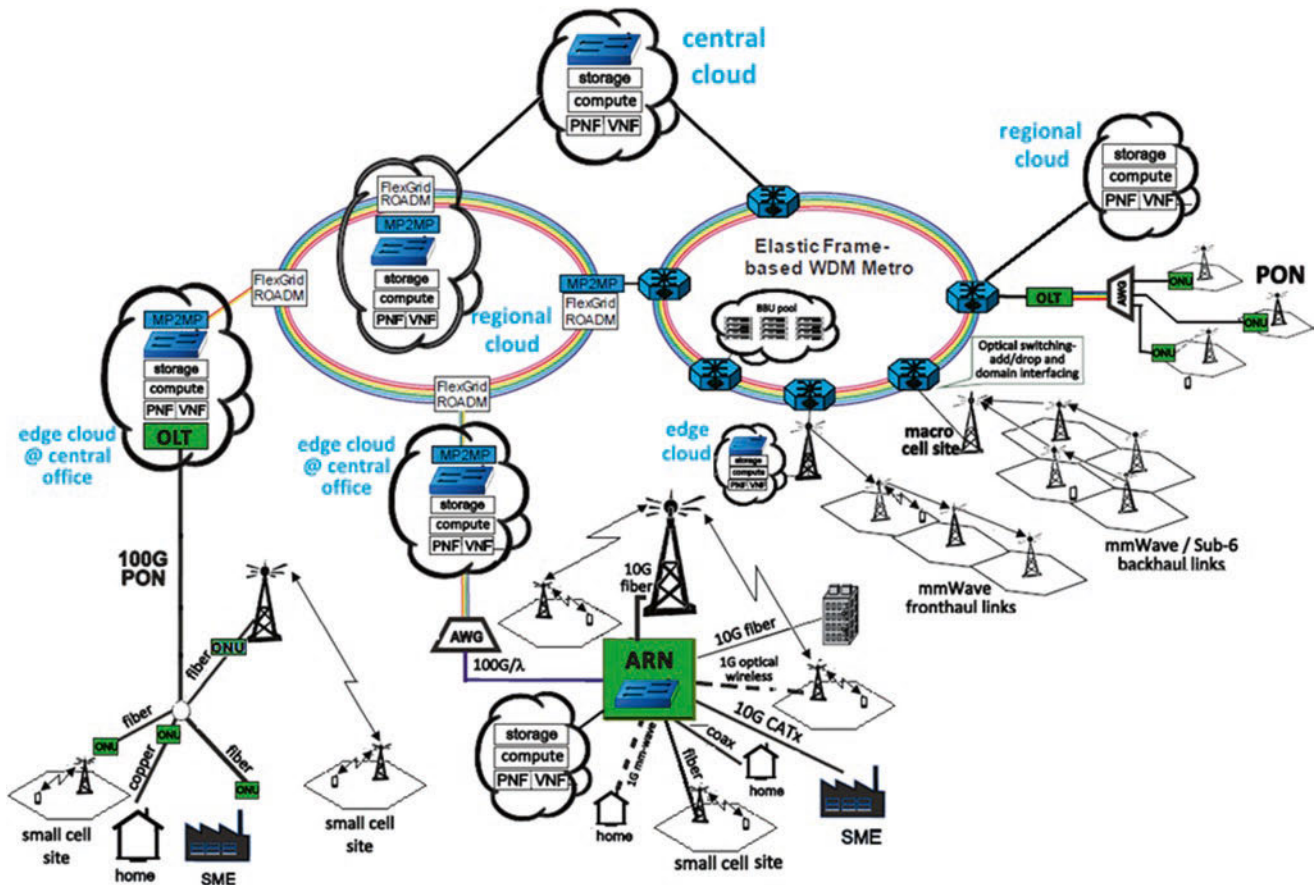


Fig. 2.38 5G physical layer architecture

The focus on fixed networks is reflected by the fact that many operators have a modern FTTH (fiber to the home) network, which strives to integrate fixed and mobile users into 5G networks through the convergence of fixed and mobile technologies. The fixed network includes domains: core, metro, and access. At the same time, metro and access domains are realized by a combination of passive and active electro-optical technologies introduced into the access and metro domains: passive optical network (PON), active remote node (ARN), multipoint-to-multipoint (MP2MP) technologies, and elastic frame-based optical switching nodes.

PON includes an optical line terminal (OLT), an optical line power splitter to a remote site (base station and trunk), and a separate line to optical network units (ONUs). The 5G network uses a new generation of passive optical network (NG-PON2), which provides the distribution of channels along the wavelength – wavelength-division multiplexing WDM (new generation of passive optical network) – which increases the network throughput up to several tens of Gbit/s. The idea of combining remote ARNs with the operational capabilities of Ethernet to perform statistical multiplexing with aggregated traffic transmitted to high-level aggregation nodes can be considered promising. Additionally, if a macro-network site is developing in a small data center, then this provides very low latency.

All access architectures terminate at a node called the central office (CO), where traffic from multiple BS is aggregated. As an alternative, we consider using ARN as a remote data center (instead of a CO), which allows you to bring network resources closer to the end user and expand the ability to calculate and store data in a remote network, near the radio network.

Given the variety of types of subscribers connected to the 5G network, such networks must be very flexible, be broadband, and be as close to the subscriber as possible. This challenge in a 5G network is accomplished in the traditional way via radio access network (RAN) small networks or cloud-based cloud radio access network RANs (C-RANs) when they are considered remote radio heads (RRHs). The RAN is connected to a centralized macro-network via a transport channel. With this separation, the main signal processing is performed in the RAN, which reduces the load on transport channels, and the C-RAN becomes a classic radio access network.

The use of new frequency bands will require the use of a new effective system of their interaction with the already mastered low-frequency part (below 6 GHz), as well as the management of the matrix structure of the antenna with MIMO

technology, based on the principle of coordinated multi-point (**CoMP**). Although these tasks were mainly solved in distributed RAN (**D-RAN**), the introduction of cloud-based C-RAN technology will reduce system overhead and power consumption. Comparing radio access systems, we can say that D-RAN introduces less time delays in data transmission and that C-RAN reduces power consumption and increases system efficiency. At the same time, implementing fronthaul-based C-RAN creates huge problems with converged fixed and mobile networks in the 5G system. It would have to transport waveform samples as ordinary payload, summing up to required data rates exceeding 1 Tbit/s per macro-cell site. Hence, there is a need to reduce the optical bandwidth utilization of the fronthaul. The solution to this problem is to split the baseband processing while maintaining the performance of the radio network.

This will functionally divide the radio access networks into D-RANs, where the information processing is transferred, performed locally at the remote radio head (RRH), to the fully centralized C-RAN, where some or all signal processing is shifted from a RRH to a baseband unit (**BBU**) in a central unit. The functions that the BBU implements, forming the transport stream, can be performed by a virtual network function (VNF) using software. Thus, the BBU node will combine the functions of a physical and virtual network, providing control over the position of the mobile subscriber, at the same time being part of the dynamic radio access network C-RAN.

All access architectures end in a so-called central office (CO) where the traffic from multiple base stations is aggregated. In the uplink toward the CO, a WDM-PON is employed so that legacy fiber infrastructures can be utilized.

The metro network domain (elastic frame-based WDM domain) provides connectivity between the access branches and the regional/core data centers (**DCs**) as well as to the global Internet (Fig. 2.35). Current metro solutions deploy dense wavelength-division multiplexing (**DWDM**) in metropolitan areas. Up to 100 wavelength channels can be transported in parallel over the same fiber to accommodate the large aggregated traffic from the access branches. However, the huge capacity of urban DWDM networks reduces the flexibility of the structure used. Expansion of the capabilities of optical networks can be achieved using optical switches (WSS) with wavelength selectivity and reconfigurable optical add/drop multiplexers (**ROADMs**). At the same time, DWDM also allows different virtual network topologies (e.g., ring, bus, and tree) to be used on the same fiber infrastructure.

Let's consider the possibilities of using the tools available in the physical network to control the management and coordination of the mobile network for various interaction scenarios and user characteristics. Architecture of fixed network for 5G suggests five categories of computing resources: central, regional, border (remote) cloud and cloud nodes, as well as physical network functions (PNF).

A central cloud node lives in a centrally located data center. Such a center hosts a large collection of processing, storage, networking, and other fundamental computing resources.

A regional cloud node is available in densely populated metropolitan, urban, and sub-urban areas, e.g., attached to a metro ring.

An edge cloud node is implemented inside an access branch of the fixed network, serving, e.g., a city quarter, and thus even closer to the end user. In typical deployments, it would be situated inside the CO.

A cloudlet (nano-cloud) is a mobility-enhanced small-scale cloud data center that is located at the edge of the network co-located with the macro-cell sites. In the case of active deployment based on the active remote node (**ARN**), both the cloudlet and the macro-cell sites would be co-located with the ARN.

Central, regional, edge clouds, and cloudlets provide physical resources to execute VNFs for management and orchestration. From the core downward to the fixed access domain, central, regional, edge clouds, and cloudlets grow in heterogeneity of hardware and hypervisor; the geographical deployment and topological structure become more complex as well.

Identifying the required mix of VNFs and physical network functions (PNFs) in the 5G network architecture and implementing them exemplarily in a representative scenario are critical research topics on the path toward successful 5G deployment.

5G requires a very flexible architecture for implementing the different scenarios and specifically the *dynamic Cloud-RAN*. In the following, it is described exemplarily how the whole 5G mobile network could be organized as a dynamic Cloud-RAN with distributed network functions.

Investigations indicate that small cells can be efficiently controlled by the superordinate macro-cells, with best results in ARN-based deployments. In this case, the mobility management and handover functions could be implemented as VNFs (eventually including PNFs) in the cloudlet co-located with the macro-cell site.

Radio channels, depending on the tariff of a particular user, can be organized (Fig. 2.38), for example, when using an optical network node (ONU) via PON with an edge cloud at the central office. The base stations of the cell with a small coverage radius are connected to the ONU using a fiber connection and then over the air with the mobile UE. Small businesses with fewer than 500 employees also connect to the ONU via an optical line and the home network via a copper cable (twisted pair).

With a more complex fixed network with a large number of users, when the ONU is replaced by an active remote ARN node, performing the functions of MP2MP switch, it will allow connecting, in addition to the specified users; it can form a network with the edge cloud and a collective cable television (CAT), organize megacots, and wirelessly connect the small cell site to ARN.

Transport networks of mm wavelengths are organized separately, forming transport channels to the backhaul link sensors and to the fronthaul link actuators. Base stations mmWave links are connected wirelessly to BS macro-cell site.

The described local fixed networks are embedded in regional or metropolitan cloud structures interacting with the terrestrial public switched telephone network, the Internet, and the central cloud.

Checking Questions

1. Enumerate main wireless network and their features.
2. Which communication systems are called “mobile”?
3. What is the difference between concepts “radio network” and “radio system”?
4. Denote the main difference between systems of radio communication and radio access.
5. What is the reason of appearance of the wideband networks of radio access?
6. Can we attribute the system of the GSM system to the systems of radio access?
7. Which are the principal differences between UMTS and LTE systems?
8. Explain what the spreading and scrambling sequences are used in IS-95 and CDMA2000 systems?
9. Denote approaches for spectrum spreading used in systems with coding channel division, and explain principles of noise-like signal formation.
10. What is the difference of systems with channel switching from systems used in the packet switching?
11. What are the differences in architecture of network of standards 2G, 3G, and 4G?
12. What are the differences of frequency duplexing and frequency channel division?
13. Explain the reason of intensive development of wideband networks of radio access and the differences of WLAN and WWAN.
14. Explain according to which criterion can we select the spreading sequences, for instance, Walsh functions, in systems with the code channel division.
15. Explain the features of the group signal in systems with the OFDM technology.
16. What are the main differences of FDM and OFDM technologies?
17. What do “the signal base” and “spreading coefficient” mean?
18. Why is spectrum aggregation actively used in LTE-A and 5G networks?
19. Are cognitive networks an evolution of LTE-A networks or a new direction in networking? Will the requirements for terminal devices (UE) change in such networks?
20. What caused the refusal of OFDM technology in the 5G system?
21. Why does 5G not use WWANs for smart home networking, but creates a separate WPAN (Internet of Things)?

Appendix

Application 2.1: Having studied Sect. 2.5.1.2, make use of the expression for the signal base (B) (the spectrum spread factor (SF))

$$SF = B = \frac{\Delta f_{mes}}{\Delta F_{mes}},$$

where $\Delta F_{mes} = 1/T_0 = 1/3.69 \mu s$ is the message bandwidth, which for the GSM900 standard is determined by the bit length $T_0 = 3.69 \mu s$, and $\Delta f_{mes} = 271 \text{ kHz}$. The spectrum width of GMSK signal (Δf_{mes}) determined by the symbol duration of the radio signal coincides with the bit duration T_0 : $\Delta f_{mes} = 1/T = 1/T_0 = 271 \text{ kHz}$.

The rate of the informational sequence coincides with the rate of symbol sequence of the physical channel: $R = (1/T)/(1/T_0) = 1 = B = SF$.

Application 2.2

Calculate the spreading factor (SF) of the signal for the system IS-95 under the condition of the speech signal transmission with the rate $R = 19.2$ kbit/s. Where is the signal noise immunity higher? Explain the reason of this phenomenon.

Having studied Sect. 2.5.2.1, make use of the expression for the signal base (B) (the spectrum spread factor (SF)).

$$SF = B = \frac{\nu}{R} = \frac{1/\tau_{el}}{1/T_0},$$

where the rate of informational video pulses is $R = 1/T = 19.2$ kbit/s. In the IS-95 system, the pulse duration in the radio channel after video pulse expansion by the Walsh sequence with duration of 64 symbols is $\tau_{el} = \frac{T}{L} = \frac{1}{RL}$, and the rate of elementary pulses (chips) is $\nu = 1/\tau_{el}$.

The expansion coefficient is

$$SF = B = \frac{\nu}{R} = \frac{19,2 \cdot 10^3 \cdot 64}{19,2 \cdot 10^3} = 64 \approx 18dB.$$

An increase in the width of spectrum of the radio signal in comparison with the spectrum of the same video signal in base band, in the example under consideration, by 64 times, significantly reduces the effect of interference at the receiver input and, above all, discrete interference characteristic of mobile communication systems.



Chapter 3

Structural Diagrams of Radio Receivers

We begin our serious acquaintance with the various radio receiving devices in this chapter from consideration of the most typical and widely distributed architectures of structural diagrams. For a start, we examine receivers for fixed radio communication systems (Sect. 3.1), at that, architectures of retuned radio front-ends of receivers are discussed in detail separately and then for receivers without retuning of input filter to the receiving signal.

Section 3.2 is devoted to receivers for mobile radio communication systems. We consider the architectures of transceivers for digital mobile radio communication systems, features of super-heterodyne types of receivers, receivers of direct conversion, receivers with digital intermediate frequency, receivers with conversion to low IF. Multi-range and multi-standard receivers are considered separately.

The most detailed Sect. 3.3 describes the manifold receivers for different standards. Among them, super-heterodyne receivers for radio stations of GSM standard subscribers, receivers with the GSM technology, receivers with the CDMA technology, receivers for mobile multimedia radio broadcasting systems, receivers of WiMAX and DECT standards, receivers for the IEEE 802.15.4 standard for the ZigBee protocol, trunking receivers, receivers for LTE systems, receivers of ultra-wideband systems, navigational consumer receivers, receivers for bio-medical investigations, differential quadrature receivers for mobile detection systems.

This chapter is the useful reference book for different types of modern radio receivers.

3.1 Radio Receivers of Fixed Communication

Radio receivers for systems of fixed communication are historically the first type of distributed radio communication systems, and we begin consideration exactly from them as from the most simple. But we need not to think that all problems of the radio receiver construction are already solved: the permanent development of technologies and semiconductor materials with various properties continually opens new possibilities for utilization in the modern radio electronics and, in particular, in the theory and practice of radio receiving devices. The results described in this section are important for students of modern universities as basic information about radio receivers and how the symbiosis of knowledge and modern technology has made these devices compact, multifunctional with low power consumption. The use of digital modes of modulation in the formation of a radio signal and preserving the basic principles of constructing receivers, has revolutionized the methods of signal processing, increasingly replacing circuit design methods with algorithmic ones.

3.1.1 Structural Receiver Diagrams with Front-End Filter Tuning

3.1.1.1 Tuned Radio-Frequency Receivers

It was historically happened that, after invention of amplifying electronic tubes in 1913, the receivers of the *direct amplifier* (Fig. 3.1) had got the most distribution. In such a receiver, the analog signal processing is performed on the frequency of received signal (amplification, filtering). At necessity of receiver operation in the frequency range (variation of receiving signal frequency), the simultaneous and similar frequency retuning of all amplifying stages is performed.

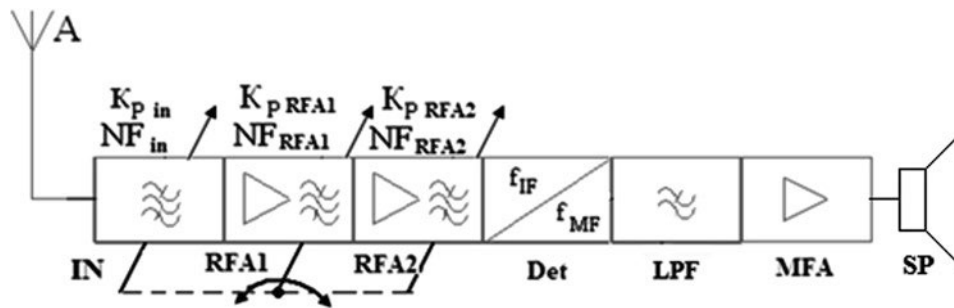


Fig. 3.1 The structure of direct amplifier receiver

Since in the range of low frequency (MW, LW of radio broadcasting range) external noises essentially exceed the natural receiver noises, then the number of stages providing the required signal level on the detector input is not large. Therefore, to retune simultaneously and similarly several circuits of the radio-frequency amplifier stages is comparably simple task. Practically, all receivers before 1929–1932 were constructed on this principle.

The input circuit (IN) usually contains the single oscillating circuit tuned on the frequency of received signal and has the power gain $K_{P,IN}$ and the noise factor (NF). The input circuit must provide the power transfer from an antenna and extract of some frequency band, which contains the useful signal. The input circuit implemented as the single oscillating circuit or as the system of double coupled circuits must be retuned similarly and simultaneously with the radio-frequency amplifiers. At transfer into the higher frequency range, this approach becomes more complicated for realization from technical point of view.

Amplifiers of radio frequencies (RFA) contain the oscillating circuits in its load, which are tuned on the frequency of the received signal and together with the input circuit ensure the required level of the useful signal on the detector input and suppression of the spurious signal of the adjacent (in frequency) transmitter (the adjacent channel, AC).

The detector (Det) provides the *conversion* of the received signal forming the output voltage, which value varies according to the modulation law of some parameter (the most often, the amplitude) of the input signal. The useful component in the output current of the detector with the modulation frequency of the transmitter carrier is allocated by a low-pass filter (LPF).

The amplifier of the modulation frequency (MFA) (the term “base band” is accepted in the literature for mobile communication) provides the required signal level on the input of reproducing device (for instance, the receiver loudspeaker) with permissible distortions (the coefficient of harmonics).

To the main shortcomings of the direct conversion receivers, we can nominate:

- Difficulty in providing the constant sensitivity in the wide frequency range.
- Impossibility to obtain the high sensitivity in the microwave range.
- Impossibility to obtain the constant pass-band width and high selectivity in the wide frequency range.
- Increase of the self-excitation danger in the radio front-end with operating frequency growth.

Mentioned shortcomings are mainly caused by necessity of retuning of the main stages of the receiver, which define its amplification and selectivity at operation in the frequency range.

With improvement of electronic engineering (appearance of high-frequency amplifying tetrodes, pentodes, and the assimilation of higher frequency ranges) from 1930 led to active implementation of radio receivers, which used the super-heterodyne principle.

3.1.1.2 Receivers of the Super-Heterodyne Type

Receivers with the single-frequency conversion. Radio receiver (RR) of the *super-heterodyne* type had got the wide application because the better technical indices could be arranged by simpler means compared to other structural principles.

The structural diagram of such a type of the receiver is shown in Fig. 3.2.

The operation principle of the super-heterodyne receiver is based on the fact that after preliminarily filtering, the receiving signal is transferred into the frequency region, where it can be processed most effectively on the fixed frequency (usually, lower compared to frequency of receiving signal).

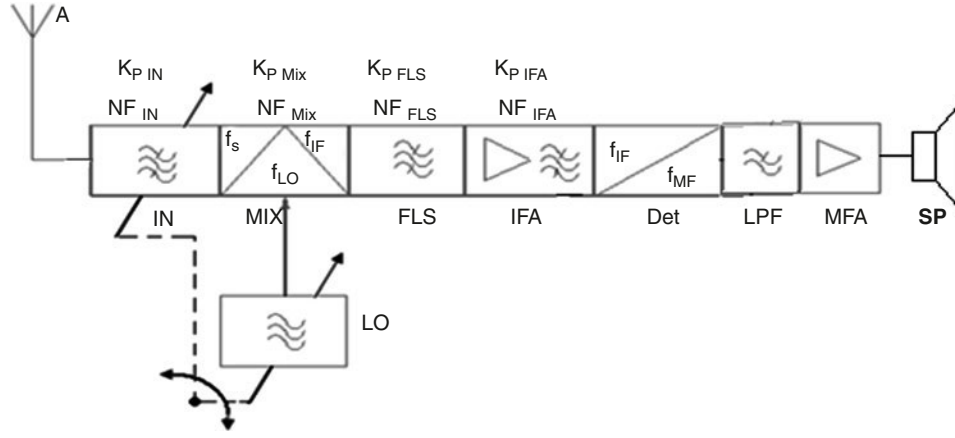


Fig. 3.2 The structure of the super-heterodyne receiver

The input circuit realized as the linear two-port is the device coupling the antenna-feeder system with the input of the first receiver stage. Preliminary filtering is performed into the input circuit (**IN**); over the **IN** in Fig. 3.2, we designated the transfer coefficient in power ($K_{P,IN}$) and the noise factor (NF_{IN}); this circuit may include the low-noise amplifier (**LNA**) of the radio-frequency. The voltage containing the receiving signal together with the variety of others, which fall into the pass-band of the input circuit, passes to the input of the frequency converter (**FC**) consisting of the mixer (**Mix**) and the local oscillator (**LO**). The mixer contains the nonlinear element (**NE**) or the parametric element, in which the *multiplication of voltages* of the input signal $u_s(f_s, t)$ and the local oscillator voltage $u_{LO}(f_{LO}, t)$, providing oscillations with *combination frequency* at its output (for harmonic impacts):

$$f_{FC} = |mf_{LO} \pm nf_s|, \quad (3.1)$$

which are created by the useful signal components with frequencies nf_s and harmonics of the local oscillator frequency mf_{LO} (where $m \in \mathbb{Z}$, $n \in \mathbb{Z}$). The voltage of the single combination components, which is called *intermediate frequency*, having the constant frequency f_{IF} (the most often at $m = 1$ and $n = 1$), is extracted from the mixer output current, which contains the large number of conversion products, with the help of the filter of lumped selectivity (**FLS**) at its output:

$$f_{IF} = |f_{LO} - f_s| \approx \text{const} \quad (3.2)$$

where we specially note that f_{LO} and f_s must be tuned simultaneously and then is amplified in the intermediate frequency amplifier (**IFA**). At retuning in the range, the frequency constancy of the useful signal voltage ($f_{IF} \approx \text{const}$) is provided by the similar and simultaneous variation of the resonance frequencies of **IN** circuit and the local oscillator.

Conversion of the useful component of the input signal should be performed with conservation of the variation law of the modulating parameter (amplitude, frequency, phase), i.e., its spectrum is transferred without variation of the ratio between spectral components into new frequency region.

The main advantages of super-heterodyne types of receivers:

1. Oscillating circuits including in the intermediate frequency section do not retuned at receiving the signals of different frequency from the range of the operating frequencies of the receiver; this allows receivers designing for the new frequency range changing only the adjustment of the input circuit and the local oscillator.
2. Owing the fixed tuning of circuits of the IF section, the amplitude-frequency response (AFR) of the whole radio section remains practically unvaried and defined by the AFR of the FLS at **IN** circuit retuning on the new operating frequency, providing with its help the required value of adjacent channel suppression;
3. Choice of the intermediate frequency, which is much less than the frequency of receiving signal, increases the amplifier stability of the IF section because of decrease of spurious feedback influence, including in the active element of the amplifying stage, and, relatively, increases the stable gain value.

Mentioned advantages of super-heterodyne type of receivers are the reasons of their wide applications in the radio communication technique. The higher sensitivity of receivers, the stability and reliability of receiver stages, etc. are provided with their help. However, introduction of the mixer in the radio front-end, which is the nonlinear or parametric devices, led to

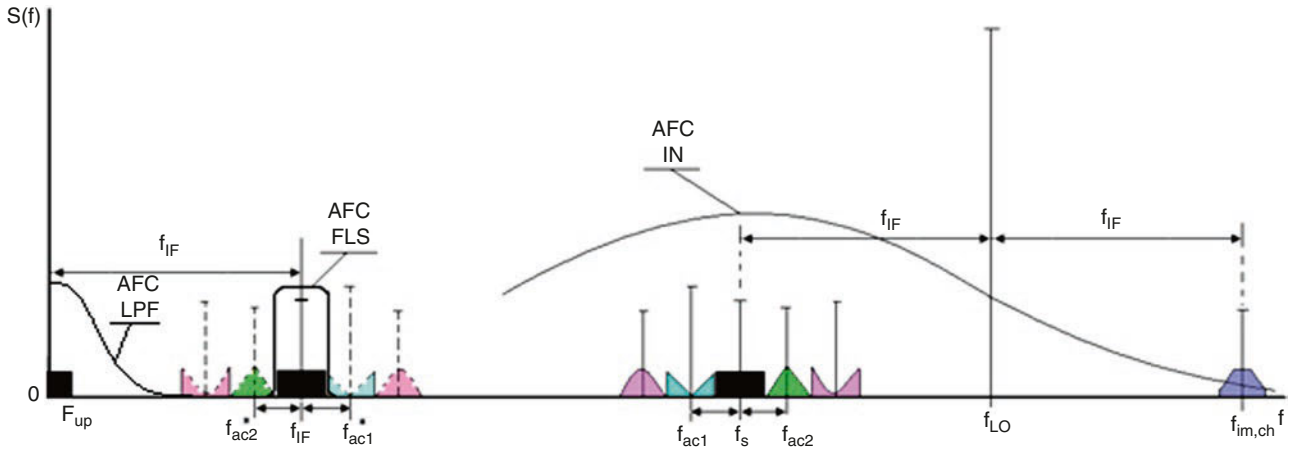


Fig. 3.3 Interferences in the adjacent channel

appearance of collateral channels (**CRC**) caused by the voltage multiplying procedure of the RF signal and the local oscillator signal (and their harmonics). Let us consider the main reasons leading to appearance of the similar interferences at operation of the receivers of the super-heterodyne type and approaches for struggle with them.

Collateral reception channels. At the input of the receiver of super-heterodyne type, as in any other receiver, besides the EMF of the useful signal, signals act, which are created in the antenna due to the electromagnetic field conversion from other acting transmitters. *Impacts from other transmitters which are nearest* in frequency are the lumped interference for the useful signal and are called the *adjacent channel (AC)* (Fig. 3.3, f_{ac}).

The position of adjacent channel (AC) in the range of received frequencies for frequency division (**FDM**) receivers is most often determined by the bandwidth allocated to the working channel. For radio broadcasting receivers, for example, with the amplitude modulation (AM) [1], the channel are considered as the adjacent at offset with regard to the useful one by ± 9 kHz while for receivers with frequency modulation by ± 250 kHz. The single-signal selectivity Se_{ss} defines the degree of its suppression. At signal reception with other type of modulation, the adjacent channel is considered as shifted from the central frequency of the received signal by the bandwidth occupied by the signal.

At *upper heterodyne (LO) tuning* ($f_{LO} > f_s$) and the input impact containing of the large number of signals with the restricted spectrum (Fig. 3.3), the useful signal (the carrier on the f_s frequency) is transferred on the f_{IF} frequency. Voltages from other transmitters at the receiver input, which fall in the pass-band of the input circuit (the pass-band is usually estimated on the level of 0.707 of the normalized value of the filter transfer coefficient $B_{0.7}$), after multiplication with the local oscillator voltage, are also transferred into the new frequency region. At that, both the useful signal spectrum and spectra of accompanied channels is exposed the inversion (the upper side band becomes lower, while the lower becomes upper). The useful signal spectrum is extracted by FFS having the pass-band somewhat bigger (approximately by 10%) then the frequency band occupied by the useful signal. Other channel spectra are suppressed. As we see from Fig. 3.3, the *AC suppression* in receivers of the super-heterodyne types is performed in the *IF strip*. It is much simpler to obtain the AFC with the large squareness coefficients ($K_{sq,0.001} = 1.5K_2$) on the intermediate frequency f_{IF} using the distributed or lumped filtering.

The main amplification and extraction of the useful signal (suppression of adjacent stations) occurs in the intermediate frequency (IF). To suppress the *adjacent channels*, we may use the approach of *distributed filtering*, which is typical for the broadcasting receivers of the zero and first groups of complicity, when each of IFA stages is loaded on the single or on two coupled oscillating circuits (usually at critical coupling between circuits). There are various stage creations on circuits tuned on f_{IF} or detuned with regard to f_{IF} (with single circuits or detuned pairs and three circuits). In this case, AFC of the linear section can be realized as the maximally flat AFC or with permissible variations of the transfer function within the limits of the bandwidth at high linearity of the phase-frequency characteristic (PFC). At high requirements to sensitivity, the IF section contains several amplifying stages, which complicates its technical implementation and decreases the stability of separate stages and the radio front-end as a whole.

The approach of *lumped filtering* (Fig. 3.2) provides the suppression of the *adjacent channel* in the IF section using of multi-chain LC filters, piezo-ceramic, electro-mechanical, or piezo-mechanical miniature filters included as a load of the frequency converter or in the first stage of IFA. Intermediate frequency amplifiers with the filter of lumped selectivity (**FLS**) are used in those cases when it is necessary to provide the high K_{sq} at small number of stages. A multi-chain filter included at the mixer output or in the load of the first stage of the intermediate frequency path amplifier (IFA), in addition to suppress-

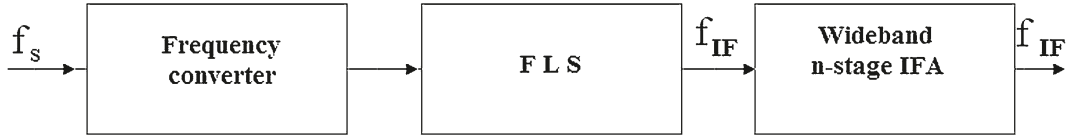


Fig. 3.4 The structure of multi-chain FLS in the IF strip

ing the adjacent channel (AC), provides, on the one hand, matching the characteristic resistance FLS with the output resistance of frequency inverter (FI), and on the other with the IFA input (Fig. 3.4).

Additional IFA stages providing the necessary amplification of the radio front-end are performed as wideband and practically not-influenced of the AFC in the pass-band of the IF section.

The expediency of application of multi-circuit FLS in the IF section is determined from the relation:

$$d_{\text{con}} \leq B_{\text{RFE}} / 2\sqrt{2}f_{\text{IF}}, \quad (3.3)$$

where B_{RFE} is the pass-band of the radio front-end (the IF section), f_{IF} is the nominal tuning frequency of FLS (the intermediate frequency), and d_{con} is the own (constructive) damping of each FLS circuits.

The modern realization of FLS is the choice of piezo-ceramic, quartz, and other types of the miniature filters. Piezo-ceramic FLS has the small dimensions, the low mass, and the stable AFC and allows simplification of construction and adjustment of the receiver. Application of the lumped filtering by FLS installation on the output of the frequency converter essentially simplifies the amplifier construction in the IF section. Amplifiers can be implemented with application of resistor stages in the larger part of the IF section besides the last. The last stage of the IF section is usually realized using the transformer coupling with the detector. Most often, the resonance stages are used in the IF section with the single oscillating circuit in the load, which has the low Q-factor, which decreases the linear distortions of AFC of the radio front-end and permits to obtain the more gain per stage. Simultaneously, this decreases the AFC overshoot values beyond the FLS pass-band, which reduces the self-excitation danger on the spurious frequencies (out-of-band AFC growths). If the relation (3.3) is not fulfilled even at some variation of the constructive circuit Q-factor, then the IF section is performed on the base of the distributed filtering.

Advantages of super-heterodyne receivers are added by shortcomings, which are peculiar to receivers of such a type only.

Structural diagrams (Fig. 3.2) of receivers contain at their input the filtering circuit (the input circuit) providing the reduction of interference influence of various nature and the characteristic of the impact. The *most important* function of the IN chain circuit is the *suppression of spurious channels*, which arises as a *consequence of frequency converter* usage in receivers of the super-heterodyne types.

If there is no preselector of the mixer (Fig. 3.2), the effect of discrete interference with frequencies at its input

$$f_{\text{IF}}, f_{\text{LO}} + f_{\text{IF}}, 2f_{\text{LO}} \pm f_{\text{IF}}, 3f_{\text{LO}} \pm f_{\text{IF}}, \dots \quad (3.4)$$

leads to appearance of component with an intermediate frequency f_{IF} in the output current. These discrete interferences are called collateral reception channels f_{col} (at upper tuning of the local oscillator, $f_s < f_{\text{LO}}$).

This leads, as it is shown in Fig. 3.5 as an impact of the useful signal with the frequency f_s , to appearance at the *output* of the frequency converter of the *interference frequency voltage* f_{IF} .

Only one of the influences, at the frequency of the useful signal f_s of all received $f_{\text{rec}} = f_{\text{LO}} - f_{\text{IF}}$ in this case, which is the difference frequency of the local oscillator and the signal, forms a component at the intermediate frequency $f_{\text{IF}} = f_{\text{LO}} - f_s$ (down converter) from this effect f_s (Fig. 3.3). The nearest component from the signal frequency mentioned in (3.4) is located symmetrically with regard to f_{LO} , which is the impact (interference f_{int}) on the frequency $f_{\text{int}} = f_{\text{im, ch}} = f_s + 2f_{\text{IF}} = f_{\text{LO}} + f_{\text{IF}}$ called the frequency of the *image channel*. This component creates the greatest danger (Fig. 3.5), since when multiplying the voltages LO with frequency (f_{LO}) in mixer with the voltage signal frequency ($f_s = f_{\text{LO}} - f_{\text{im}}$) or with the spurious channel ($f_{\text{im, ch}} = f_{\text{LO}} + f_{\text{im}}$), a difference (intermediate) frequency is created from the useful signal $f_{\text{im}} = f_{\text{LO}} - f_s = f_{\text{LO}} - f_{\text{LO}} + f_{\text{im}} = f_{\text{im}}$ and from the “mirror channel” signal $f_{\text{im}} = f_{\text{im, ch}} - f_{\text{LO}} = f_{\text{LO}} - f_{\text{im}} - f_{\text{LO}} = f_{\text{im}}$. At the same time, the separation of intermediate frequency voltages obtained from different sources using FLS is impossible due to their equality. The suppression degree of the spurious impact on the mixer input is defined by AFC on the input circuit filter. We can neglect the influence of spurious conversion channels of higher order due to significantly higher attenuation compared with the image channel.

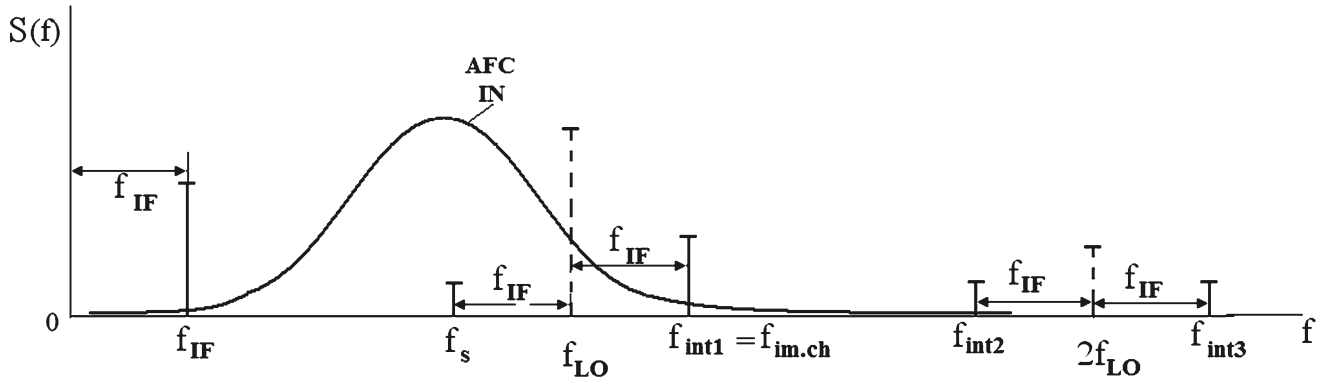


Fig. 3.5 Position of collateral channels in a super-heterodyne receiver (Fig. 3.4) under the condition ($f_{LO} > f_s$)

Another, very dangerous spurious channel, which exists only in receivers of the super-heterodyne type, is the channel of *direct passage* $f_{rec} = f_{int,dp} = f_{IF}$ ($f_{int} = f_{rec}$). In this case, the mixer operates as the amplifier of the input impact with the $f_{rec} = f_{IF}$ frequency and simultaneously as the converter forming frequencies $f_{LO} \pm f_{IF}$ which do not fall in the LPF pass-band.

Another collateral conversion channels on frequencies $f_{col}; f_{int2,3} = 2f_{LO} \pm f_{IF}, f_{int4,5} = 3f_{LO} \pm f_{IF}$ usually do not require additional measures for their suppression, since the preselector filters (IN circuit) ensure their minimal influence on the receiver operation.

For the *linear mode* of frequency converter operation on *signal*, when the signal frequency harmonics do not arise ($n = 1$), the number of collateral channels reduces to two ($f_{im, ch}$ and f_{rec}) compared to the nonlinear mode ($n \neq 1$). The *influence of collateral conversion channels* is manifested in the form of beats between closed in frequency oscillations in the IF section from converted input impacts (for instance, f_s and $f_{im, ch}$), of listening of spurious stations simultaneously to the useful signal, the mutual signal modulation falling in the receiver input, etc. The absence of spurious stations on frequencies of collateral channels does not create mentioned distortions, but decreases the factual receiver sensitivity due to power increase of resulting external noise on the detector input. This is a consequence of noise transfer from the frequency band of collateral channels into the intermediate frequency. Reduction of distinguishability in the detector input ($D_d = P_s/P_n$) is defined by the noise intensity and the selective properties of the input circuit and the IF section.

In receivers with the single conversion (Fig. 3.2), the intermediate frequency value is always constant, and it is chosen, as a rule, in frequency regions free from operating radio stations. So, for the automobile radio broadcasting receivers [64], the list of recommended intermediate frequencies is 0.070 ± 0.002 ; 0.076 ± 0.002 ; 0.465 ± 0.002 ; 0.468 ± 0.002 ; 1.840 ± 0.005 ; 2.900 ± 0.010 ; 10.700 ± 0.010 ; and 24.975 ± 0.100 MHz. Existing potential danger of falling in the mixer input of the interference with the frequency $f_{int} = f_{IF}$ requires to take the constructive measures for its level reduction up to safe value, which is defined by standards. Taking into consideration that the *channel of the direct passage* ($f_{dp} = f_{IF}$, Fig. 3.5) in HF and VHF ranges is far enough from the useful signal frequency, it is easily suppressed in the IN circuit. If the input circuit does not provide of required suppression of the direct passage channel (Se_{dp}), then it is achieved with the help of simplest notch filter, which is included before the mixer. Possible variants of implementation of the preselector circuits (front-end) with the additional notch filter tuned on $f_{int} = f_{IF}$ are shown in Fig. 3.6.

Figure 3.6 shows the electrical diagrams of front-end containing the input circuit and the RFA stage with tuned single circuits with the auto-transformer coupling with the active element. The front-end coupling with an antenna is chosen as external capacitor realized with the help of the capacitor $C_{con A}$. The suppression of the direct passage channel ($f_{col} = f_{IF}$ in the circuit Fig. 3.6a) is provided by the parallel circuit ($L1, C1$), tuned on f_{dp} and connected in series with the parallel circuit ($L2, C2$). The circuit $L2, C2$ is tuned on the frequency of useful signal and has the small transfer coefficient on the frequency $f_{dp} = f_{IF}$. At that, depending on its Q-factor, the circuit $L1, C1$ can have the low enough losses on the frequency of the useful signal.

In the circuit in Fig. 3.6b, the notch filter on the series circuit $L1, C1$, tuned also on f_{dp} , is connected in parallel to the input of the RFA stage, and having the small resistance on the resonance frequency, it shortens the input signal to the ground on the f_{dp} frequency.

Linear distortions in the RF front-end. The front-end in the region of low and medium frequencies is usually realized as the passive circuit (input circuit), to which the frequency converter is connected (RFA is absent). Realization of the input circuit in the form of the single oscillating circuit (SOC) permits to provide the image channel suppression according to requirements of standards on radio engineering equipment by a choice of its loaded (equivalent) Q-factor ($Q_{eq, im, ch}$). At that, the value of the loaded Q-factor (taking into account the inserted losses in the circuit from the external circuits) should be less than 80. Otherwise, this caused difficulties at its technical realization. Simultaneously, at that, we should provide the

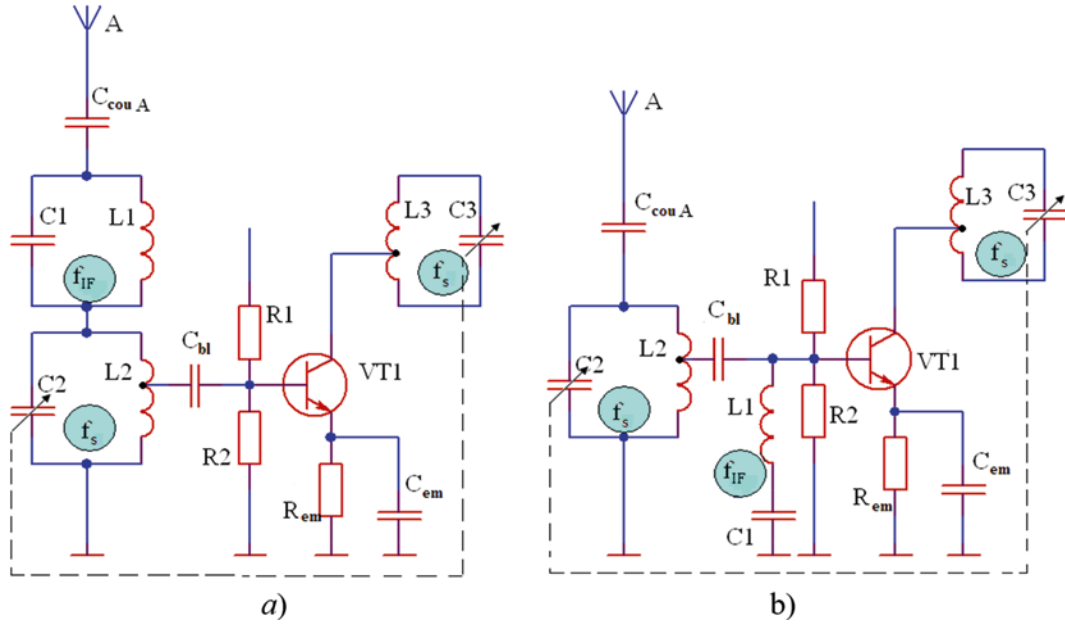


Fig. 3.6 Possible realizations of front-end containing the input circuits and the RFA stage with tuned single circuits with the parallel circuit (a) and series circuit (b)

value of linear distortions (irregularity of AFC) in the input circuit not more than it is defined in the standard to the equipment. This leads to the necessity of reduction of the loaded Q-factor ($Q_{eq,l}$) at irregularity of the front-end AFC.

At signal reception in LW, MW, and SW with AM, it is necessary to take into consideration that the irregularities of the truth curve $M = 14$ dB are summed from irregularities caused by the linear section (M_{lin}), by the detector (M_d), by the output signal amplifier (M_{ampl}), and by the loudspeaker (M_{loud}):

$$M = M_{lin} + M_d + M_{ampl} + M_{loud} \quad (3.5)$$

At providing the required selectivity, the irregularity of the linear section (the radio frequency front-end) is desired to be $M_{lin} = 9 \dots 13$ dB, and from, this $M_d + M_{ampl} + M_{loud} \leq 5$ dB. The loudspeakers introduce the significant contribution in the linear distortions, which create AFC distortions up to 14 dB, and therefore, to reduce irregularities created by the loudspeaker, one can take special measures. For this, the loudspeakers with the wider bandwidth are used, the amplifiers are constructed with the low output resistance, and recently, the most often used measure is application of several loudspeakers, which reproduce the only part of the audio range. Utilization of the tone quality regulators, which increase amplification in the region of upper frequencies, also reduces the linear distortions. The described measures should provide the resulting distortions (according with requirements of standards for radio broadcasting receivers) in the band of the linear section $M_{lin} = 9$ dB, at radio signal reception on frequencies upper than 250 kHz and $M_{lin} = 13$ dB at operation on frequencies lower than 250 kHz.

The irregularity of frequency response of the RF front-end (linear parth) includes $M_{lin} = M_{pres} + M_{IFA}$, where M_{pres} is irregularity created of preselector, and M_{IFA} is the irregularity of frequency response of the IF strip of receiver. In the area of moderate-high frequencies, we usually choose $M_{pres} = 3$ dB, and $M_{IFA} = 6$ dB to obtain the larger attenuation of adjacent channel signals and to stability improve of the truth curve. For radio broadcasting receivers of the LW range, it is admissible the requirement reduction of preselector AFC irregularities up to $M_{pres} = 6$ dB, which is connected with difficulties of filter implementation having the low Q-factors. In UHF range, because of the signal reception with wideband FM, the front-end irregularity should be $M_{pres} \leq 1,5$ dB.

The realized value of the equivalent (loaded) Q-factor of a single oscillatory circuit must simultaneously provide the required suppression of the image channel, and the permissible nonformity of frequency response of preselector, that is, fulfills the conditions:

$$Q_{eq,im,ch} < Q_{eq} < Q_{eq,load}, \quad (3.6)$$

where $Q_{eq,im,ch}$ is the equivalent Q-factor providing the suppression of the image channel and $Q_{eq,load}$ is the equivalent loaded Q-factor, which realizing the requirement on irregularity frequency response of preselector. The front-end is usually realized in the form of two single oscillation circuits connected to the input circuits and the RFA (Fig. 3.7).

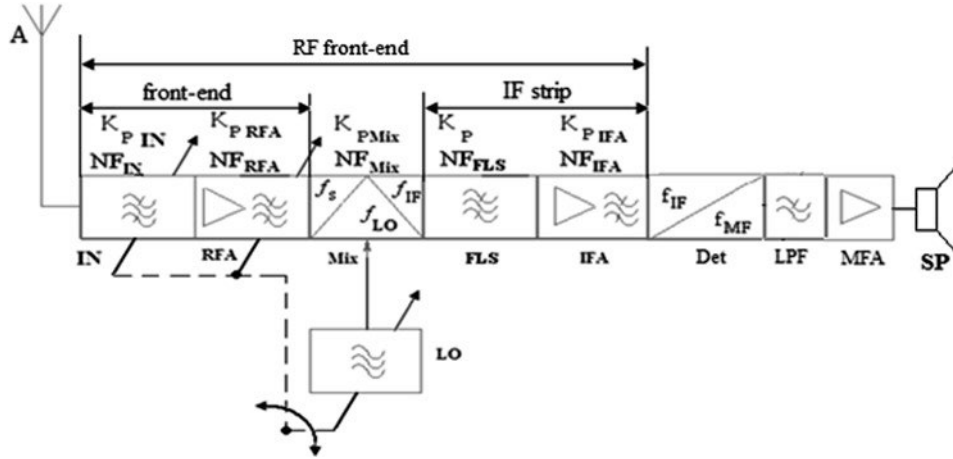


Fig. 3.7 Structural diagram of the super-heterodyne receiver with RFA

At that, as the decoupling element between the single oscillating circuits, the AE (bipolar and the field-effect transistors (FET)) are used with the low level of proper noises, making the circuit parameters independent. Insertion of RFA allows reduction of equivalent Q-factor (Q_{eq}) of each used single oscillating circuit, which, according to constructive considerations, are often chosen the same at providing of required suppression of the image channel ($Se_{im, ch}$) and specified irregularity of the front-end (preselector) AFC (M_{pres}). Reduction of the loaded Q-factor simplifies the realization of single oscillation circuit; the resulting AFC becomes more flat, wideband. Insertion of RFA increases the factual sensitivity [73] approximately by the order for AM receivers without degradation of stability. Insertion in the preselector of additional RFA stage improves the noise characteristics of the receiver as the value of the real noise factor now is mainly defined by the RFA noise factor, but not the largest noisy stage of the frequency converter. For receivers of meter and decimeter ranges, insertion of the RFA stage, besides the growth of factual sensitivity and selectivity ($Se_{im, ch}$), permits to increase the local oscillator stability.

In receivers of the third-fourth groups of complexity, the input circuit is usually realized with the non-tuned pass-band due to the small covering coefficient, but somewhat larger than the received frequency range. In receivers of zero and first groups of complexity, the input circuit is performed as narrowband and non-tuned. At that, the RFA is often implemented on the base of the cascode circuit, which improves the local oscillator and the input circuit decoupling because their frequency proximity at absence of RFA causes the power radiation with the frequency of the local oscillator by the receiver antenna. The single oscillating circuits (SOC) in the input circuit and in the RFA load are usually performed as the same, which simplifies the receiver and reduces its cost. If SOC application in front-end does not provide the required $Se_{im, ch}$, it is necessary to use the more complicate filtering system. However, application of double pass-band filters leads to increase of the receiver noise coefficient. The main reason for the exponential growth of mobile and satellite communication system was the appearance on market of field-effect transistors on a silicon substrate, which have a low noise figure and cost compared to BT. Receivers of mass manufacture are usually constructed on the described (Fig. 3.2) structural diagram.

Constructive realization of the preselector (front-end). As it follows from previous considerations, the *main destination of input circuits* in the super-heterodyne receiver is *suppression of the image channel*, which shifted from the useful signal frequency f_s by the value of $2f_{IF}$. In LW, MW, and SW ranges, this is provided most often by application of the single-circuit filters, which is caused by simplicity of input circuit realization and simultaneously reduces the variations of its AFC at tuning in the range, as it occurs at using of coupled circuits. Application of the non-tuned antenna for operation in the frequency range at variations of its parameters as well as input parameters of AE (transistor VT1, Fig. 3.6) at variation of the receiving signal frequency leads to deformation of the input circuit AFC. Variations of AFC at retuning in the range depending on the resonance frequency of the antenna and at the influence of the spurious feedback in the transistor may be unpredictable, which causes troubles at application in the input circuit even double-circuit filters and especially multi-circuit. Another reason, which restrict the application of such a construction of input circuits, is the choice of the coupling value between circuits and values of its parameters (Q , $B_{0.7}$) to achieve the constancy of input circuit indices at retuning in the range.

Implementation of the required value of the image channel suppression is simpler solved by the RFA introduction (single or double) with using of SOC in the load, which simultaneous increases the distinguishability coefficient. However, the number of circuits tuned with the help mechanical capacitors (or inductors) cannot be more than 4–5. This is caused by the strong mutual influence of capacitors belonging to the front-end circuits and the local oscillator, which are constructively performed on the same rotation axis, and the requirement of identity of variations of the variable capacitance at tuning of all

circuits simultaneously. Application of varicaps allows the increase of simultaneously tuned circuits; however, due to parameter variations of the volt-farad characteristics of the varicap matrices and their nonlinearity, the number of tuned circuits are limited by the same values. Besides, for receivers with smooth retuning in the range, the significant influence on front-end properties (the bandwidth, the transfer function, the retuning, the linear distortions) is exposed the problem of *circuit conjugation of the front-end and the local oscillator*.

Shown in Fig. 3.3 is a variant of spectra location of useful and other impacting signals which corresponds to the upper local oscillation tuning ($f_{LO} > f_s$), when the local oscillator design does not cause technical difficulties. The high local oscillator (LO) setting simplifies the implementation of preselector and reduces the sub-range number, into which we must divide the operating frequency range for achievement the required of $Se_{im, ch}$ value, especially when the overlap ratio is essentially large than unit ($k_{cov} \gg 1$), where $k_{cov} = f_{max}/f_{min}$. However, with the growth of operating frequency, for instance, in the system of the mobile communication of the GSM1800 standard, it is technically difficult to provide the required indices of the local oscillator with upper tuning; therefore, we use the lower tuning ($f_{LO} < f_s$). Then the side band inversion is observed in the spectrum of converted signals and frequencies of the image channel $f_{im, ch}$ and the useful channel f_s change places (Fig. 3.3).

Implementation of intermediate frequency section. The complexity of technical realization of the front-end circuits ($Q_{eq} > 80$) for fulfillment of standard requirements on suppression of the image channel at given irregularity assumes the variation of the structural diagram scheme of the receiver (Fig. 3.7) by introduction of the additional RFA in the input circuit, which allows the decrease of Q_{eq} of each circuit including in the input circuit.

At designing of the radio broadcasting receivers, the value of the intermediate frequency is defined by standards and equals to $f_{IF} = 465$ kHz at reception of the AM signals and $f_{IF} = 10.7$ MHz for FM signals. A choice of the standard values of intermediate frequencies allows simplification and cost reduction of the equipment manufacture owing to a possibility of normalized block application and the specialized equipment for technological adjustment and checking.

At absence of rigid requirements on a choice of the intermediate frequency, its value must be chosen from the following considerations.

Transferring the spectrum of the input signal in a superheterodyne type with a single conversion to an intermediate frequency (lower, then the frequency f_s) simplifies the implementation of receiver, reduces the number of amplification stages in RF front-end, increases its stability and reduces cost. The chosen value of the intermediate frequency, which is usually dictated by standards for serial production, should satisfy the requirement of the absence of powerful operating stations near chosen f_{IF} . A choice of higher f_{IF} at fulfillment of mentioned requirements creates troubles in suppression of the adjacent channel. The *up-conversion*, when we choose f_{IF} as the intermediate frequency, which value is upper than $f_{s, max}$, essentially simplifies the struggle with the image channel and other collateral channels of conversion and reduces their number in the operating frequency range. However, in this case, the number of conversions increases, and the task of filters' realization for the first intermediate frequency becomes more difficult.

Rather important, especially, for receivers of the single-band (SB) AM signals, is a choice of the local oscillator frequency. At upper local oscillator tuning, when $f_{LO} = f_s + f_{IF}$, the achievement of specified tuning range $f_{s, min} \dots f_{s, max}$ simplifies, but at that, the spectrum side band inversion occurs of signal frequencies, which must be taken into consideration at further processing. In addition, the absolute instability of the local oscillator frequency increases. Thus, the main reasons leading to application of several frequency conversions are:

- Impossibility of simultaneous fulfillment of the contradictory requirements on image and adjacent channel suppression.
- Impossibility to provide the simultaneous suppression of the image channel at permissible AFC irregularity of the front-end section.
- Requirement for achievement of required accuracy of the frequency tuning.
- The necessity to apply more than five in professional receivers (and three in radio broadcasting receivers) of tuned circuits.

Receivers with double-frequency conversion. In professional receiver at high requirements to image channel suppression, the double-frequency signal conversion is often used (Fig. 3.8).

Application of the double-frequency conversion provides the required suppression of the image channel in the front-end due to utilization of the high value of the first intermediate frequency

$$f_{IF1} > 3f_{c, max}, \quad (3.7)$$

where $f_{s, max}$ is the maximal frequency value of operation frequency range. Such receivers are called *infradynes*. At that, the combinational components of the following types are eliminated: $2f_{s, i}$; $2f_{s, i} \pm f_{c, k}$; $2f_{s, i} \pm f_{s, k}$; $f_{s, i} \pm f_{s, k} \pm f_{s, q}$; where $f_{s, i}$, $f_{s, k}$, $f_{s, q}$ are frequencies of signals, which *do not create the components with the intermediate frequency* due to interaction on the nonlinear element of the front-end.

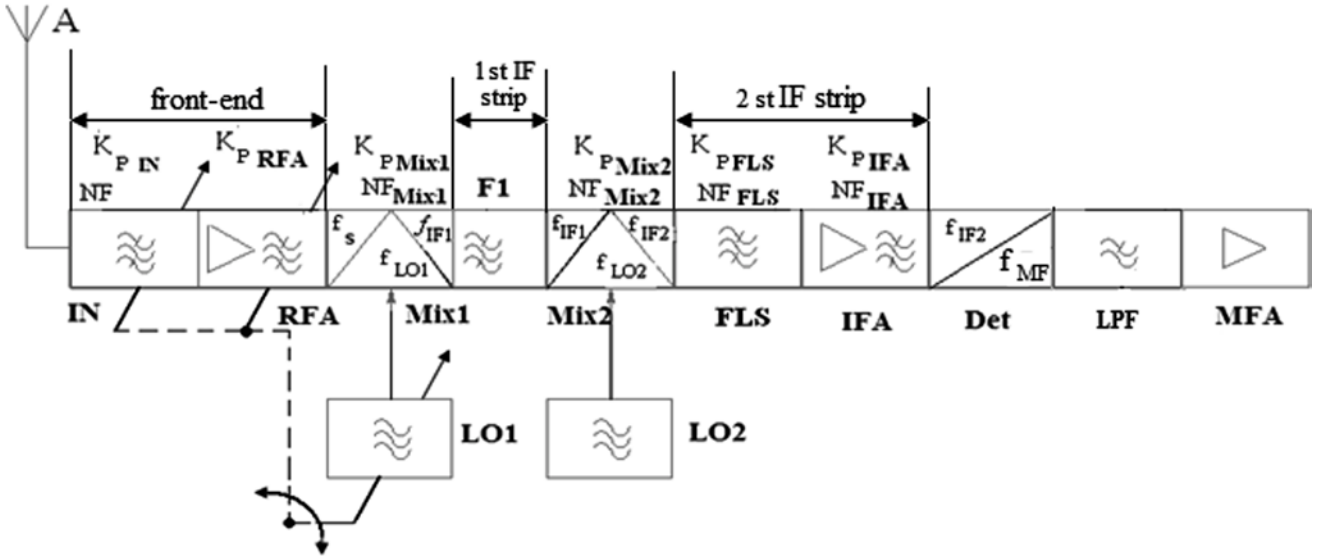


Fig. 3.8 The structural diagram of the super-heterodyne receiver with the double-frequency conversion

In some cases, the f_{IF1} decrease is admissible but not less than $2f_{s\min}$, where $f_{s\min}$ is the minimal value of the operation frequency range.

The frequency of the first local oscillator is chosen from the condition

$$f_{LO1} = f_{IF1} + f_s. \quad (3.8)$$

This allows elimination on the influence of frequency components of the form:

$$2f_{LO} \pm f_{c,i}; 2f_{LO} \pm 2f_{s,i} \pm f_s; 2f_{LO} \pm f_{s,i} \pm f_{s,m} \pm f_{s,q}.$$

To suppress the arising image interference, it is expedient to choose the second intermediate frequency from the condition

$$f_{IF2} > (f_{s\max} - f_{s\min})/2. \quad (3.9)$$

To exclude the spurious influence of harmonics of the second local oscillator, its frequency should be chosen from the condition

$$f_{LO2} > [(f_{LO1} \pm f_s) + f_{IF1}]/2. \quad (3.10)$$

The fulfillment of these conditions will allow avoiding the appearance of false frequencies on the load Mix 1, which are absent in the input signal and are collateral reception channels of reception for second intermediate frequency.

The high first intermediate frequency makes it relatively easy to achieve the necessary $Se_{im,ch1}$ rejection at the first intermediate frequency at the interface without changing the parameters of the filters in the input circuit and the LNA load that were used in single conversion structure. The converted signal on the frequency f_{IF1} is extracted by the $F1$ filter on the output of the first mixer and passes to the input of the second mixer. The second mixer Mix2 performed the transfer of f_{IF1} to the lower second intermediate frequency f_{IF2} , which facilitates the suppression of the adjacent channel and formation of required AFC. The double conversion leads to appearance of the *second image channel* on the frequency $f_{im,ch2} = f_s - 2f_{IF2}$, which is located now symmetrically with regard to the second local oscillator frequency f_{LO2} . Attenuation of the second image channel in the front-end is relatively small (Fig. 3.9), which can be explained by its near location of the useful signal frequency f_s .

The second frequency converter transfers the second image frequency $f_{im,ch2}$ into the frequency $f_{LO1} - f_{im,ch2} = f_{LO1} - (f_s - 2f_{IF2}) = f_{IF1} + 2f_{IF2}$, which should be suppressed in the first intermediate frequency section by the filter $F1$. For chosen high f_{IF1} at low second intermediate frequency f_{IF2} , the suppression of $f_{im,ch2}$ may be not enough, which will require the application of the third frequency converter.

Obviously that fulfillment of (3.7) condition on the operating frequencies of systems of mobile communication (Table 1.2) is impossible; therefore, on the output of the frequency converter FC_1 , the band-pass filter (BPF) is used, which facilitates the suppression of unwanted combination components, and a choice f_{IF1} is provided on the base of recommendations [55]. Application of the nomograms of combination components simplifies the choice at variation of the local oscillator

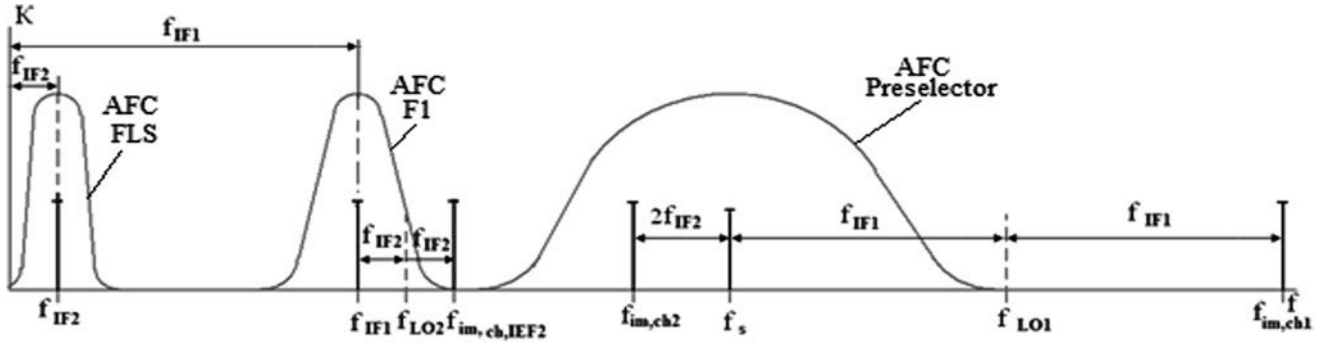


Fig. 3.9 Frequency diagram of the super-heterodyne receiver with double-frequency conversion

frequency, which permits the intermediate frequency choice in the aperture restricted by the bandwidth of converted frequencies of the input signal. Relatively high first intermediate frequency allows performing the necessary suppression of the image channel $f_{im,ch1}$ by the front-end filters. Converted signal on the frequency f_{IF1} then is extracted by the F1 filter in the intermediate frequency section and then it once more down-converted by the converter FC2. The low second intermediate frequency f_{IF2} simplifies formation of required AFC (close to rectangular in shape) and the pass-band of section with the help of FLS. The feature of double conversion consists in appearance of the second image channel $f_{im,ch2} = f_s - 2f_{IF2}$ shifted from f_{IF1} by $2f_{IF2}$ and located symmetrically relative to the frequency of the second local oscillator f_{LO2} (Fig. 3.9).

In the preselector, the second image channel is not exposed of noticeable attenuation because the second intermediate frequency f_{IF2} is relatively low and $f_{im,ch2}$ is located enough close to the frequency of the received signal f_s .

On the output of the first FC1, the frequency $f_{im,ch2}$ is transformed in the frequency $f_{LO1} - f_{im,ch2} = f_{LO1} - (f_s - 2f_{IF2}) = f_{IF1} + 2f_{IF2} = f_{im,ch,IF2}$ which should be suppressed in the first intermediate frequency section by the F1 filter, which is mainly intended for this.

The infradyne advantage consists in simplification of the front-end. So, in the ranging receiver, the tuning can be performed only by tuning of the local oscillator circuit, which essentially simplifies and reduces the price of device construction. In this case, the necessity of application of the switched inductors drops away for the sub-range replacement. This design reduces the number of stages in the front-end and eliminates the use of unreliable and bulky mechanical switches.

Another advantage of the infradyne is the reduction of covering coefficient k_{cc} on the frequency of the first local oscillator, which allows the restriction by the local oscillator circuit tuning (without sub-range switching) at operation in the frequency range. Such a construction of the local oscillator significantly reduces the receiver tuning time on the receiving frequency, which is important for adaptive and automated communication systems. At that, application of wideband front-end containing RFAs essentially increases requirements to linearity of amplifying elements' characteristics at large dynamic range. The choice of preselector structure, a number of signal conversion, intermediate frequencies, and requirements to stability of AFC and PFC are defined with account of the (3.3) condition and the approach to construct the intermediate frequency section.

The approach of lumped filtering of the adjacent channel is rather widely applied in broadcasting receivers including the first group of complexity and permits to obtain the required technical indices creating the compact devices. If the receiver has the increased requirements on the level of phase distortions (e.g., in receivers of multi-channel messages), the FSL application can lead to distortions of the primary signal shape.

In receiver with adjustable bandwidth as well as for reception of radio broadcasting programs with FM, the intermediate frequency amplifiers (IFAs) with the pass-band filters provide the required suppression of the adjacent channel.

IF amplifiers with the single-circuit stages tuned on the single frequency are characterized by the simplicity in adjustment, the good transient and phase characteristics, and the stability of AFC and PFC at variation of AE parameters. At that, the smooth (in shape) resulting AFC of IF section has no enough squareness and amplification area. Therefore, such amplifiers are used at small amplification and small requirements to squareness.

Amplifiers with double-circuit pass-band filters (DPSF) have more squareness and high stability of AFC. Increase of coupling between circuits (more than critical) increases the AFC squareness, but at that, the adjustment becomes complicated and AFC, and PFC dependence on AE parameters grows.

Amplifiers on pairs and triples of detuned circuits are used only in wideband sections ($B/f_{IF} > 0.1$) with high amplification (more than 10^5). Such chains of amplifiers are complicate in adjustment and are exposed by the strong influence of AE parameter variations.

3.1.1.3 Receivers of Direct Conversion

In systems of fixed communications, the receivers find application, which use the principle of super-heterodyne type, but at choice of the *local oscillator frequency equaled to the frequency (and the phase) of the received signal*, when the signal frequency conversion is performed on the zero frequency (Fig. 3.10). Such a type of receivers is called the *homodyne*.

The selector realized with application of the input circuit of the resonance type and the radio-frequency amplifier with the resonance load (usually, the single oscillating circuit) provide the preliminary filtering of the useful signal on the background of interferences. On the mixer output realized as the multiplier, the informational signal with the modulation frequency is formed, which is extracted with the help of the low-pass filter (LPF).

To provide synchronism of the input signal and the local oscillator voltage, the useful signal is extracted with the help of the narrowband filter in the synchronism circuit, and it passes to the local oscillator (the heterodyne) for locking of the oscillation frequency by the external weak signal. Effective operation of such a receiver, especially in the frequency band, can be achieved only at coincidence of the local oscillator frequency with the frequency of the carrier oscillation of the receiving signal with accuracy to the phase. The useful signal spectrum with the central frequency f_s acting on the mixer input, after multiplication with the local oscillator signal, is transferred onto zero frequency with further filtering in MFF. The low-frequency signal extracted by the filter and amplified in the modulation frequency amplifier (MFA) passes to reproduction devices of the received signal. Synchronism and in-phase of two voltages on the mixer inputs (Mix) are rather difficult to provide at reception of analogous signals, especially at operation in the frequency range. Therefore, homodynes has got the wide distribution in digital communication systems, where the high-stable master oscillators are used, and the frequency and phase control is performed after digital processing in the post-detector section (from informational section).

3.1.2 Structural Diagrams of Receivers Without Filter Tuning on the Receiving Frequency

Evidently, the receiver of *direct conversion* operating in the frequency range (Fig. 3.1) cannot be realized on the non-tuned circuits since, together with the receiver tuning on the frequency of receiving signal, they provide selectivity on the adjacent channel. With the help of filters (single- and double-circuit) in each amplifying stage, the resulting AFC is formed defined as the product of gains of all n -stages of the linear section $\prod_{i=1,n} [K_i(f)]$ and providing the required suppression of the adjacent channel.

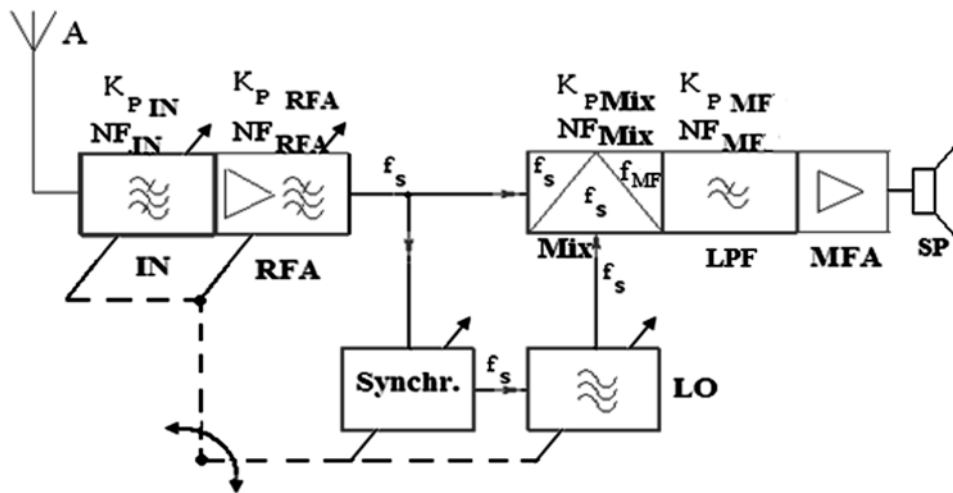


Fig. 3.10 The homodyne receiver with frequency conversion to the zero frequency

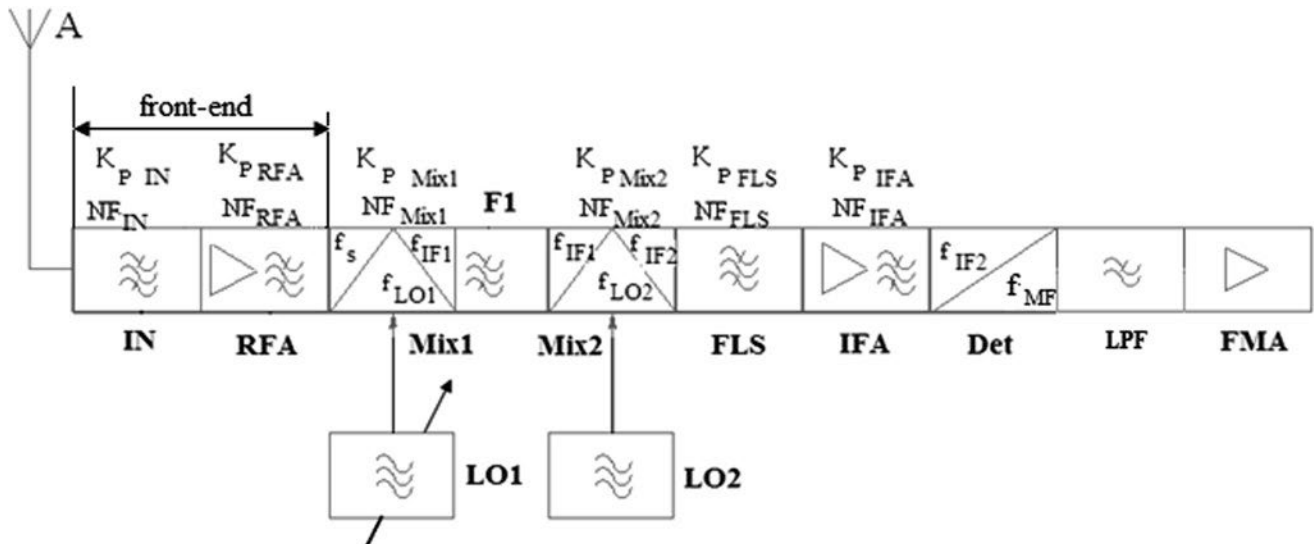


Fig. 3.11 The architecture of the super-heterodyne receiver with double conversion

3.1.2.1 Receivers of the Super-Heterodyne Type

The architecture of *super-heterodyne* receivers with non-tuned preselector (Fig. 3.11) is similar to the receiver structure realized with application of tuned filtering systems in the input circuit and RFA (Fig. 3.8).

The main difference of the receiver from having the tuning consists in the fact that the pass-band of front-end is chosen the essentially higher of the pass-band occupied by the input signal (it occupies the pass-band of the receiving *frequency range*). Application of wideband filters in the preselector (front-end) implemented as multi-circuits on the surface-acoustic waves, etc. permits to exclude the necessity of its tuning at receiver operation in the frequency range, while the high AFC slope realizable in this case provides the necessary suppression of the image channel on the first intermediate frequency. In this case, the problem conjugation of the preselector circuit and the local oscillator is eliminated, which required an increase in the bandwidth of front-end and, accordingly, led to decrease the sensitivity of the receiver.

Tuning on the useful signal is performed by variation of the local oscillator frequency. For the input signal with the frequency f_s , which multiplies in Mix1 by the local oscillator voltage and has the same frequency f_{LO1} that the combinational component, for instance, $f_{IF1} = f_{LO1} - f_s$, happens to be equal to the central frequency of the F1 filter, and this signal falling in the pass-band of F1 is extracted and then is exposed by the second conversion, amplification and so on. The main destination of F1 filter consists in suppression of the second image frequency signal ($f_{im, ch2} = f_{LO2} \pm f_{IF2}$) up to required value and in some attenuation of adjacent channels, and since $f_{IF1} = \text{const}$, then its implementation does not cause difficulties. All other conversion products happen to be outside the F1 pass-band, where they are suppressed to the required level.

Suppression of adjacent channels is achieved with the help of FLS, which is usually made as multi-chain LC, piezoceramic, or electromechanical filters. If the very narrow pass-band is required (hundreds of Hertz), then we can use piezocrystal (quartz) filters.

Difference in the receiver construction with tuning of the front-end circuits and without tuning consists in the method of receiver tuning only in the operating frequency range and in concrete realization of filtering devices. Achievement of the required value of $S_{eim, ch1}$ usually assumes the application of the double-frequency conversion, since, especially with the operation frequency growth, it is difficult to simultaneously satisfy the contradictory conditions of required suppression of the image channel and irregularity σ of preselector AFC in the operating frequency range.

3.1.2.2 Receiver of Direct Conversion

Receivers of analog signals of *direct conversion* with non-tuned preselector (Fig. 3.12) did not find the practical application due to the complexity to achieve the optimal relations for the frequency and the phase of the input signal and the reference oscillator.

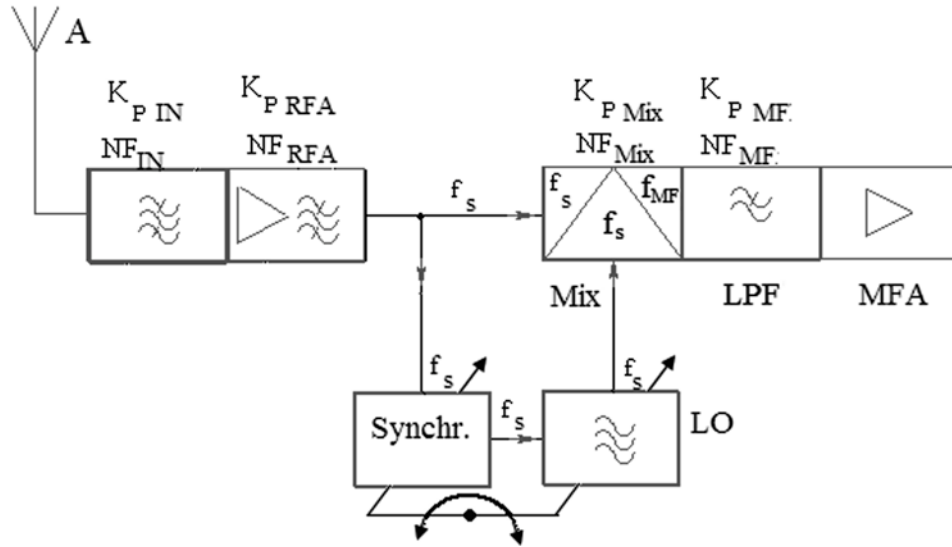


Fig. 3.12 The receiver of direct conversion with non-tuned preselector

However, such a diagram is actively implemented by developers of mobile terminals of the digital cellular communication systems and the radio access systems, which are used for operation in several frequency ranges and with different standards simultaneously. In this case, the frequency control of local oscillators, which is made as the voltage-controlled oscillators (VCO), is provided from the digital section, and application of tuned narrowband filter in the synchronization circuit is not necessary.

3.2 Receivers of the Mobile Communication Systems

3.2.1 The Architecture of Transceivers for Systems of Digital Radio Communication

The transceivers contained in the structure of municipal networks, personal networks, the cellular communication networks, and all new developments of the equipment of global communication networks use the digital methods of speech signals, data, and multi-media processing to form the digital modulated sequence. In order to support the constancy of the envelope of the transmitter output power (with the peak-factor close to zero), we can use the noise-resistance methods of shift-keying carrier modulation (FSK, GMSK, PSK, QPSK, nQAM). Parameters of the modulating binary sequence in the transmitter (duration and the pulse shape, with return to zero or without return) are chosen in such a manner that filter responses at demodulator output would provide the pulse orthogonality for received sequence.

In systems of mobile radio communication, the subscriber terminal should provide the duplex communication mode at keeping small physical dimensions. Therefore, terminals of such systems are implemented as transceivers (receivers-transmitters), which have the system of separation of the directions of reception and transmission (Fig. 3.13).

At operation in several frequency ranges, the *device for section separation* should not only separate the reception and transmission sections, but also connect the appropriate antenna (on the base station) and the pass-band filter of the high-pass and low-pass filters with parameters, which are extracted the operation bandwidth of reception and transmission. The receiving part of the radio-frequency transceiver block, which realizes the above-described construction principles, should provide the subscriber signal reception without adjusting of its operating frequency, to detect signals with different types of modulation. Besides, most often, it must provide the terminal operation in various frequency ranges and process signals of different networks and standards without any manual adjustments.

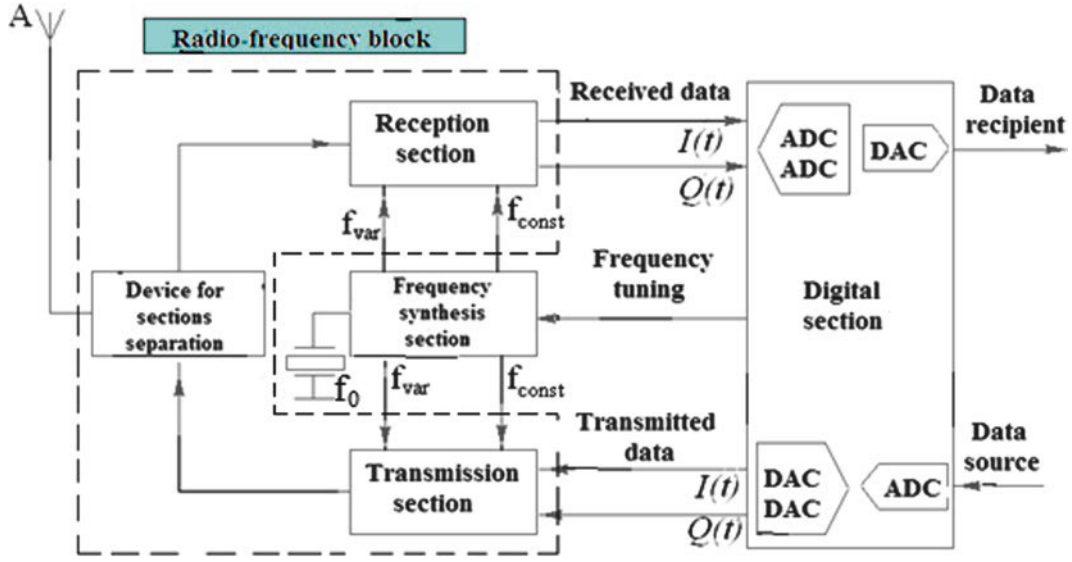


Fig. 3.13 The transceiver architecture with separation of reception and transmission directions

3.2.2 Features of Super-Heterodyne Receivers

Comparing, for instance, the structural diagram of the *super-heterodyne* receiver (Fig. 3.11) with the receiver using digital methods of signal processing (Fig. 3.14), we can note that the structure of the radio front-end does not change. Now only the detector is used as the quadratic demodulator forming at output both in-phase and quadrature digital sequences $I(t)$ and $Q(t)$ of the received signal.

An antenna creates the voltage in the receiver input, which contain the useful signal together with many of others spurious influences. The input circuit together with RFA provides the extraction of the operating frequency band containing the useful signal and the suppression of the image channel, which is defined by the choice of the first intermediate frequency, for instance, for lower local oscillator frequency tuning $f_{im, ch1} = f_{LO1} - f_s$, and also the suppression of the direct passage channel f_{IF1} .

The value of the first intermediate frequency is chosen as constant. Because of multiplication on the Mixer1 nonlinear element of the signal frequency f_s and the local oscillator f_{LO1} , the combinational components with frequency $f = mf_s \pm nf_{LO1}$ are formed in its output current, where $m, n \in \mathbb{Z}$. The first intermediate frequency is usually $f_{col} = f_{IF1} = f_s - f_{LO1}$, $f_{IF1} = f_s - f_{LO1}$ ($m = n = 1$, the simple conversion) chosen from the condition that in the spectrum of the Mixer 1 output current, the components, which are the collateral conversion channels for Mixer 2, are absent. The reason of such an approach to the choice of f_{IF1} is the fact that the condition of infradyne realization (3.7) is impossible to perform for any receiver of the super-heterodyne type with double-frequency conversion for systems of mobile communication. Therefore, f_{IF1} is chosen with the help of nomograms in the interval of values, which do not contain the collateral conversion channels, and the list of piezo-ceramic filters and SAW filters produced by various manufacturers.

The non-tuned preselector (front-end) often has the very wide pass-band compared with the pass-band allocated by the single channel, for instance, for the GSM900 standard: the bandwidth of operating frequency range is 25 MHz, and the allocated channel is only 200 kHz. Including the low-noise amplifier RFA in the preselector under impact of powerful signal (interference) sources in the operating range or out-of-band emissions of closely located transmitters can create nonlinear distortions due to restriction of the dynamic range of the AE stage. Therefore, a choice of f_{IF1} should be performed with the account of potential danger of falling in the receiver input of lumped interferences, which are created by other base stations and lead to appearance in the FC1 input of components in the operating frequency ranges or closely to f_{IF1} . It is necessary to take into consideration that in the output current of the first frequency converter (FC1), the following combinational components may occur, which at inconvenient choice of f_{IF1} will coincide with frequencies of collateral channels of conversion for FC2.

Digital section fulfills the post-detector signal processing, and independently on the method of modulation, it should eliminate the difference between the frequency of VCO1 connected to the Mix1 and the frequency of the input signal. This

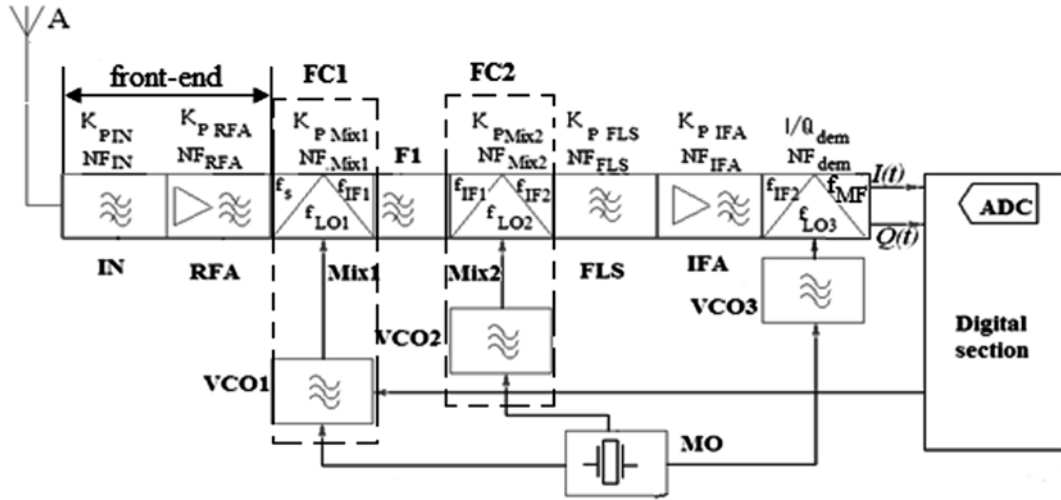


Fig. 3.14 The super-heterodyne architecture of the double-conversion receiver for digital signal processing

allows obtaining of the frequency value of the converted signal exactly equaled to the mean filter frequency on the mixer output. On the contrary to analog systems, where high stability of the intermediate frequency is performed by the automatic frequency control (AFC) system (on intermediate frequency of the local oscillator frequency), in digital systems, it can be achieved by application of frequency synthesizers stabilized by the quartz and spanned with the phase-locked loop. The high frequency stability of the VCO1 is provided by the quartz stabilized master oscillator (MO) and its exact value by the value of varicap capacitance included in the VCO1 loop created in the digital section.

The operating frequency band acting at the input of FC1 transferred into the region of lower frequencies is extracted by the F1 filter with the central frequency f_{IF1} equaled to the mean frequency of the converted band. The multiplication of the converted frequency band with the center frequency f_{IF1} by the voltage of second local oscillator with frequency f_{LO2} reacts in the spectrum of the output current FC₂ a large number of combinational components containing the frequency band of the useful signal selected by the bandpass filter (FLS) with the central frequency f_{IF2} , for example, $f_{IF2} = f_{LO2} - f_{IF1}$ (at double-frequency conversion, it is not necessary for realization $f_{LO2} > f_{IF1}$). Reduction of linear distortion created by FLS due to difference of the mean filter frequency and f_{IF2} can be achieved by supporting of the exact value of frequency f_{LO2} provided by VCO2. The control of the current values of VCOs frequencies is performed from the digital section. The pass-band filter FLS (the filter of lumped selectivity) on the output of Mix2 provides the suppression of the adjacent channel according to technical requirements on the produced equipment.

The signal from the detector output frequently realized as I/Q-demodulator fulfilling the separate processing of in-phase and quadrature signals with regard to the phase of the reference oscillator passes to the summator. The output signal under influence of phase shift inaccuracy of channels and differences of amplitudes will differ from the signal at the transmitter modulator input. The demodulator circuit is factually the synchronous detector, for which effective operation the coincidence with the accuracy to the voltage phases with the frequency f_{IF2} and VCO3 voltage is required. The double-frequency conversion is not necessary condition of the radio front-end realization of the mobile terminal, since the growth of conversion number of input signal decreases the receiver dynamic range and increases the level of own noises. Everything is defined by requirements to receivers, the main of which are the selectivity of frequencies of the collateral channels, the sensitivity, the bandwidth of the radio front-end, and the value of admissible and real noise factors.

3.2.3 The Receiver of Direct Conversion

3.2.3.1 The Receiver of Digital Communication Systems

Widening of mobile communication standards, the transfer to transceiver construction operating in different frequency ranges, and receiving signals with the different band and various technologies of the multi-station access revived the interest to diagrams (Fig. 3.15) of receiver of direct conversion.

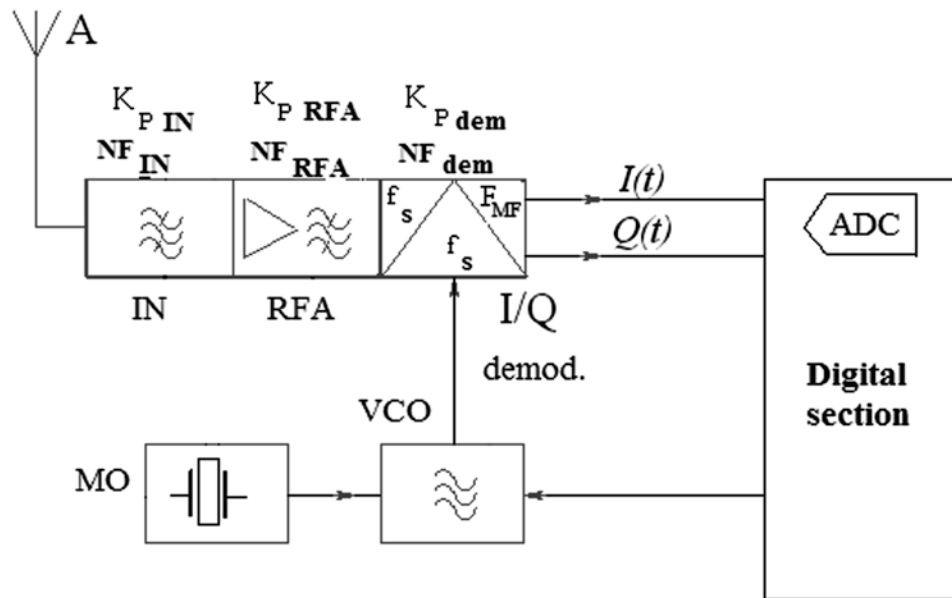


Fig. 3.15 The architecture of the receiver of direct conversion with digital signal processing

At high frequency stability of the master oscillator (MO), the synchronism of the input signal and the signal produced by VCO is provided simply enough by variation of the varicap capacitance value connected to the VCO circuit. In the digital section, the constant voltage is formed, which value is defined by the frequency difference of the input signal and the signal produced by the frequency synthesizer spanned by the PLL system. The main limitation in application of such an architecture is the restriction in the IC nomenclature manufactured by manifold producers, UHF converters (I/Q demodulators), which are capable to provide signal processing at low input level. The suppression of the adjacent channel is performed on the low frequency with the help of high-effective analog and digital filters of high order. The conversion to the zero frequency will take off the problem of the image channel suppression, which is typical for super-heterodyne receivers due to its absence ($f_{Lo} = f_s$). We can provide the signal reception in the frequency ranges belonging simultaneously to several standards by the choice of front-end parameters, mainly, the pass-band and the value of its central frequency. This permits to construct the mobile terminals operating in the global mobile network.

An interest to implementation of the subscriber receivers using the direct conversion of the receiving signals is caused by several reasons: the necessity of receiver designing with lower number of blocks in the radio front-end and operating in several frequency ranges as well as in networks of different standards. The technological progress allows the IC production (frequency converters), which have the low-noise factor at action of input signals with low level. Such an architecture is dominating in radio receivers of WLAN (Wi-Fi) and WPAN (Bluetooth, DECT) network, when the conversion of receiving signal is provided into low intermediate frequency (low IF conversion) or into zero frequency (DC conversion).

The most distributed filtering method of signal processing in receivers of analog and digital signals, which is used for the adjacent channel suppression, requires the single-, double-, and even triple-frequency conversion of receiving signals. Application of electro-acoustic, piezo-electric, and quartz or having the microstrip construction pass-band filters tuned on the appropriate frequency allows suppression of image, semi-image, adjacent and other channels, but it causes trouble in application of the unified integral technology at manufacture of the active block of the radio front-end and passive two-ports. Utilization of the quadrature signal processing and the microelectronic low-pass filter (LPF) in the receiver of direct conversion (so-called, synchrodyne) allows effective solution of mentioned problems. At that, the increased requirements to characteristics of oscillators and converters are made in the receiver of direct conversion.

The main advantages of the direct conversion receiver are:

- The simplified front-end structure because the necessity of the image channel suppression is absent.
- The main filtering (suppression of the adjacent channel) is transferred into the low frequency region, which allows extraction of the modulating frequency component of receiving signal with the help of LPF. Such a filter fulfilled on the integral technology compatible to the standard MOS-technology can be implemented as the trans-conductive amplifier [96].

3.2.3.2 Quadrature Modulator

In the subscriber radio stations realized as the transceiver, the frequency synthesizer usually performs the function of the master oscillator for the receiver and transmitter simultaneously. For formation of the *single-side signals with suppressed carrier in the transmitter*, the quadrature modulator circuit is used (Fig. 3.16).

For input (modulating) signal $u_{in}(t) = U(t) \cos \Omega t$, the in-phase component on the Mix1 input $u_I(t) = U(t) \cos \Omega t$ acts, while the quadrature component $u_Q(t) = U(t) \cos (\Omega - \pi/2) = U(t) \sin \Omega t$ acts at the Mix2 input. When the digital stream is the modulation signal, the amplitude of the input signal $U(t)$ takes the only two values: zero and unit. For the local oscillator signal

$$u_0(t) = U_0 \cos \omega_0 t, \quad (3.11)$$

the voltage on the Mix1 output

$$u_{Mix1}(t) = 0.5 \cdot K_{mix} U(t) U_0 [\cos(\omega_0 - \Omega)t + \cos(\omega_0 + \Omega)t],$$

and on Mix2 output

$$u_{Mix2}(t) = 0.5 \cdot K_{mix} U(t) U_0 [\cos(\omega_0 - \Omega)t - \cos(\omega_0 + \Omega)t],$$

where K_{mix} is the transfer coefficient of each mixer. On the output of the balance modulator, signals corresponding to the upper $u_+(t)$ or lower $u_-(t)$ side band are formed:

$$u_+(t) = K_{mix} U(t) U_0 \cos(\omega_0 + \Omega)t, \quad (3.12)$$

$$u_-(t) = K_{mix} U(t) U_0 \cos(\omega_0 - \Omega)t. \quad (3.13)$$

The I/Q demodulators, which contained two channels, are used for recovering of the modulating signal on the receiving side.

3.2.3.3 Synchrodyne with a Quadrature Demodulator

The simplified structural diagram of the homodyne receiver is presented in Fig. 3.17.

EMF is induced in the antenna and creates the voltage at the receiver input, which is amplified in the low-noise amplifier (LNA). Besides the low-noise level, this amplifier should have the high gain to reduce the own receiver noise level and the pass-band filter connected with the output circuit, which restricts the bandwidth of receiving frequencies. The input circuit fulfilling the function of receiver input matching with the antenna has also filtering properties extracting some band of the operating frequencies containing the useful signal. The voltage from the LNA output is separated into two channels and passes to inputs of Mix1 and Mix2 mixers. At the same time, along with the input signal, each mixer is supplied a voltage

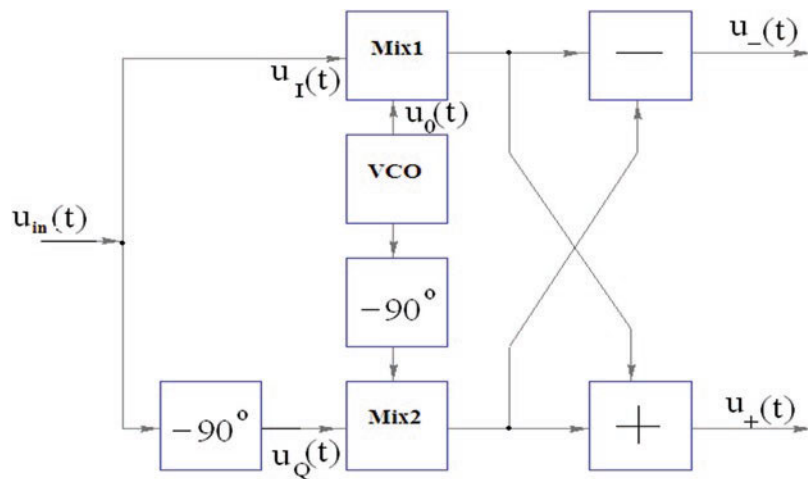


Fig. 3.16 Formation of the single-side signals with suppressed carrier in the transmitter

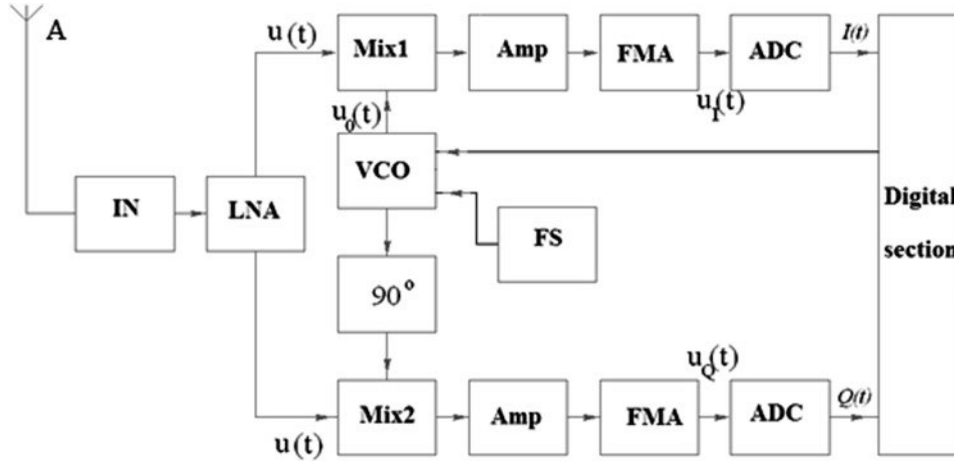


Fig. 3.17 The simplified structural diagram of the homodyne receiver

from the VCO, the frequency of which is regulated by a control voltage from the digital section. The initial VCO frequency value is specified by frequency synthesizer (FS), which forms the frequency grid, which base frequency is defined by the frequency of the master oscillator (MO) and is stabilized by the quartz.

At action on the mixers' inputs of the signal, corresponding to received upper side band (3.14)

$$u_{in}(t) = aU(t)\cos(\omega_0 + \Omega)t, \quad (3.14)$$

where a is the proportional coefficient of the local oscillator voltage defining by (3.12), the voltage on the Mix1 output will be

$$u_{Mix1}(t) = 0.5aK_{mix}U_m(t)U_0[\cos\Omega t + \cos(2\omega_0 + \Omega)t], \quad (3.15)$$

on the Mix2:

$$u_{Mix2}(t) = -0.5aK_{mix}U(t)U_0[\sin\Omega t - \sin(2\omega_0 + \Omega)t]. \quad (3.16)$$

After signal amplification in the amplifier with controlled gain and suppression of the total component with the frequency $2\omega_0 + \Omega$ in LPF, the quadrature components are

$$u_I(t) = 0.5aK_{mix}U(t)U_0\cos\Omega t, \quad (3.17)$$

$$u_Q(t) = -0.5aK_{mix}U(t)U_0\sin\Omega t. \quad (3.18)$$

for the initial signal passed to the ADC input.

The in-phase component $I(t)$ and quadrature component $Q(t)$ of radio signal with a modulation frequency, converted in each arm into digital sequences $I(t)$ and $Q(t)$, respectively, enter the digital path, where further signal processing (detection, interleaving, decoding data and speech, etc.) is performed.

3.2.3.4 Features of Homodyne Receivers

Achievement of such receiver characteristics as the small number of blocks in the radio front-end, the absence of the image channel, a possibility to adjacent channel suppression by simpler measures, and the application of the integral technology for their realization simultaneously lead to the appearance of other difficulties for obtaining of specified technical indices. Reduction of application effectiveness of such receivers is caused by:

- The leakage of the local oscillator signal to the LNA input.
- The absence of complete parameter identity of the in-phase and quadrature converter channels.
- The influence of redundant (flicker) noises.
- The influence of the phase noises of the frequency synthesizer.
- The effect of nonlinear distortion.

The leakage of the local oscillator signal to the LNA input, in contrast to receivers of the super-heterodyne type, may show the essential effect on distortion value caused by overloading of low-frequency amplifiers, even at deep gain adjustment. At multiplication of the input signal $u(t)$ and the local oscillator voltage $u_0(t)$, on the Mix output (Fig. 3.18), the low-frequency (Ω) voltage and the component of $2\omega_0 t + \Omega$ are formed, and the last is then filtered in the LPF.

The leakage of the local oscillator voltage to the mixer input leads to its multiplication with the local oscillator voltage, which is applied on the local oscillator input and, accordingly, to appearance of the component with the $2\omega_0$ frequency and the DC component. Such the “spurious” DC falls into the operating channel band and at large values can significantly influence on the receiver noise immunity.

To obtain the low-noise factor of the radio front-end reduced to the input, it is necessary that the LNA/Mixer gain would be about 30 dB. If the suppression of the local oscillator signal in this structure is -60 dB, then for amplitude of the local oscillator voltage of 1 V, the *value of the bias voltage* on the mixer output (DC component) will be 31.6 mV. For the gain from the antenna to ADC (Fig. 3.17) of about 100 dB, this will cause the overloading of the next stages even for AGC action. Connection in the input of I/Q modulator of the Gaussian filter (application of GMSK modulation), which AFC has the maximum on the zero frequency, increases the bias voltage value. If the subscriber station in the mode of reception/transmission is moving, then if antenna emits the local oscillator power, the signal received by the antenna due to the Doppler effect will have somewhat another frequency, which will cause the appearance of the slowly changing component on the mixer output.

To eliminate the bias voltage, the following methods are applied:

- Addition of blocking capacitors on the mixer output.
- Addition of high-pass filters.
- Compensation of the bias voltage.

Application of the blocking capacitors and HPF, which cutoff frequency is 0.1% of the data transmission rate, leads to necessity of utilization of large area on the chip and, as the consequence, the large time constant of the filter. This may cause

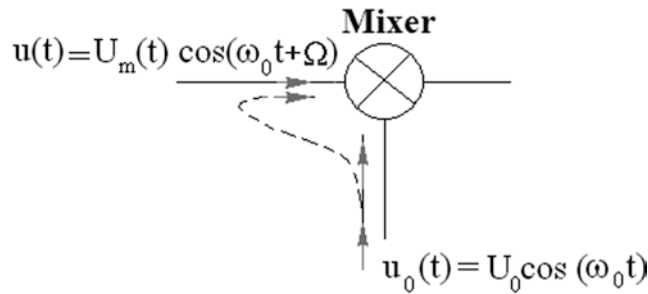


Fig. 3.18 Formation of signals during the signal multiplication in the mixer

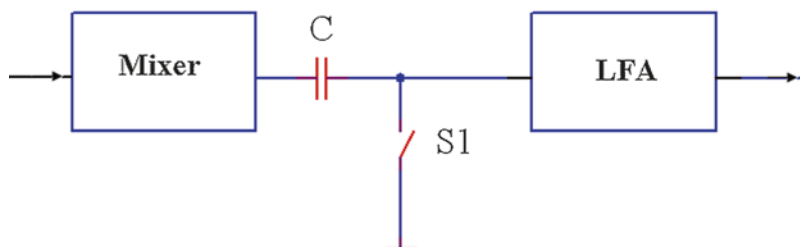


Fig. 3.19 The circuit for suppression of the spurious bias

additional losses of several initial bits in the pulse train after comparably long break at reception, which is caused by the absence of initial bias and the long transient. Some attenuation of the initial bias effect can be achieved by application of binary FM signals (BFSK), having the amplitude spectrum with the maximum on the frequency 4.5 kHz. The DC component of the BFSK amplitude spectrum has the value by 25...30 dB less than maximal value.

For suppression of the spurious bias, we can use the circuit shown in Fig. 3.19.

The operation principle is based on the fact that in systems with time channel division (TDMA) in the time intervals, when traffic is interrupted to arrange the new channel, the capacitor C at the mixer output is grounded, eliminating the bias voltage. In the active step, the SI switch opens eliminating the appearance of DC component on the LFA input. Owing to the fact that the C capacitor has rather large capacitance, it occupies a quite significant area on the chip at integral implementation and, accordingly, has the relatively low own Q-factor. Losses in the capacitor lead to arising of thermal noises, which level is inversely proportional of the capacitance value, which indicates on necessity of its increase. Nevertheless, the mentioned method allows elimination of the slowly changing bias component caused by the Doppler effect.

The identity absence of the demodulator channel parameters (in-phase and quadrature) in the receiver can lead to appearance of distortions, which cause the reduction of SNR and, as a consequence, the sensitivity decrease.

Let us represent the input pulse signal in the form of a sum of quadratic components:

$$u(t) = b \cos \omega_0 t + c \sin \omega_0 t,$$

where b and c are coefficients, which take values ± 1 . Since the phase shift of the receiving signal can in general lead to its distortion and the noise level increase, it is expedient to form the phase shift in the local oscillator section (Fig. 3.17), where the signal has the constant parameters. The local oscillator voltage (3.11), which passes on mixers of in-phase and quadrature channels, takes a form:

$$u_{0I}(t) = U_{0I} \cos \omega_0 t,$$

$$u_{0Q}(t) = U_{0Q} (1 + \varepsilon) \cos(\omega_0 t + \theta),$$

where ε and θ define the amplitude and phase deviations caused by the channel mismatch. The input signals of each channel ADC (Fig. 3.17) will be $U_I(t) = 0.5 K_{mix} b$,

$$\begin{aligned} U_Q(t) &= 0.5 K_{mix} [(1 + \varepsilon) c \cdot \cos \theta - (1 + \varepsilon) b \sin \theta] \\ &= 0.5 K_{mix} [c \cdot \cos \theta - b \sin \theta], \\ U_I(t) &= 0.5 K_{mix} b, \end{aligned}$$

Taking into consideration that at I/Q demodulators inputs are usually used the logarithmic amplifiers, the influence of the amplitude error does not practically manifest. Phase distortions (phase shifts) lead to variation of digital sequence shape (Fig. 3.20).

However, the SNR decrease occurring due to the amplitude-phase conversion in receivers of direct conversion turns out to be less noticeable for the same phase shifts compared to receivers of the super-heterodyne type. In any case, errors, which are caused by the different phase shift in channels, and the mismatch of signal amplitudes $I(t)$ and $Q(t)$ disturb the canonic type of the receiving signal constellation at down-conversion, thus increasing the coefficient of bit errors. For GSM systems, the mismatch of channel parameters is considered as admissible up to 2° in phase and 0.3 dB in amplitude.

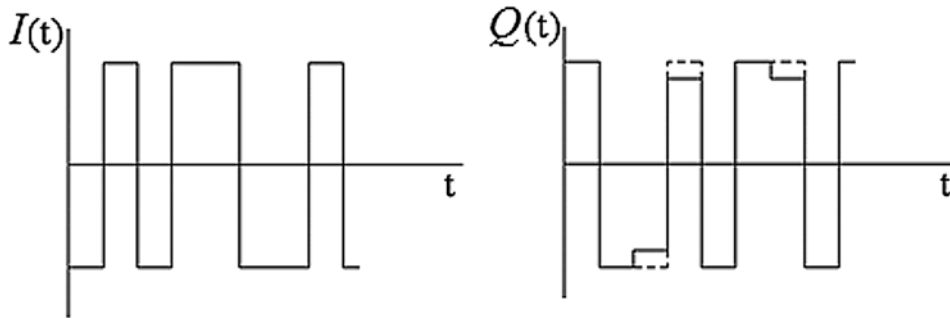


Fig. 3.20 Phase distortions lead to the shape variation of the digital sequence

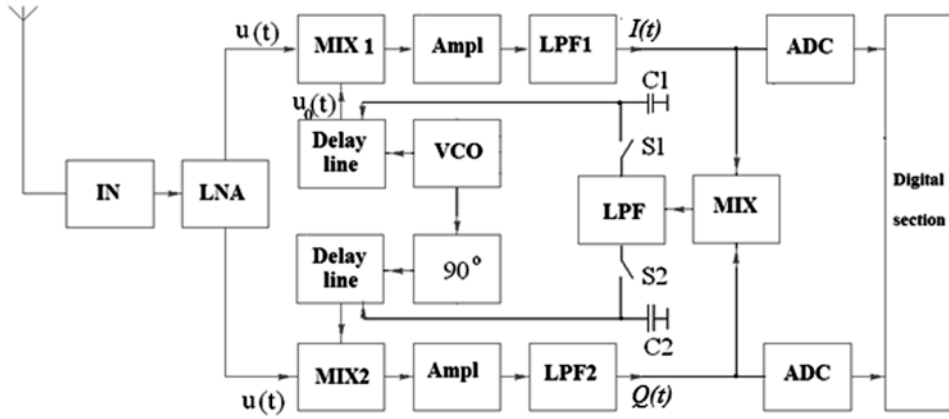


Fig. 3.21 The architecture of the receiver with digital signal processing and the calibration circuit

To eliminate the mismatch, we can use the calibration circuits, for instance [63, 64], on the base of PLL (Fig. 3.21).

The additional (compared to Fig. 3.17) mixer (Mix), the smoothing low-pass filter, adjustable delay lines (phase correctors), two capacitors (C1, C2), and two switchers (S1, S2) are used in this architecture. In the calibration mode (switchers are closed), the testing signal close in frequency to the local oscillator signal is applied on mixers Mix1 and Mix2 inputs. From outputs of LPF1 and LPF2 filters, the quadrature components of the testing signals are applied to the mixer creating at output the voltage proportional to the phase difference between $I(t)$ and $Q(t)$. The smoothing LPF on the mixer output creates the voltage varying the value of phase shifts in the delay lines. In the operating mode, switchers S1 and S2 are open, and the control voltage is supported by slowly discharged capacitors C1 and C2. We should note that permissible mismatch of channel parameters for GSM systems is up to 2° and 3 dB.

Nonlinear distortions in the receiver of direct conversion. In contrast to receivers of the super-heterodyne type, where the main type of nonlinear distortions are intermodulation distortions [64] of the third order, the distortions of the second order are strongly expressed in receivers of direct conversion. At action on the preselector input, which has the nonlinearity (AE of LNA), of the signal $u(t) = U \cos \omega_1 t + U \cos \omega_2 t$, the low-frequency oscillation at the LNA output arises with the difference frequency:

$$y(t) = a_2 U^2 \cos(\omega_1 - \omega_2)t. \quad (3.20)$$

In the super-heterodyne receiver, the component of the converted useful signal on the mixer output is transferred into the intermediate (difference) frequency f_{IF} and – another combinational component – into the summing frequency, which are filtered out by the IF filter in the IF section. In a direct conversion receiver, the lack of identity of mixer parameters (e.g., in branch of a differential stage) leads to leakage of a difference (close to zero) frequency signal to its output, which falls into the bandwidth of the LPF channel. Application of the balance circuit of the RFA can weaken the leakage effect at that increasing the consuming power and the necessity of utilization of the symmetric input device. However, the application of an antenna with non-symmetric output in the subscriber radio station leads to the necessity of the symmetric transformer in the input of the balanced LNA, which decreases the receiver sensitivity and degrades of the mass-dimension parameters due to utilization of the add-on elements. The influence weakening of intermodulation products in the LNA output can be achieved in the simplest case by application of the blocking capacitor between LNA and the mixer.

Flicker noises in the receiver of direct conversion may show the significant influence on the receiver sensitivity, since the conversion happens in low frequency, where its influence is the most noticeable.

The implemetation of passive and active IC components based on use of a unified MOS technology simplifies the evaluation of noise perfomance of such stages. Application of MOS technology with admissible dimensions less than $1 \mu\text{m}$ at the gain width up to hundreds of μm allows approximation by the expression [35]:

$$S_F(f) = \frac{K'}{WLC_{ox}} \frac{1}{f} \quad (3.21)$$

of the spectral density of the flicker-noise average power, where K' is the specific conductance (the slope) of the transistor; W, L are geometrical dimensions of the transistor channel, the width and length of the channel; and C_{ox} is the specific capacitance of the dielectric layer (oxide).

The flicker-noise power defining for the frequency band from 10 Hz to 200 kHz, which corresponds to the radio-frequency band of the single channel of the GSM standard:

$$P_F = \int_{f_1}^{f_2} \frac{K}{WLC_{ox}} \frac{df}{f} = \frac{K}{WLC_{ox}} \ln 2 \cdot 10^4.$$

The power spectral density of thermal noises, which becomes equaled to the value of the spectral density of flicker noises at about 1 MHz:

$$S_f(f = 10^6) \approx S_T = 4kt \frac{2}{3g_m},$$

where g_m is the conductance of the transistor channel, Sim; $k = 1.38 \cdot 10^{-23}$ J/K·Hz is the Boltzmann constant; and T is the absolute transistor temperature, K°.

From this, the power of the flicker-noise component

$$P_F = 4kT \frac{2}{3g_m} (1 \cdot 10^6) \ln 2 \cdot 10^4, \quad (3.22)$$

since in the same band, the thermal noise power is

$$P_T = 4kT \frac{2}{3g_m} 2 \cdot 10^5.$$

Thus, the thermal noise power in the band of $\Delta f = 200$ kHz (the band allocated on the single channel of data transmission) is approximately by 50 times less than the flicker-noise power in the same frequency band. To decrease the flicker noise, we can use MOS-transistors with the larger channel width, which, nevertheless, simultaneously degrades the frequency properties of transistors. The most effective method of flicker-noise reduction is the correlation signal processing on the converter input, which acts in the mode of LO frequency doubling.

Phase noises and discrete collateral components of the frequency synthesizer show the essential influence not only on the value of the distorted bits coefficient, but on the level of blocking in the reception section. Reference signals in the frequency synthesis sections of radio-frequency devices should have the monochromatic spectrum. However, in real systems, the phase spectrum is always present in the output signal spectrum of the frequency synthesizer, which leads to error appearance at reception together with other factors decreasing the permissible reception quality defining by standards. There are two main mechanisms of phase noises influence of the reference signals. The first is the fact that the presence of phase noise causes broadening of the spectral line of the oscillator, which leads to multiplication of the useful and adjacent channels, and as a consequence, an increase in the noise level of the mixer. Conversion of the noise component of the local oscillator voltage with the useful signal in the mixer leads to the same result. It is possible to increase the “purity” of the spectrum formed by VCO by using its tuning circuit with high selective properties. The second mechanism is called the inverse noise conversion. In this case, the noise components with the intermediate frequency are formed owing to conversion of the phase noise spectrum components. The value of contribution in the total noise component of the inverse conversion products happens to be noticeably less than the result of direct conversion of frequency synthesizer noises.

The influence of mixer load parameters. The properties of the low-frequency part of the section, the LPF and ADC, have the essential influence of the receiver noise immunity. The optimal choice of the LPF order and the ADC digit capacity allows provision of the best SNR at minimal power consumption. At implementation of LPF, the active filters 4÷5th order with the Butterworth characteristics have found the widest application, which have the minimal irregularities in the pass-band [91]. The LPF of 5th order in the receiver of direct conversion of GSM standard decreases SNR by 0.7 dB. Multifunctional multi-range receivers intended for processing of digital streams of GSM (EDGE) and UMTS standards should apply the filter of the 6th order, which provides the same value of SNR reduction at complication of the LPF construction.

The transfer of the adjacent channel filtering in the band of the main spectrum leads to increase of the discretization frequency and the ADC digit capacity. The digit capacity is a number of bits of amplitude level representation of the analog signal; the discretization frequency is defined by the repetition frequency of the clock pulse oscillator of the reference oscillator connected to the input of the ADC synchronization block.

3.2.4 Receiver with the Digital Intermediate Frequency

Development of technique and technology of digital ICs led to the fact that final mixing and filtering provided in the IF stages can be performed already in the digital domain. In receivers with the digital IF (the digital IF receiver), the digitization of the IF signal is directly performed. As the local oscillator voltage source, the direct digital frequency synthesizer (DDS) is used

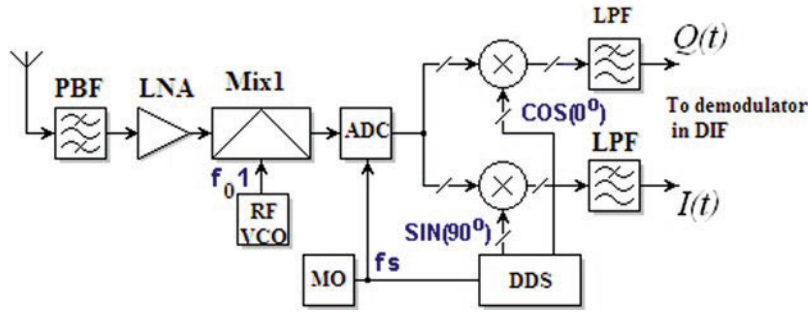


Fig. 3.22 Generalized structure of the receiver with digital IF

which is sometimes called the oscillator with digital (programmed) control numerically controlled oscillator (NCO). This device is implemented completely with usage of digital technique, and it is produced by a series of manufacturers in the form of specialized IC. The oscillator forms the digital samples of two sinusoids with the exact phase shift by 90° (Fig. 3.22).

There is an important fact that the intensity of sine sample formation is always defined by the reference frequency f_s independently on the nominal of generating frequency f_0 . The nominal output frequency is specified by variation of the phase advance per sample. The small phase advance per sample corresponds to low frequencies; the large advance to high frequencies. The value of phase advance per sample is proportional to the output frequency and is programmed in the range from 0 to $f_s/2$.

The digital mixer, which factually consists of two digital multipliers, is the important component of such a receiver. Digital samples of the input signal from ADC are mathematically multiplied with digital samples of sine and cosine passing from the output of the digital local oscillator. In contrast to analog mixers, which create also a variety of unwanted components on the output, the digital mixers are practically ideal devices and produces only two output signals of summed and difference frequencies.

The reference ADC signal f_s is applied on the DIF local oscillator. Digital sine samples from the LO output are defined by the reference frequency f_s and are generated with the rate equaled to sample frequency of ADC, which is synchronized by the single reference signal f_s . Application of the digital IF, besides all others, allows avoid of the I and Q channel imbalance, which leads to well suppression of the image channel. Quadratic components formed with the help of the DDS, and they pass to the digital mixer including two digital multipliers, where mathematical multiplication operation is performed. In the structure under consideration, the ADC plays the role of a digital sequence generator, spectrum of which can be obtained by installing a LPF at its output. The main obstacle to using such a structure in modern data transmission networks is the requirement that the sampling frequency f_s must be twice the frequency of the signal at the ADC input (Kotelnicov's theorem). The use of a high-speed AFDC will simplify the structure of the receiver, eliminating (or seriously reducing) the analog path, while preserving the function of digital mixer as a shaper of quadrature (I and Q) digital sequences. However, this architecture requires the high-speed ADC, and this causes the increase of the consuming current of all reception section.

3.2.5 Receivers with Conversion to the Low Intermediate Frequency

Receivers with conversion to low (but not zero) frequency have better characteristics compared to above-considered receivers and simultaneously allow realization of front-end fully on the integral technology [91]. Such receivers conveniently combine advantages of the super-heterodyne receivers and advantages of direct conversion receivers. At that, the image channel suppression, as a consequence of the mixer addition in the RF front-end, happens to be not in preselector with utilization of the filtering method, but mainly in the intermediate frequency section with application of the quadrature signal processing. In the cellular communication systems of the GSM standard, the intermediate frequency f_{IF} are usually chosen in the interval 250...400 kHz, which is approximately equaled to the twice value of the modulation frequency of the transmitted signal. Such a choice of f_{IF} allows elimination of shortcomings of direct conversion receivers related to the leakage of the LO signal (appearance of bias voltage), the sensitivity decrease of the homodyne receiver due to flicker-noises, etc. The structural diagram of the receiver with input signal conversion on the low intermediate frequency with application of *phase methods of the image channel suppression* (the Wayver architecture) is shown in Fig. 3.23.

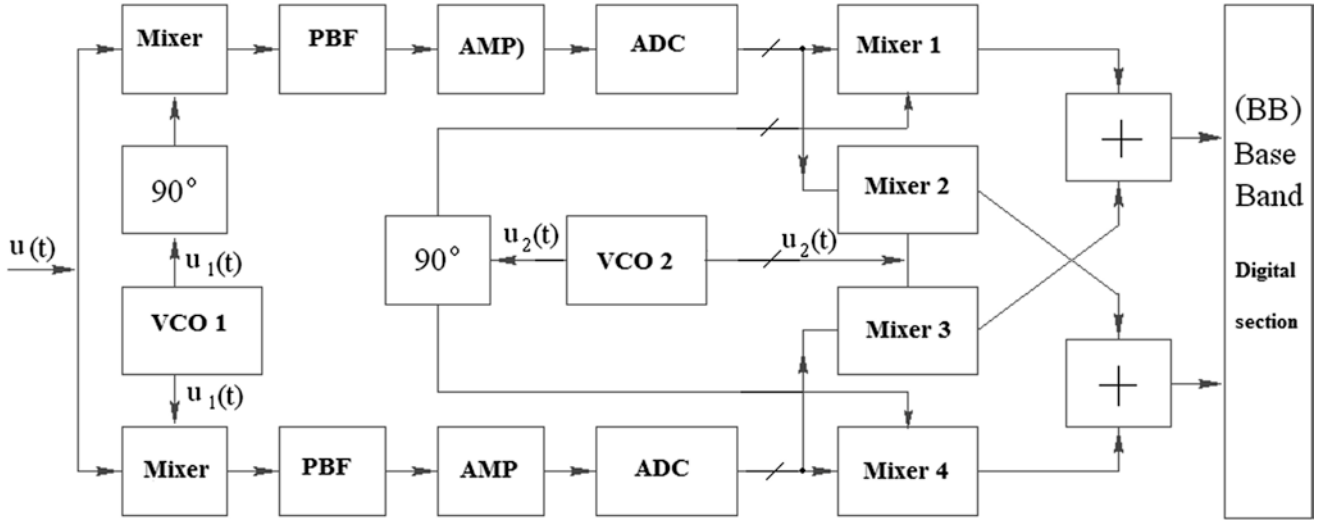


Fig. 3.23 The structural diagram of the receiver with input signal conversion on the low intermediate frequency with application of phase methods of the image channel suppression (the Weaver architecture)

If the following signal acts at the receiver input

$$u(t) = U_{in} \cos(\omega_0 + \Omega)t, \quad (3.23)$$

and the LO voltages are

$$u_1(t) = U_1 \cos \omega_1 t \quad (3.24)$$

and

$$u_2(t) = U_2 \cos \omega_{IF} t, \quad (3.25)$$

where $\omega_{IF} = \omega_0 - \omega_1$, signals passing in the LPF inputs has the frequencies: $\omega_0 + \omega_1 + \Omega$ and Ω . Taking into consideration the significant difference in frequency of these signals, the effective extraction of the modulating signal is achieved at not high requirements on AFC of the pass-band (low-pass) filter. After amplification, the analog signal is converted into the digital form in ADC and passes to inputs of mixers, which output voltages are the result of multiplication of low-frequency signal and the voltage $u_2(t) = U_2 \cos \omega_{IF} t$. The task of quadrature components' formation in the digital form is solved technically significantly simpler than at utilization of the signals in the analog form. At outputs of the digital mixers Mix1...Mix4, the following voltages are formed:

$$U_{mix1} = K_{mix} [\cos \Omega t - \cos(2\omega_{IF} + \Omega)t],$$

$$U_{mix2} = -K_{mix} [\sin \Omega t - \sin(2\omega_{IF} + \Omega)t],$$

$$U_{mix3} = K_{mix} [\cos \Omega t + \cos(2\omega_{IF} + \Omega)t],$$

$$U_{mix4} = -K_{mix} [\sin \Omega t - \sin(2\omega_{IF} + \Omega)t].$$

At the output of the first summer, signals from outputs of Mix1 and Mix3 mutually compensate the components of the image channel frequencies ($2\omega_{IF} + \Omega$), and components on the modulation frequency proportional to $\cos \Omega t$ are summed. The second summer creates the quadrature signal proportional to $\sin \Omega t$, simultaneously eliminating the signal with the frequency ($2\omega_{IF} + \Omega$). Non-high requirements to the low-frequency pass-band filters simplify their realization because the main requirement for them is the suppression of the output of the first pair of mixers of the summed component with the frequency ($\omega_0 + \omega_1 + \Omega$), as well as signals of the adjacent channels.

Another variant of the receiver structure with input signal conversion into low intermediate frequency using the Hartley architecture for the image channel suppression allowing realization the radio front-end on the single chip (Fig. 3.24) is based on application of the *complex (polyphase) filters* [96].

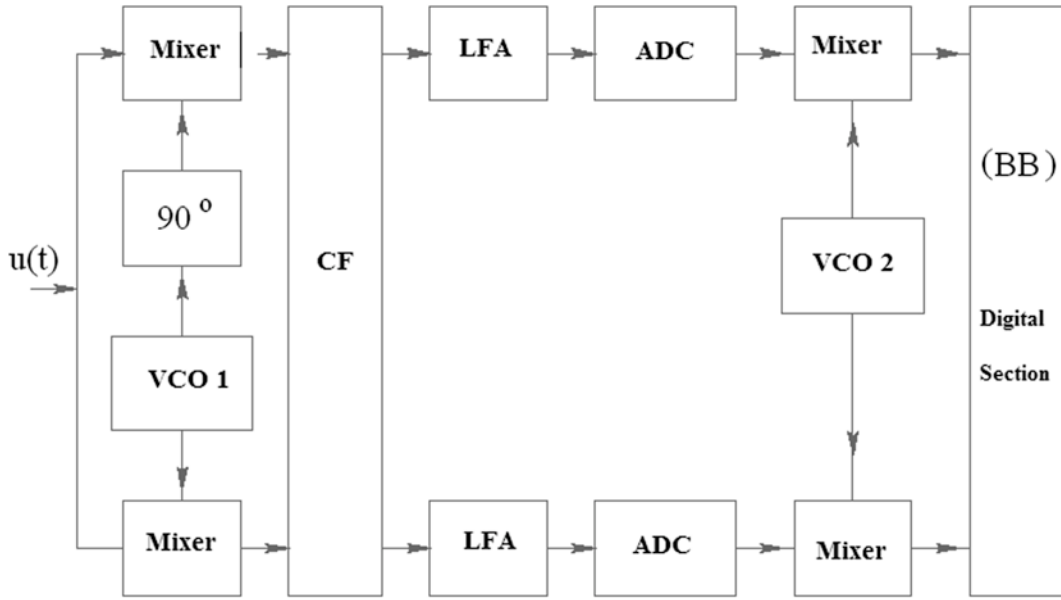


Fig. 3.24 Another variant of the receiver structure with input signal conversion into low intermediate frequency using the Hartley architecture for the image channel suppression

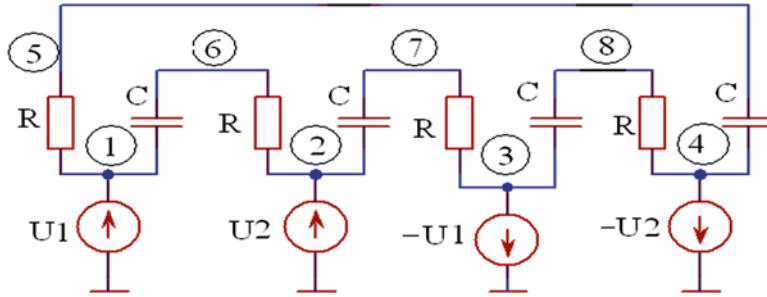


Fig. 3.25 The simplest structure of the complex filter

At action on the receiver input of the signal (3.23) and the output VCO1 voltage (3.24), the signal formed of the mixer outputs having the frequency values, $\omega_0 + \omega_{LO1} + \Omega$ and Ω , passes to the complex filter (CF), which simplest structure is presented in Fig. 3.25.

The in-phase U_1 and quadrature U_2 components (Fig. 3.25) creating on outputs of the first pair of mixers pass on inputs of the first and second filters:

$$u_1(t) = U_{in} \cos(\omega_{IF} + \Omega)t + \cos(\omega_0 + \Omega + \omega_{LO1})t,$$

$$u_2(t) = -U_{in} \sin(\omega_{IF} + \Omega)t + \sin(\omega_0 + \Omega + \omega_{LO1})t.$$

Voltages of required phase are passed in points 1–4 of the complex filter circuit.

Between points 5–7, the differential in-phase signal is formed, and between points 6–8, the differential quadrature signal. Assuming that the frequency of the second local oscillator is equal to ω_{IF} , the combination components $\omega_{IF} + \omega_{IF} + \Omega$ are suppressed in the digital section, and components with the modulation frequency Ω are extracted. It [97] is shown that the construction of the complex filter on the base of trans-conductive amplifiers – the cascade filter of 12th order – provides the image channel suppression not less than by 45 dB at filter realization on the base of MOS-technology and occupied area on the chip less than 1.7 mm² together with the control system.

Signals $u_3(t) = -u_1(t)$ and $u_4(t) = -u_2(t)$, shifted in phase by 180° with regard to voltages applied to inputs 1 and 2 of the complex filter, are applied in inputs 3 and 4. In integral implementation, the antiphase voltages with the same amplitude are formed by the phase-inversed stage on the silicon MOS-transistors with the separated load (Fig. 3.26).

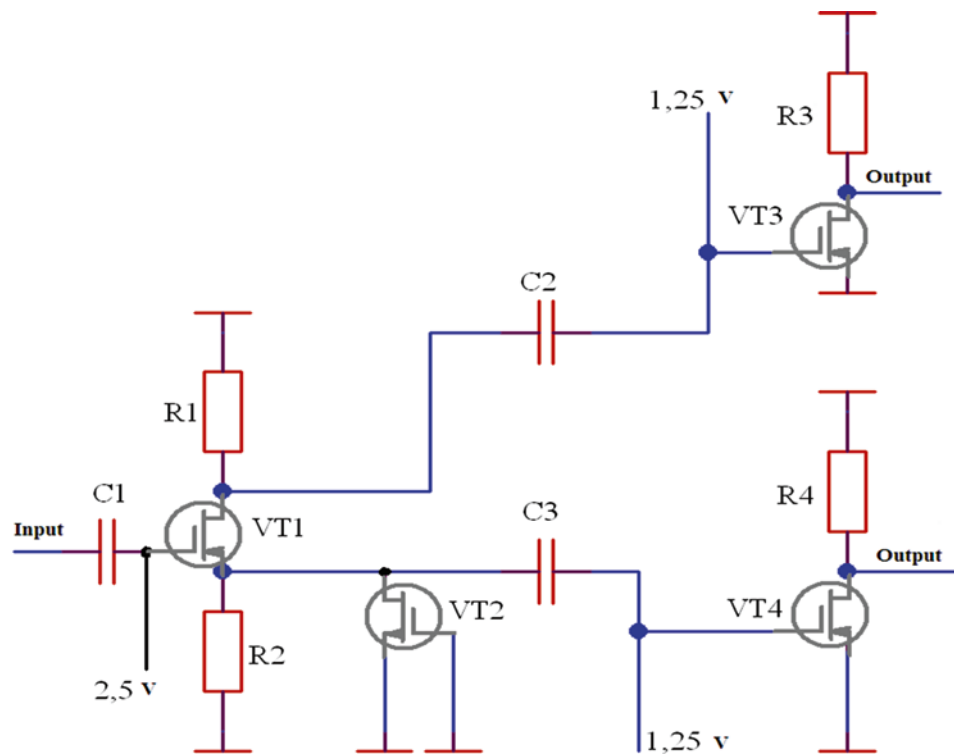


Fig. 3.26 The phase-inverted stage on the silicon MOS-transistors with the separated load

The CF parameter dispersion essentially influences on CF application on the image channel suppression degree. So, at mismatch in amplitude and phase between CF branches by only 1%, the image channel suppression is accordingly 40 dB and 35 dB. Therefore, if suppression degree in receivers with conversion to the zero frequency is the same in branches, then in receivers with conversion to the low intermediate frequency, the achievement the required SNF value in an input of the digital section should be estimated by the worst value of suppression in branches. To preserve the SNR, in absence of the identity of the arms of the quadrature structure in the CF and the equality of the signal levels on its differential inputs, this turns out to be a difficult task due to the limited accuracy of the implementation of the CF, caused by high requirements for the manufacture of MOS transistors.

This disbalance can be noticeably reduced increasing the ADC digit capacity from usual 12–14 to 14–16 bits. Another variant of mismatch reduction in branches of the I/Q-demodulator is utilization of the independent adaptive digital correction of delay durations and signal levels in in-phase and quadrature channels of the receiver.

3.2.6 Multi-Range and Multi-Standard Radio Receivers

3.2.6.1 Input ASN and FEM Modules of Transceivers

Mobile communication network of the third-generation IMT/UMTS, owing to significant increase of the data transmission rate and spectrum representation and spreading of rendering services, finds more and more numbers of followers, which increases the park of mobile devices of wideband access networks. At that, terminals developing for the 3G networks must render services of the mobile communication simultaneously in the second generation networks. This supposes an existence of technical opportunity for new terminals to operate not only in different frequency ranges but in various radio interfaces and in various radio access networks.

Creation of the cellular phone construction operating in different frequency ranges becomes essentially relevant in connection of assimilation of new frequency range of 1800 MHz (GSM1800) in the European part and implementation of personal communication systems in the range of 1900 MHz in countries of Northern America (PCS 1900) and in range of 1800 MHz: DCS1800.

In the simplest case, the device of section separation (at operation in the single-frequency range) represents the switch, which switches sections of reception and transmission.

At present, there are two types of devices for section separation: modules of switching and filtering realizing in the integral implementation as modules for antenna switch (ASM) and input front-end module (FEM).

Figure 3.27 shows the structure of the switching module ASM in the very simplest variant when we use the single-pole switch on six positions, which connects the single antenna input to six radio front-ends.

The switching module of the antenna has usually no elements of the pass-band filtering and serves for antenna switching to the transmitter output or to the receiver input at time channel division in the simplest case of the subscriber device operation in the single range. This switch should have low signal losses (insertion loss) in the closed state and to provide the well decoupling between the transmitter and the receiver. An example of such a switch for two positions receive/transmit is a switch made from GaAs FETs [98] in IC RF2436 have very low losses (less than 0.5 dB on the frequency of 900 MHz) are widely used in cordless phones of the DECT standard and in wideband computing networks (WLAN). The transfer function of this switcher is +27 dBm at reception and transmission section decoupling of 24 dB. The switcher RF2425 can be used for switching of two antennas for reception and transmission in the single or different frequency ranges (adaptive antennas). At operation in the frequency range from 0 to 2 GHz and for the power source voltage from 1,5 to 6 V from the unipolar source, it provides the transfer function on the 1-dB level of +31 dBm.

The classical FEM structure for the GSM system using the single antenna; two duplexers, which separate the reception and transmission bands; and two switchers is shown in Fig. 3.28. For operation in the range of 850 MHz, we use the switch

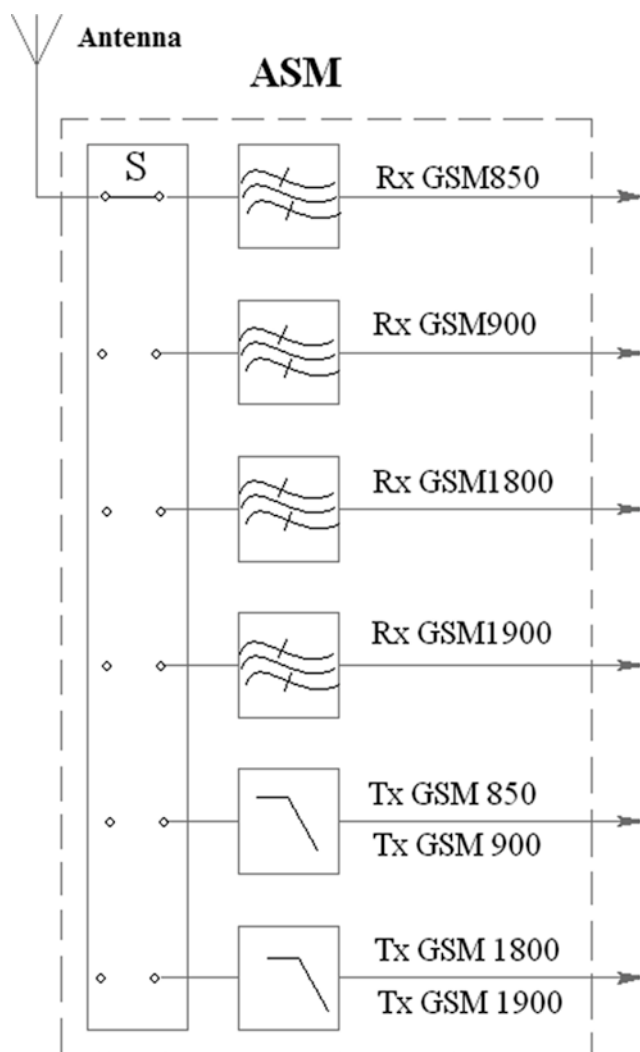


Fig. 3.27 The structure of switching module ASM in the very simplest variant, when the single-pole switch **S** is used on six positions connecting the signal antenna input to six radio front-ends

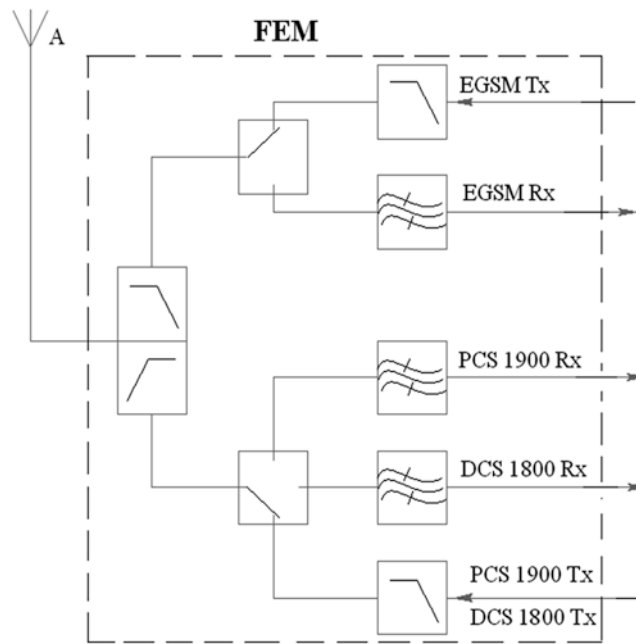


Fig. 3.28 The classical FEM structure for the GSM system using the single antenna, two duplexers, which separate the reception and transmission bands, and two switchers

on two positions, and for channel arrangement in PCS 1900, DCS 1800 systems, we can use the switcher on three positions and two duplexers, which separate two receiving sections of different frequency ranges. The considered structure has the increased loss level in the reception section caused by in-series connection of the duplexer and the switcher. Construction of radio-frequency FEM modules is complicated at subscriber terminal operation in different frequency ranges and at different standards. This relates to not only with a necessity to extract the appropriate operation range on reception and transmission but with the difference in energy plans of GSM and UMTS systems.

Even at operation with the single GSM standard, each reception range requires the input of the radio front-end in the insertion of the individual pass-band filter, usually, on SAW.

At present, the widest application at construction of mobile terminals have got receivers implemented with the single- or double-frequency conversion of the input signal. Application of the multiple frequency conversion in the *receiving section* is mainly caused by a necessity to have a preselector operating in the wide frequency range at enough high requirements of the collateral channel suppression. Besides, at cellular structure, which is typical practically for any mobile communication system, the intra-channel (so-channel) distortions arise besides distortions caused by the collateral reception channels.

In ASM structure, which is called the switching module, when the subscriber terminal operates in the single-frequency range, the switcher's number can be reduced to single (only on two positions), when an antenna alternately connects on the reception or the transmission. Depending on the specific manufacturer, the pass-band filters, for example, can be absent in ASM, and only duplex filters can be present when using the frequency duplexing, and in TDMA variants, the switcher can be absent (at operation in the single-frequency range).

The variant of FEM structure (Fig. 3.28) with the single antenna, two duplexers, and two switchers on four directions (Four-Band GSM Front-End Module), the one of which should provide three possible states.

3.2.6.2 Multi-Range Receiver of Direct Conversion

Implementation of new cellular communication standards based on the transmission methods and the basing equipment of the previous generation of the communication systems simultaneously required the modernization of subscriber terminals. They must simultaneously provide the speech traffic transmission and operation in networks of the wideband and wireless access of different frequencies and standards.

Increase of the data transmission rate and the growth of utilization effectiveness of the radio-frequency spectrum provide a possibility of obtaining by the subscriber of new types of mobile services. This is a consequence of more advanced tech-

nologies of the multi-station access and new types of modulation. From the point of view of the mobile terminal construction, the structure of the direct conversion receiver is optimal. The multi-range receiver providing operation of the mobile communication system in network of the GSM standard with different carriers of EGSM (Extended Global System for Mobile Communications), DCS (Digital Communications System), and PCS (Personal Communications System) is the example of realization such a structure. The difference in construction of such a receiver compared to the single-band receiver consists only in more complicate realization of the front-end module (FEM), and, relatively, after converter, the receiver structure does not depend on the chosen standard.

The super-heterodyne receiver using one or several carrier frequencies with extraction of the operating band for each sub-range and having the specified sensitivity and selectivity should include the comparably large variety of add-on devices (filters) and additional amplification stages, which increases the power consumption and the terminal cost will be increased. The need to change the frequency plan of the receiver during operation depending on the operating frequency range makes its structure unsuitable for use in multi-band receivers. From the technical point of view, this variant is the most complicated requiring even only for signal reception of the GSM900 standard and GSM1800 extraction of fourth frequency sub-ranges. The diagram of the receiver with the lower IF leads to a necessity of inserting of blocks, which provides the image channel suppression, which may lead to appearance of the blocking signal at the mixer output and simultaneously to degrade the noise characteristics of the radio front-end.

There are other variants for the radio front-end construction, for example, with parallel connection of receiving sections for the single sub-range or the standard in the single wideband section. Application of the super-heterodyne receiver architecture at integration the receiving section requires additional connection of wideband low-noise amplifiers at their inputs. However, it increases the noise power on the receiver input due to the front-end broadbandness. Addition of the wideband pass-band filter on the LNA input somewhat reduces the problem sharpness, but saves the adjacent channel interference influence per one channel, which leads to a necessity of more rigid filtering in other channels (Fig. 3.29).

Multi-range antenna (or two antennas on own frequency range) is connected to the band-pass filter (BPF), which restricts the LNA pass-band, on the output of which the additional pass-band filter (BPF1...BPF3) are connected, which extracts the operating frequency bands. The receiver of the mobile communication, which receives signals of the digital CDMA and the analog AMPS standards, is implemented on such architecture.

The interface part of this receiver is performed according to the parallel circuit and includes three channels for signal processing with the bandwidth of 925...960, 1806...1880, and 1930...1990 MHz (Fig. 3.30).

Each channel contains the frequency filter, the wideband LNA, and a mixer. For extraction of the first channel, the LPF is used, while for second and third channels HPFs. At that, zeroes of the transfer functions of filters realized as ladder structures are located on frequencies of image channels.

The direct conversion receiver used for operation in several frequency ranges satisfies the modern requirements to the radio front-ends at minimal number of add-on elements and the sections of signal processing. Application in the input of the direct conversion receiver of the wideband low-noise amplifier is the possible variant of the radio front-end structure at integration

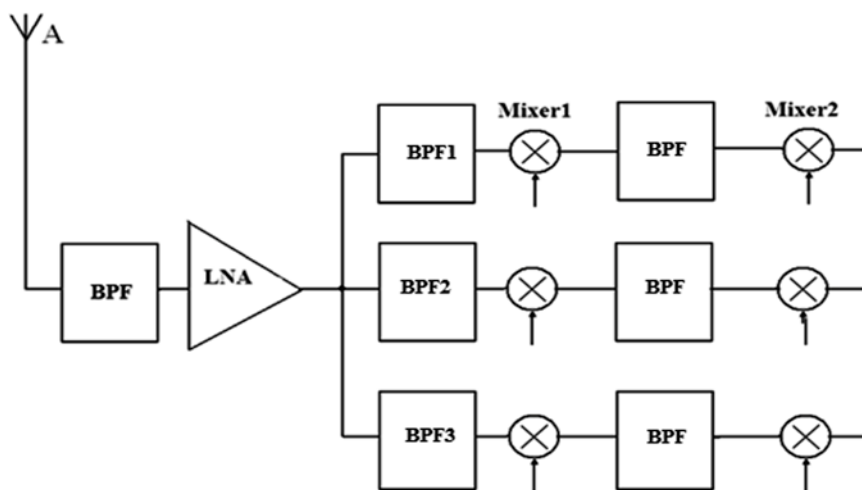


Fig. 3.29 Addition of the wideband pass-band filter on the LNA input

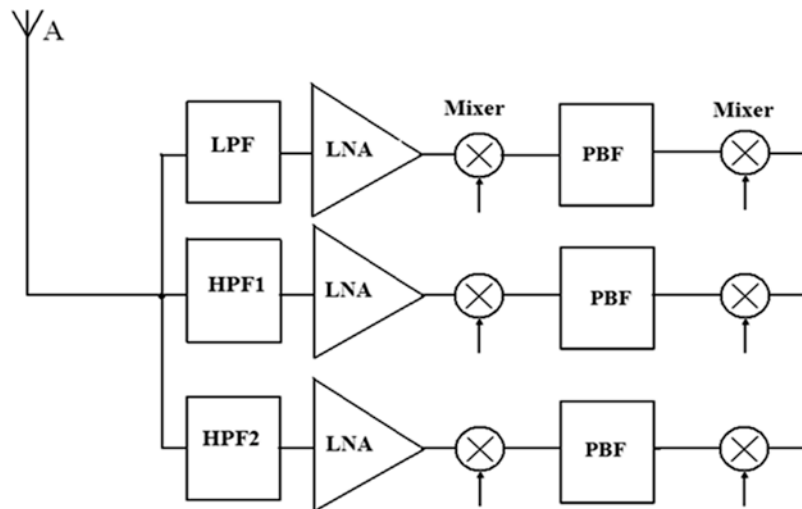


Fig. 3.30 The interface part of this receiver is performed according to the parallel circuit and includes three channels for signal processing

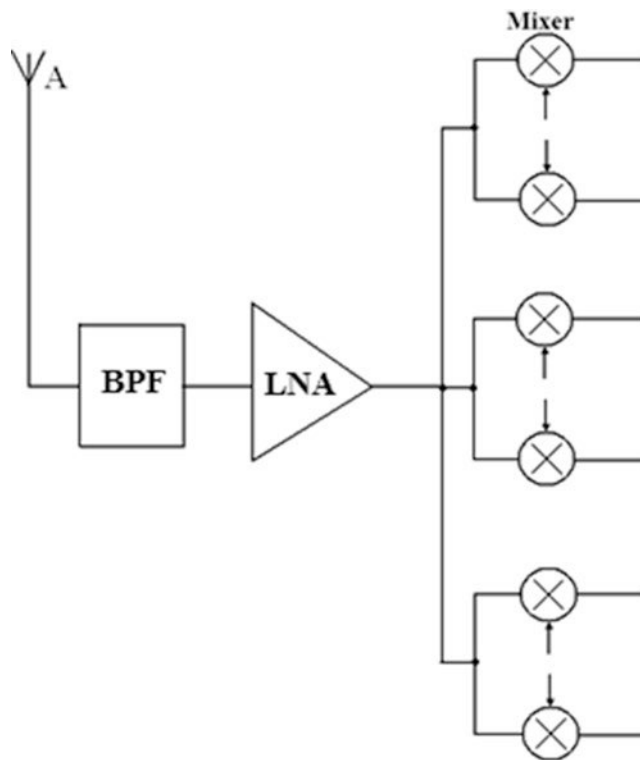


Fig. 3.31 The structure with application in the input of the direct conversion receiver of the wideband low-noise amplifier

of closely located ranges in frequency, which allows the reduction of their number to the one, decreasing the terminal power consumption compared to the previous case (Fig. 3.31).

The signal from the multi-range antenna is fed to the input of the band-pass filter (BPF), which reduces the level of interference from extraneous channels to acceptable values at the input of the LNA, from the output of which these are distributed to the inputs of the quadrature demodulators.

The multi-range transceiver construction for signals of GSM/EGSM/DCS/PCS standards with preliminary extraction of two frequency ranges is known also with further transfer of receiving signal frequency into the low intermediate frequency. The main difference at signal reception of a combination of GSM standards consists in the fact that VCO applied for

operation in the appropriate range is tuned in frequency with the help of multipliers and dividers providing the specified value of the ratio between the input signal carrier frequency and the local oscillator frequency. Even at transfer into a low intermediate frequency, when complex filters implemented using integral technology can be used to suppress extraneous interfering ranges, ensuring optimal performance of RF front-end when changing the operating range (bandwidth, passband ripple, etc.) is a rather difficult task. Therefore, main adjustments are performed in the modulation frequency section.

3.2.6.3 Multi-Functional Receivers

Implementation of multi-functional receivers operating not only in different frequency ranges, but with different standards (for instance, EDGE and UMTS), significantly complicates its architecture. To support standards with so different parameters, the modulation frequency bandwidth, the dynamic range, and the carrier frequency, the receiver must contain the high level of universality. In this case, the single IC should contain the multi-functional blocks providing the reception and processing of signals of different types of standards (as much as possible).

There are two main approaches to realization of the multi-functional architecture:

- Reception in several channels simultaneously (the concurrent system).
- Reception of the only single standard during the duration of the single symbols.

Simultaneous reception in several channels is the simplest and at the same time is the most technically difficult variant of the multi-functional receiver construction. At that, the complexity increases with the growth of a number of simultaneously received channels. The simplest variant of such a structure, for instance, for signal processing of the GSM and WCDMA standards, must contain at least two channels of signal processing of the modulating frequency: the first for GSM signal processing and the second for WCDMA signals. At utilization of quadratic signal processing, the number of channel increases to four. Evidently, the subscriber terminal using the principle of the super-heterodyne receiver, being simultaneously the multi-range receiver, can be realized (Fig. 3.29) with application of additional add-on elements: the pass-band filters, LNA. Construction of the multi-functional receiver with utilization of conversion to the low intermediate frequency is possible only for terminals operating in the GSM/DCS/PCS standards in connection with the fact that its IF value is about 200 kHz, while the frequency bandwidth in the UMTS standard is 5 MHz. As we mentioned earlier, there are definite difficulties at implementation of the complex filter on the converter output to the low intermediate frequency, and the transfer in the multi-functional mode essentially complicates the struggle with the image channel, since its parameters correction will be required at signal processing of different standards.

To develop the multi-functional receivers, it is necessary to retune the frequency-selective units to increase of sensitivity, the dynamic range, and so on, which is possible both in the intermediate frequency section and after the quadrature demodulator. However, retuning of the pass-band filters in the intermediate frequency section, which is usually realized as add-on elements, with saving of required characteristics is rather a complicated engineering task. This indicates that for the multi-functional receivers, the architecture of direct conversion of the input signal is preferable, when retuning of the frequency-selective block is performed in the modulating frequency section. The final architecture choice of the multi-functional receiver is defined by series of factors. The main one for the RF front-end is the front-end (preselector), which reduces the level of the LO signal, which leaks into the mixers input and creates a parasitic coupling between the units due to the equality of the input signal frequencies and the LO frequency. For the low-frequency part of the interface, the requirement of the wide retuning LPF range is determining.

The multi-functional receiver having the architecture of the direct conversion capable to receive signals of four RF ranges of GSM900, DCS1800, PCS1900, and WCDMA standards with two different modulating frequency bands includes the block of single-ended LNAs with variable gain, the quadrature demodulator and the local oscillator, the channel retuning LPF, and the channel amplifiers. The control of retuned block parameters is performed by the digital information section ("Base Band" section). LNA contains the four input and two selective stages. Depending on the carrying frequency value, one of the input stage is connected to the high-frequency input, while others are grounded, at that. At the reception of the GSM900 standard, the non-tuned resonance circuit is the load of the first stage, while at reception of other signal standards is the resonance circuit, which tuning frequency varies under action of the discretely adjustable circuit capacity. At outputs of quadrature demodulators, the channel filters of the fifth order are connected, which are implemented on op-amps as the active RC circuits. Adjustment of its parameters (the filter time constant, the bandwidth, etc.) is performed by discrete variation of the LPF capacitance by the control signals from the information section. The following amplifiers of the modulation

signal have the programmable amplification with the control step of 3 dB (or 6 dB) depending on the receiving signal standard.

Transfer from the speech message transmission to the high-rate data transmission, assimilation of new frequency ranges, and the modulation types simultaneously with the need of operation with different standards, using the single subscriber terminal, indicates the preferable architecture of its implementation – the receiver of direct conversion.

Such an approach of the receiver construction with signal transfer into the zero frequency is already actively used for WLAN, WCDMA, Wi-Fi, WiMAX, Bluetooth, UWB, and ZigBee systems.

For cellular communication GSM system, the signal transfer into low intermediate frequency is more often used with further signal processing in the I/Q-demodulator.

All mentioned system excluding Bluetooth do not require to apply the architecture of the parallel reception. Receiving the signal in the allocated frequency band (the bandwidth is defined by the radio access method and the approach of channel division for reception and transmission) in the definite time interval, the receiver occupies the single channel, while all others are in the waiting mode or switched off.

The main receiver blocks defining the carrier frequency of the receiving signal, the frequency band of the radio channel, and input of digital information channel should be retuned. Retuning of modulation frequency filter is provided simpler than in receivers with the transfer into low intermediate frequency or IF retuning in the super-heterodyne receiver. These circumstances mainly defined the application of the homodyne receiver architecture. An example of such a structure [64] is designed to receive signals (Fig. 3.32) of the CDMA2000, WCDMA and GSM900, GSM1800, and GSM1900 standards in the frequency ranges indicated in Table 1.2.

Modern subscriber radio stations usually used the single multi-range antenna operating simultaneously with several sections of reception and transmission. The subscriber radio stations of the cellular radio communication systems must provide the duplex operation mode (a possibility to simultaneously listen and to speak). The most distributed methods of the reception and transmission section division (duplexing) are frequency FDD and time TDD. On the base of chosen duplexing method of sections, the systems of multi-station access with frequency, time, or code channel division are implemented.

In the narrowband communication systems, the dual antennas (two antennas for the single operation range) are used for struggle with fading on the base stations. Such dual antennas usually have different patterns and are connected to the input of the transceiver through the switcher (Diversity Switch) of diversity antennas. The diversity reception increases the level of received signal by 3–5 dB. At that, antennas are located on the distance not less than of 10λ , where λ is the operating wavelength.

The input front-end module (InFEM) including, besides the antenna switcher, the devices for filtering realized usually as SAW filters may also contain the low-noise amplifier of the receiving sections and the power amplifier (LPA) of the transmission section. For operation of the receiver and the transmitter in the same frequency range on the single antenna in the FDD mode, the duplex filters (duplexers) are used. To separate the sections of different frequency ranges (for instance, GSM1900 and GSM1800 (Fig. 3.28)), the duplexers are used.

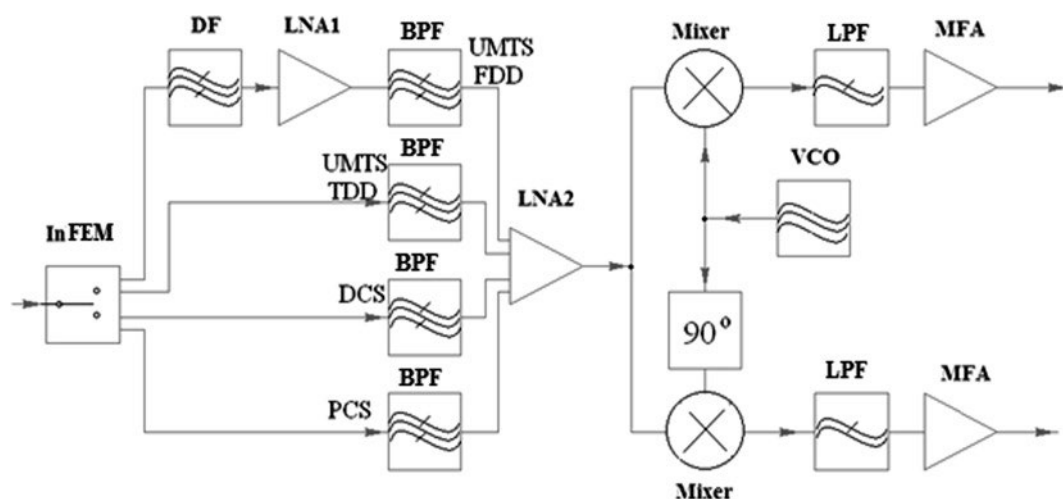


Fig. 3.32 The homodyne receiver architecture

The passive radio front-end elements are mainly defined by the geometrical dimensions and the cost of the subscriber radio station, and the consuming power is defined by the transmitter section. At present, passive components are located outside IC since their implementation on the unified integral technology turns out to be rather difficult or impossible at all. Geometrical dimensions of passive elements are commensurable with the IC dimensions, and therefore, they are realized as add-on elements, while to reduce their number, the developer usually tries to perform the terminal construction as universal for operation in several frequency ranges. We can attribute the duplexers, directional couplers, filters, inductors, and capacitors to such components. The reduction of a number of add-on elements in the radio front-end can be provided by application of the receiver architecture of the direct conversion at operation in several frequency ranges. A variant of the RF front-end architecture (Fig. 3.32) is considered as the main one in the design of subscriber terminals with several standards (GSM, UMTS). Transfer to operation in the more high-frequency range (system's Bluetooth, Wi-Fi, WiMAX), improvement of the integral technology (UHF volumetric MOП-structures) and development on its base of LNA and VCO allows realization of the planer inductors on the unified technology with the active elements [17].

3.3 Architecture Features of Receivers of Different Standards

3.3.1 The Receiver of the Super-Heterodyne Type for the Subscriber Radio Station of GSM Standards

The high-noise immunity of digital radio communication, radio broadcasting, and TV broadcasting systems compared to the similar analog systems is caused by utilization as the modulating signals of double-level and multi-level signals. The double-level modulation in the main frequency band becomes energetically more effective, and from spectral point of view, it is less effective compared to multi-level transmission systems. *Spectral effectiveness* or, in other words, the effectiveness of frequency band utilization is defined by the specific transmission rate, i.e., the quantity of transmitted bits per second per one hertz [(bit/s)/Hz]. This value is the important criterion of the system. So, for example, for double-level transmission system, using the Gaussian PSK system with minimal frequency shift (*Gaussian Minimum Shift Keying*, GMSK) with the LPF bandwidth on the modulator input of 81.3 kHz and the transmission rate of coded speech signal of 13 kbit/s, the spectral effectiveness is 0.16 (bit/s)/Hz.

Formation of digital stream. The multifunctional device is most popular among users of mobile communication systems, providing the ability to exchange data in real time with mobile and stationary correspondents while maintaining the function of providing subscribers with telephone and video communication services as well as sending short messages. Besides the realizations of mentioned functions, it should also have the low dimensions and mass and have batteries with the large time between recharging and the long operation in the waiting mode. The structural diagram of receiving and transmitting parts of the mobile station [6, 70] are presented in Fig. 3.33.

On the **transmitting** side, the variation of the audio pressure converted in the *microphone* into the electric signal passes to the ADC. According to requirements on the nominal range of the audio frequencies of radio stations, the upper frequency is the value not more than 3400 Hz, which is provided by the LPF with the cutoff frequency of 4 kHz, which is connected to the ADC input. For all mobile communication systems, the ADC discretization frequency is chosen of 8 kHz, which satisfies the Kotelnikov theorem with some margin, since it cannot be less than 6.8 kHz. The pulse sequence samples with the frequency of 8 kHz is converted in ADC into the 13-bit digital sequence (the point (I), Fig. 3.33), which defines the one from quantization levels (the total number of possible quantization levels is $2^{13} = 8192$). The digital stream rate describing the discretized and quantized analog signal by the number of bits per 1 s is $8000 \times 13 = 104$ kbit/s. In the GSM standard, the frequency-time channel division is used (TDMA/FDMA); when in the band of the single channel, eight subscribers can simultaneously operate, which are separated in time. Resulting rate of the digital signal is $8 \times 104 = 832$ kbit/s, and we understand that it is impossible to provide the digital stream transmission with such a rate in the frequency band of 200 kHz allocated by the single channel.

Decrease of the information stream rate in the GSM standard can be provided by application of the *speech encoder*. It is known that speech signal parameters are varied rather slowly, which is caused by properties of the human speech apparatus, and therefore, we can consider that in the interval of 20 ms, such parameters as the frequency, loudness, the voice tone quality do not change. In this case, we can transmit only these parameters, which will be performed by segments only 50 times per second, instead 8000. Each sample of the speech signal are coded by only 260 bits, that for single subscriber, the digital stream rate on the ADC output will be $50 \times 260 = 13$ kbit/s (the point (II)). Resulting rate of the digital stream for eight

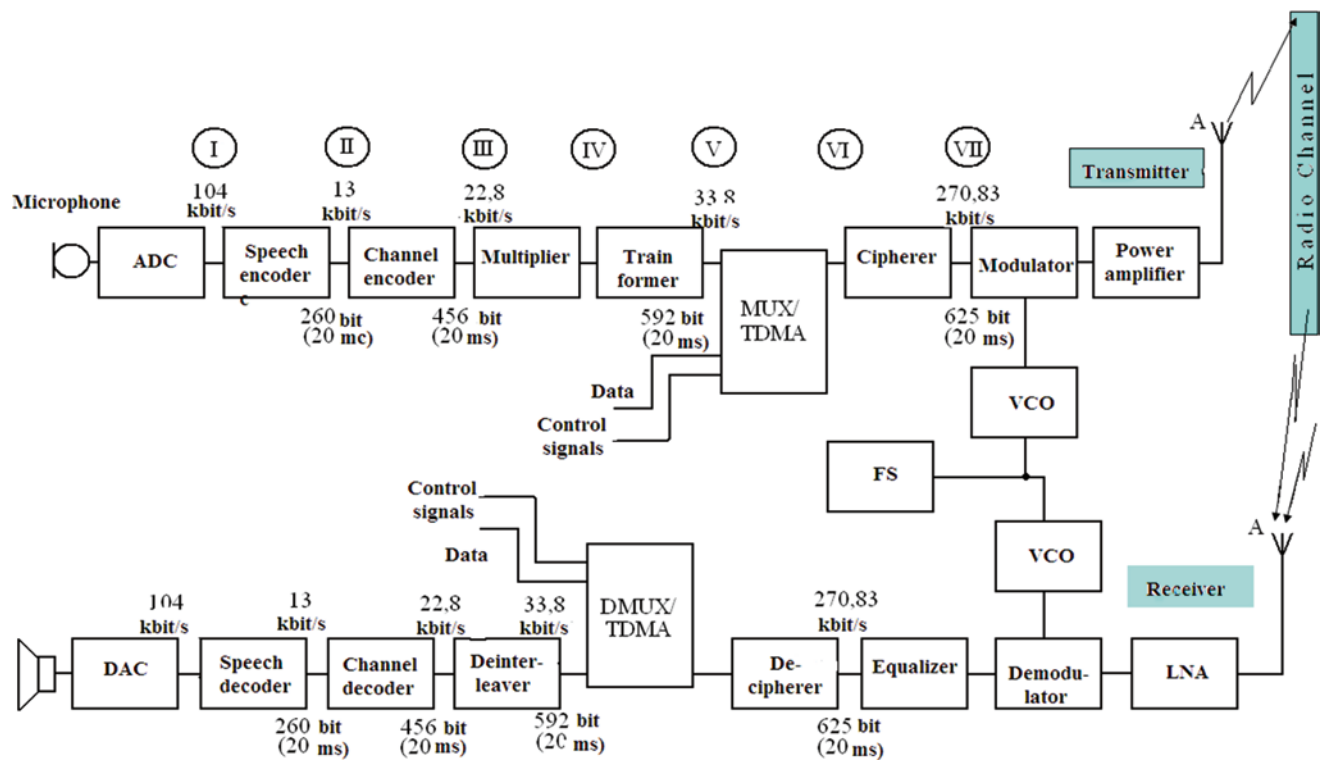


Fig. 3.33 The structural diagram of receiving and transmitting parts of the mobile station

simultaneously operating subscribers on the single subcarrier will be $8 \times 13 = 104$ kbit/s. At that, we can provide the speech quality according to requirements of the phone network of general use (PSTN).

Protection against group errors. The sequence acting in input of the *channel encoder* of 260 bits contains three bit blocks: 50, the very important bits; 132 bits, important; and 78 bits, not very important information. The first block from 50 bits is transmitted into the blocking encoder, where three additional bits are added to them, which are used for detection by the error even parity at message reception. Fifty-three bits from the first block and 132 bits of the second block and the trailer (4 zero bits) for a total of 189 bits are passed to the convolution encoder with exponent $R = 1/2$ and encoding constraint length 5. The total length of the two blocks is 78 bits. Together with the remaining 78 bits of the second block, which is not protected, they form 456-bit block (the speech fragment with the length of 20 ms). Application of the channel coding increases the digital stream rate to 22.8 kbit/s (the point (III)).

The output 456-bit stream passes to the *interleaving device* each 20 ms, which forms 8 blocks of equal length with 57 bits (Fig. 3.34).

In the transmitter, for a struggle with errors, which owing to fading are grouped into blocks, and for the effective decoding, the interleaving is used. The interleaving procedure consists of variation of generated coding sequence on informational and auxiliary bits before its passage into the radio front-end. Such a modified sequence, as any other, can be distorted in the channel by the impact of block errors. The influence of such a type of distortions can be essentially weakened by destroying of its block structure. For this, the interleaving devices are included in the transmitter, while in the receiver the deinterleaving devices. The interleaving devices can be divided into two main category: *blocking* and *convolution*.

In the *blocking interleaving device*, the binary sequence from the encoder output are recorded into the matrix with the random order of recording-reading of the multiplexer, which contains of 8 columns and 57 lines sequentially on lines, then they are read per columns (Fig. 3.34). Formed 8 segments with 57 bit each are distributed by eight sequential packets located in the allocated time slots.

In the *receiver*, the entering bits are saved in the same matrix, in which they were read in the transmitter. At matrix filling by received bits, they are read in the same order, in which they were recorded in the matrix of the interleaving device. Such a procedure of the recording consequence by lines and reading by columns leads to dispersion of error packets making them statistically independent.

Each of the blocks consists of bits shifted each other in the initial sequence by 8 bits, which allows effective struggle with errors grouped in packets due to fading, providing their dispersion, and increases the effectiveness of decoding. Blocks with 57 bits are integrated in pairs through additional learning sequence (Fig. 3.35) in bursts.

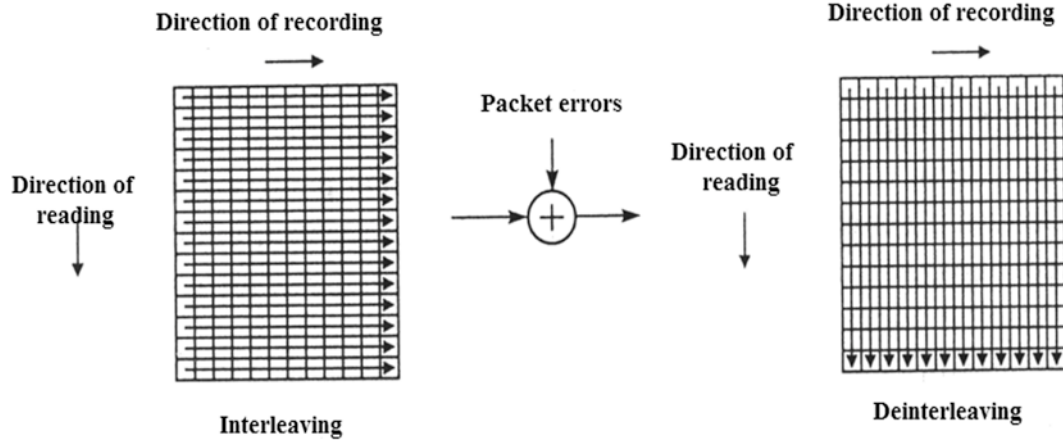


Fig. 3.34 The output 456-bit stream passes to the interleaving device each 20 ms, which forms 8 blocks of equal length with 57 bits

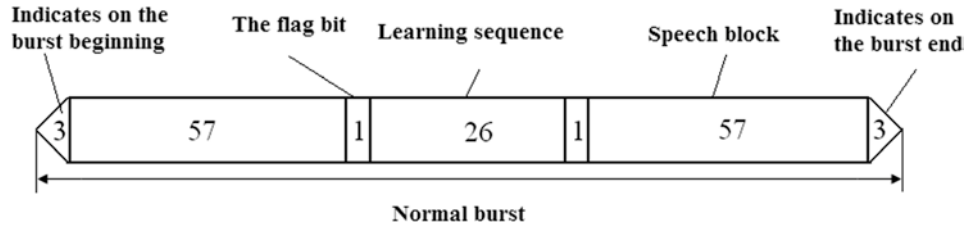


Fig. 3.35 The structure of normal burst

However, when the bits of neighboring speech blocks are placed in one packet, then when transmitting it in radio channel with a complex electromagnetic environment, packet loss increases the number of erroneously received bits by 25% (a quarter of all speech blocks), which leads to a decoder malfunction.

The number of errors can be reduced up to 12.5% if packets in one 20 ms speech fragment are formed not by combining two neighboring blocks (Fig. 3.35) containing 57 bits, but to replace one of them taken from a neighboring packet. Then a normal packet from the traffic channel (TCN) will include 8 segments of 57 bits of the current and previous block, and the last 4 packets include the bits of the current and next block. This leads to appearance of speech signal delay by 20 ms for the mobile station. However, under unfavorable transmission conditions, the burst loss means the loss of only one block, which twice reduces the BER. This index (12.5%) is the maximal error level, at which the channel decoder operates correctly. The moment of the burst start is designated by three bits; the place of digital sequence bifurcation (block substitution from another burst) is designated by the flag bit.

Including learning sequence is caused by a necessity to increase of signal noise immunity at the information exchange between the base station and the subscriber radio station. As we see from Fig. 3.46, this is provided by addition in the burst besides two blocks, carrying the speech information or data, of 26-bit learning sequence: 6 end bits and 2 flag bits (formatting). This increases the burst length from 114 to 148 bits and to the growth of digital stream rate to 33.8 kbit/c (the point (VI)). Totally, the bursts' formatting process increases the number of bits in the single segment of 20 ms from 456 bits to 592 bits.

Another method of the struggle with the error packets is based on application of the convolution interleaving device (Fig. 3.36).

The binary sequence applied per bits by a switcher to inputs of B parallel registers. Each i -th register ($i = 1, \dots, M$) delays of input bits by $(i-1)M$ cycles of the synchronizing signal. Output delayed signals are applied to the channel input through the second switcher. Deinterleaving device (the switcher together with receiver registers) distributes the received bits on registers according to the value of the delay duration, which is defined by a number of synchronizing signal cycles. The upper register delays the sequence in the system of receiver-transmitter by $B(1-M)$ cycles, while the lower register does not insert

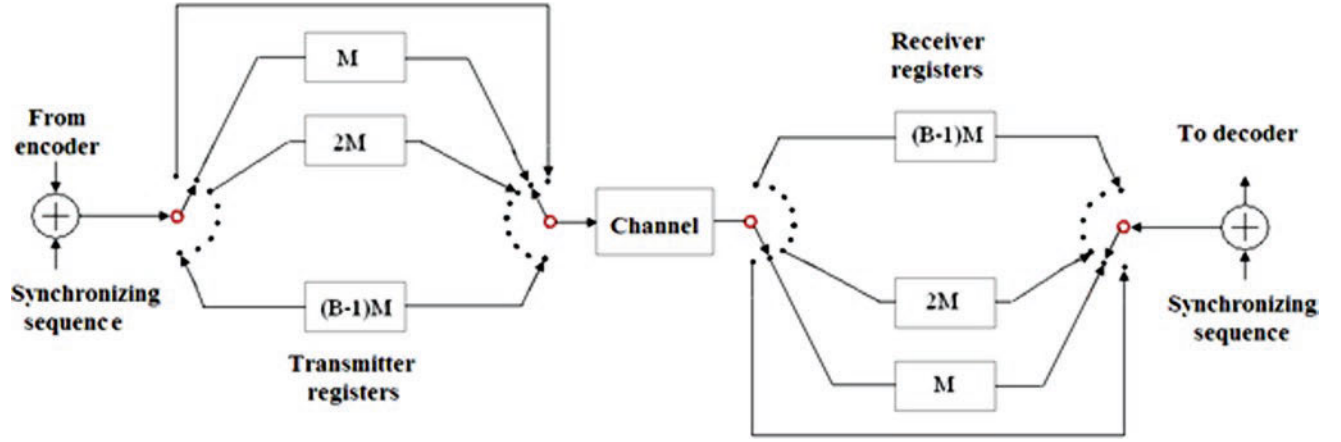


Fig. 3.36 Another method of the struggle with the error packets is based on application of the convolution interleaving device

the delay. The delay inserted by the stage of each parallel level of interleaving device and deinterleaving device becomes the constant quantity equaled to $B(1-M)$. Evidently, that as in the case of blocking interleaving, the synchronization plays an important role in the system receiver-transmitter, which provides the high performance of decoding algorithms of blocking and convolution codes.

At duration 0.577 ms of the single time slot of the TDMA cycles, we can provide the transmission of 156.25 bits (the duration of each bit equals to $T_b = 3.7 \mu s$) at the burst length of 148 bits. The remaining free time segment of 8.25 bits is called the protection interval (guard period), which is used to power control consumed from the power supply.

Another very important indicator is estimate of power arriving at the mobile station from the BS interacting with it and from three BSs of neighboring cells. The level of signals received over 16 BCCH broadcast control channels from the DS is estimated in the MS, and its value is transmitted through the opposite channel to the active BS. The BS interacting with it estimates the interference state of the channel by slots that are free at this moment and makes decision about changing the level of radiated power of its own transmitter and the transmitter of active MS.

The ciphering block does not add bits in the digital sequence performing the permutation of messages inside bursts.

At output of the ciphering block, the formatted burst contains 156.25 bits and accordingly on the standard 20 ms-segment 625 bits. This corresponds to the digital stream rate on the modulator input of 270.9 kbit/s (the point (VII)).

In the receiving part of the structural diagram of the subscriber radio station (Fig. 3.33), there are units that perform the functions to those described in the transmitter: demodulator, deinterleaving, decoding devices, etc.

Struggle with the inter-symbol interference. The signal in the input of the *receiving* device arrives via several ways exposing an attenuation and an amplification exposed to distortions, which is defined by the multipath channel nature. For transmitted signal $s(t) = A(t) \cos [\omega t + \theta(t)]$, we can receive the signal $r(t)$, which has a form:

$$r(t) = \sum_{i=0}^n \{x_i(t-\tau_i)A(t-\tau_i)\cos[\omega(t-\tau_i) + \theta(t-\tau_i)] + y_i(t-\tau_i)A(t-\tau_i)\sin[\omega(t-\tau_i) + \theta(t-\tau_i)]\}, \quad (3.26)$$

where τ_i is the time delay of i -th signal with regard to the signal arrived on the direct visibility way.

We know that the receiving signal $r(t)$ is exposed by attenuation describing by the Rayleigh distribution. If the receiving signal is enough narrowband closed in characteristics with channel properties, then attenuation of its components $x_i(t)$ and $y_i(t)$ will also obey to the Rayleigh distribution law. If several ways of signal passage is enough evidence, for example, reflection from a mounting, a hill, buildings, or similar structures, then the distribution function of signal passage ways is close to the Rician law [4]. If the time delay τ_i for i -th beam propagation is small compared to the bit duration, then the received signal is exposed by insignificant distortion. At $\tau_i \gg T_b$, the superposition of transmitted symbols occurs, which leads to the inter-symbol interference in the receiver input.

In third-generation broadband data transmission systems, noise-like signals arrive at the receiver input in different ways and have different latency in different ways. Vector addition of the main signal and its copies result in multipath fading. In addition to interference caused by feeding, the signal in radio channel is exposed to intersymbol interference due to the narrow modulating uses 0.26 μs in the BS transmitter (in UMTS system).

To eliminate the inter-symbol interference, we can either exclude the multipath propagation or apply the engineering solution making the receiver eliminate the influence of inter-symbol interference. The simplest example of decreasing of interference distortions is application of the channel model describing by the pulse response $h(t)$ at signal processing, which takes into consideration its multipath properties. If the receiving signal is described by the following equation

$$r(t) = \int_{-\infty}^{\infty} s(t)h(t-\tau)d\tau, \quad (3.27)$$

we can recover the transmitted signal $s(t)$, if we know the transfer function of the channel $h^{-1}(t)$, which is an inversion of the $h(t)$ response. The main difficulty of this function obtaining is its dependence in time. The diagram (Fig. 3.37) allowing the transfer function obtaining is called the *equalizer*.

From a number of 26 bits of the learning sequence (Fig. 3.35), there are eight preliminary-known bits of the mobile station (it is programmed at equipment manufacturing). The base station requests this sequence from 8 bits from the subscriber radio stations. In the specific cell, there is the *learning sequence*, which is an etalon for all subscriber radio stations and base stations. Received learning sequence is compared per bits with the known etalon, and the comparison results are directed into the summer. At coincidence of the first bit of the received learning sequence with the etalon, the result passes in the adder, which forms the coefficient $a_1 = 0$, then the comparison occurs of the second bits and the value of a_2 (zero or unit), etc. The current sequence, which is differential between received and etalon, is formed on the output of the detector (Det). The structure of the training sequence in the radio channel under the influence of interference can change, and hence the correct reception of speech signal can be disrupted. If a zero sequence is formed at the output of the detector, this indicates that the training sequence was received without errors, but if coefficients other than zero appear, this indicates the presence of error in the speech sequence. However, for formed coefficients $a_1, a_2, \dots, a_n$, which are now non-zero and their values are defined by the channel properties, they insert an amendment at formation of the speech sequence, which will make it more exact. Since their appearance is caused by the interference, then its influence will be now taken into account at formation of the speech signal transmitted in the same burst, when channel properties can be considered as non-changed. The accuracy of the speech sequence recovery (as the learning sequence) is defined by the rms deviation of the receiving learning sequence from the etalon and always is the value less than 100%.

The *data* transmission in the standard version of GSM occurs with rates 2400, 4800, and 9600 kbit/s. For each these rates, we apply the appropriate convolution code with the *coding coefficient* from 1/6 to 1/2. As in the case of speech transmission at all mentioned data processing (quantization, speech coding/decoding, the channel coding/decoding, interleaving/deinterleaving of blocks), the time delay appears. However, the inserted time delay in this case is not so critical. The more important problem at data transmission is the decrease of the error appearance frequency, which is achieved by appearance of the channel coding with error correction and deeper interleaving.

The multiplexer (MUX/TDMA) forms the resulting digital stream in the modulator input, consisting of the speech signal in the binary form, the digital data stream, and auxiliary symbols (Fig. 3.33). The signal from the multiplexer output passes to multiplier (the modulator), in which the harmonic signal simultaneously passes from the frequency synthesizer (or from VCO), which parameters are modulated by the digital sequence. As a result, the digital signal having the more compact

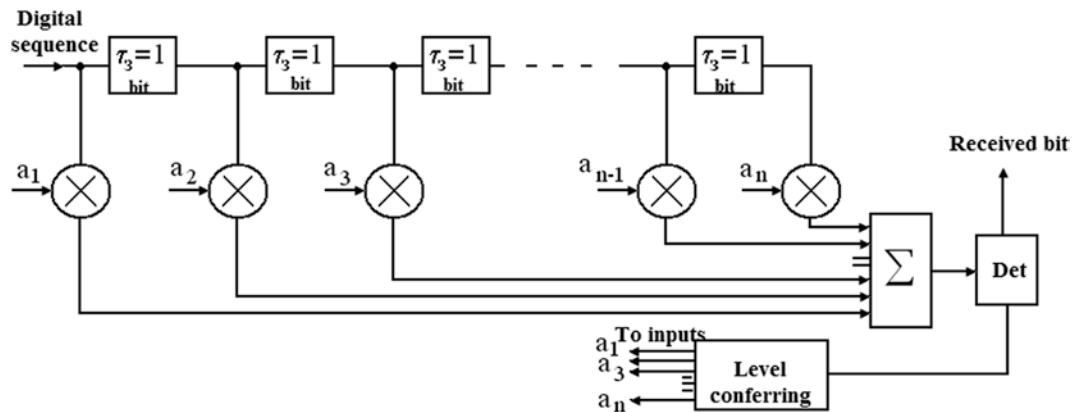


Fig. 3.37 The diagram of the equalizer

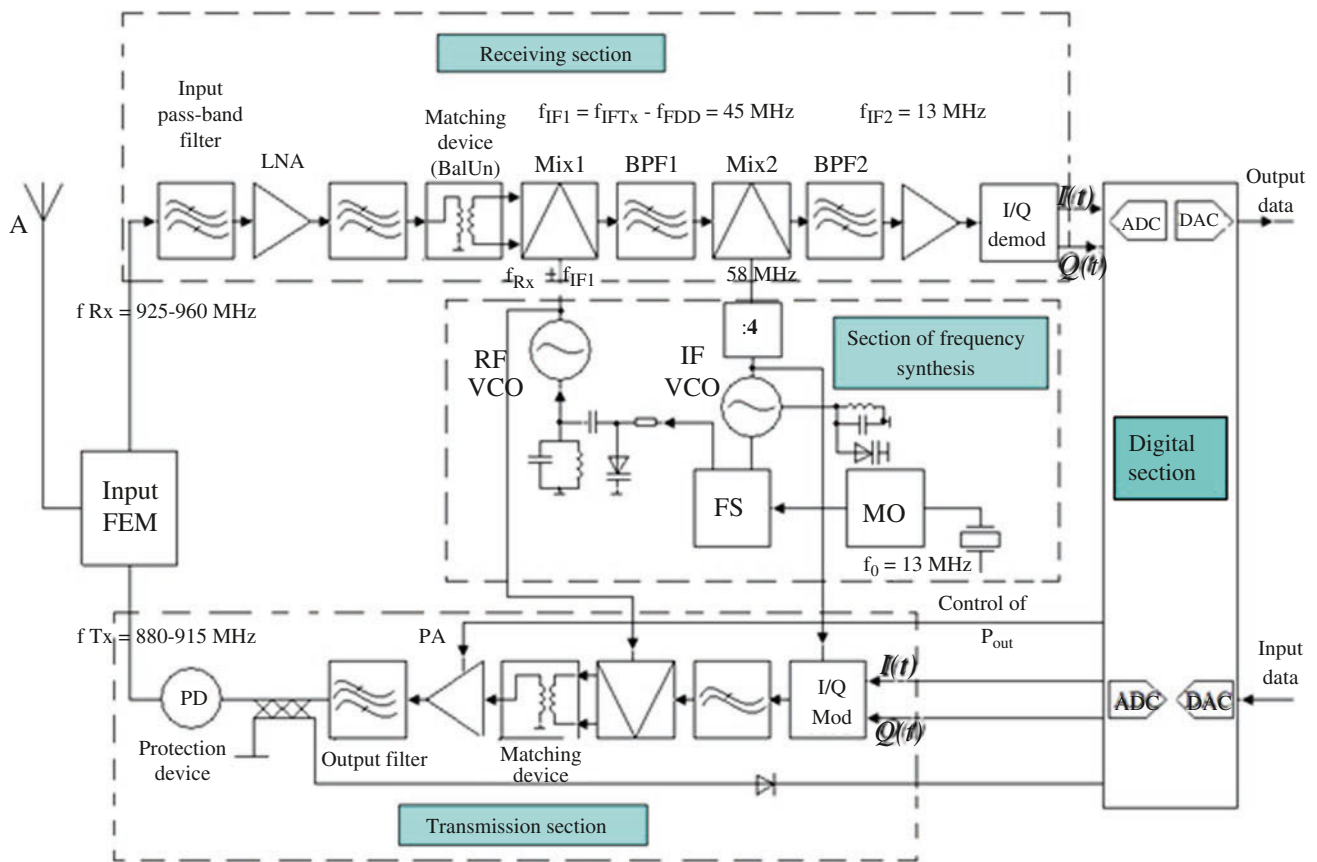


Fig. 3.38 The architecture of the modern transceiver

spectrum is transferred into the radio frequency range, and chosen type of modulation should provide as far as possible the best radio signal protection from the interference impact.

Structural diagram of the transceiver. In modern receiver-transmitters (transceivers) of the base and mobile stations of the cellular communication networks of the GSM standard independently on the specific implementation and construction principles of separate units, we can choose the *four main sections* (Fig. 3.38):

- Transmission section.
- Receiving section.
- Frequency synthesizer section.
- The digital (baseband) section.

The radio-frequency blocks with the single and double (triple) operating ranges of the cellular communication systems of the GSM and CDMA standards, as the trunking systems, have the unified construction principles, and the main differences are observed only in their architecture.

Figure 3.38 shows the structural diagram of the single-range RF block of the GSM900 standard [94]. The transceiver architecture for the base station BTS and the mobile station MS are identical, and the difference consists only in realization of separate blocks. For instance, the BTS antenna of the complicate construction mounted in buildings or the highly raised masts, providing the electromagnetic field conversion into EMF applied with the help of the feeder to the receiver input. For MS, this is miniature embedded (for instance, into the cellular phone) dipole. In considered example of the single-range transceiver, the *input FEM* is a duplexer, which separates the bands of received and transmitted frequencies. For described

system, the duplexing band is 45 MHz. The duplexer can be realized as two filters: the low-pass filter and high-pass filter or as two band-pass filters.

Transmission section. In the considered variant of transmitter section architecture of the transceiver, the double-frequency conversion is used, which decreases the power amplifier (PA) and the load influence on the VCO parameters. In the GSM standard, the Gaussian frequency shift-keying modulation with minimal phase shift (GMSK) is used, at which the binary sequence passing to the modulator from the ADC output of the digital section passes through the LPF with Gaussian characteristic. This allows the essential decrease of frequency band of the transmitted signal. To increase the noise immunity of the transmitted signal, the quadratic modulator is used. Data stream passing on the modulator from the ADC output of the digital section is separated into two streams: the first one is in-phase ($I(t)$); the second one is quadratic ($Q(t)$). At that, each stream has the symbol rate equaled to a half of the input bit rate. These streams pass to multipliers: on the input of I -channel multiplier simultaneously, the carrier signal $\cos(\omega_0 t)$ is applied, while on the second input of the Q -channel multiplier, the quadratic carrier ($\sin(\omega_0 t)$) is applied. On multiplier outputs, the signal of the I -modulator, which has a phase of 0° or 180° , is summed with signals of the Q -modulator with phase of 90° or 270° , forming the four-phase signal. The signal limited in the band passes to the second frequency converter (FC) and then to the power amplifier. The output filter restricts the band of transmitted frequencies and suppresses out-of-band conversion products. The check for transmitter operation (output power measurement and control, alarm signaling, etc.) is performed in the digital block, where the energy part passes through the directional coupler from the power amplifier output.

Reception section. In the input of the receiving section, which is constructed with double-frequency conversion, the pass-band filter is connected, which together with the filter on the LNA output provides the required suppression of the image channel. The pass-band filter on the LNA output is usually used for the specified frequency range from the composed materials or with utilization the acousto-electric transducers on the base of piezo-ceramics. Utilization of the additional inter-stage filter allows reduction of spurious (adjacent) channels influence, which is produced by other system subscribers.

The circuits of a differential or ring converters, which are usually implemented at diodes or transistor differential stages, have a symmetric input/output. Such converters (MIX1) require the use of matching devices at transistors or passive components at their input/output. For example, matching transformer BALUN (**BAL**ance- to **UN**balance) is used to connect the non-symmetric LNA output to symmetric input of the convertor. In the presented diagram, the first intermediate frequency $f_{IF1} = 45$ MHz is chosen to be equaled to duplex (FDD) shift between frequency of transmission and reception, which is used in duplex and semi-duplex communication systems, at operation of the receiver and transmitter on the single antenna. In the general case, the value of the first intermediate frequency in the receiver of the super-heterodyne type with the double conversion is chosen from the condition of absence of components in the Mix1 conversion products, which are the collateral channels of Mix2 conversion (in more details, see Chap. 11). This allows retuning of the radio-frequency block with the help of varicaps, using the only one RF VCO (in the part of receiver and transmitter sections, which do not deal with the intermediate frequency). The composite filter on the Mix1 output decreases the level of the converted frequency signal outside of the useful signal band.

The second converter (Mix2) realizing usually on multipliers, which are used the Gilbert cell (multiplier on the base of differential stages) of double-gate FET, is loaded to the pass-band filter (piezo-ceramic, piezo-electric). The LO voltage with the fixed frequency of 58 MHz on the second input of FC2 is obtained from the frequency synthesizer by its frequency division by four. The LO frequency adjustment to obtain the exact value of $f_{IF2} = 13$ MHz is provided by the PLL system. With the help of the pass-band filter on the Mix2 output, the required adjacent channel suppression is provided (at frequency method of channel division, see Table 1.2). After amplification, the signal with the central frequency f_{IF2} passes to the input of the I/Q demodulator. Since modulating signals $I(t)$ and $Q(t)$ are in quadrature (orthogonal), then the receiver can demodulate and regenerate them independently on each other, operating factually as two receivers. The signal conversion in the I/Q demodulator is performed to the zero frequency, i.e., the signal with the complex envelope is formed on the output, which contains information about an amplitude and a phase on initial modulated signal. Then, regenerated sequences **I** (In-phase) and **Q** (Quadrature) are again integrated into the parallel-in-series converter to obtain the initial data sequence used for further processing in the informational section. Here, the decoding of the signal occurs, its detection and error correction, etc.

At present, in radio communication systems, the structures of the active suppression of the collateral reception channels (image channels) find the great distribution, which allows simplification of the filter construction in the load of FCs.

The frequency synthesis section. In modern receivers, the different synthesis methods of the discrete variety of LO frequencies based on the single etalon frequency are widely used. The required values of nominal frequencies obtained by means of conversion are called the frequency synthesis. For this, we can use operation of adding, subtraction, frequency

multiplication, and division, which can be provided with the help of appropriate devices. The multiplication operation can be performed with the help of the PLL control and VCO. The synthesis device fulfilled in the form of the functionally complete block or the device is called the frequency synthesizer (FS).

In RF blocks of the communication systems (Fig. 3.38), the single reference signal (clock signal) is usually used for all frequency synthesizers, which is obtained from the reference oscillator (RO) stabilized by the quartz. As these reference oscillators in the stationary radio blocks of the large systems, BSS, the base station systems, and BSC, the base station controller, we can often use the voltage controlled crystal oscillator (VCCO) or the high-stable temperature-compensated crystal oscillators.

Frequency tuning of the RF block is provided from the FS, which controls the RF frequency of VCO with the help of the varicap in the circuit of VCO, which frequencies are varied in the range 970...1005 MHz. The second frequency synthesizer or another output of the single FS is applied for creation of the reference frequency voltage of 232 MHz. The auxiliary signal used in the receiving section (45 MHz) and another for quadratic reference frequencies for the quadratic I/Q modulator of the transmission section are formed from the reference frequency.

The double-range receiver. Figure 3.39 shows the receiving part of the radio front-end of the mobile station of the GSM standard, which receives signals of two frequency ranges.

The signal from an antenna, through the input RF module containing the switcher, which connects an antenna and the receiving and transmission sections, and duplex filters, passes to diplexers. Diplexers are the pass-band filters of LPF and HPF, which separate the one receiving frequency range from another. The signal from the output of the frequency converter passes to the pass-band filter, which extracts the intermediate frequency $f_{IF}=360$ MHz, which is mutual for both operating ranges, which is provided by the RF VCO frequency choice controlled from the digital section. The output signal is passed to the I/Q-demodulator, on which output two orthogonal digital sequences $I(t)$ and $Q(t)$ are formed, with which help the detection of signal and error correction, and control signals are produced, etc.

Formation of the global network of mobile communication assumes the device operation in networks of different destinations (WWAN, WMAN, WLAN) and various frequency ranges. This complicates the architecture of receiver radio front-end of the super-heterodyne type by increasing of the number of nodes, which backups the performed functions, which degrade the mass-dimension indices of the device and increases its cost. The alternative is the transfer to the receiver structure of the direct conversion (Fig. 3.40), which has the simple analog part, and all procedures of signal processing are performed in the digital section.

Destination of the input RF module is similar to above-considered diagrams. The pass-band filter in the input of LNA has the bandwidth, which allows extraction of all operating frequency bands of the receiver, which slightly reduces its sensitivity. The demodulator provides the separate processing of in-phase and quadrature components of the input signal, and the signal spectrum transfer to the zero frequency eliminates the problem of the image channel. Application of the coherent detection permits to distinguish of the useful signal from any other from the similar systems.

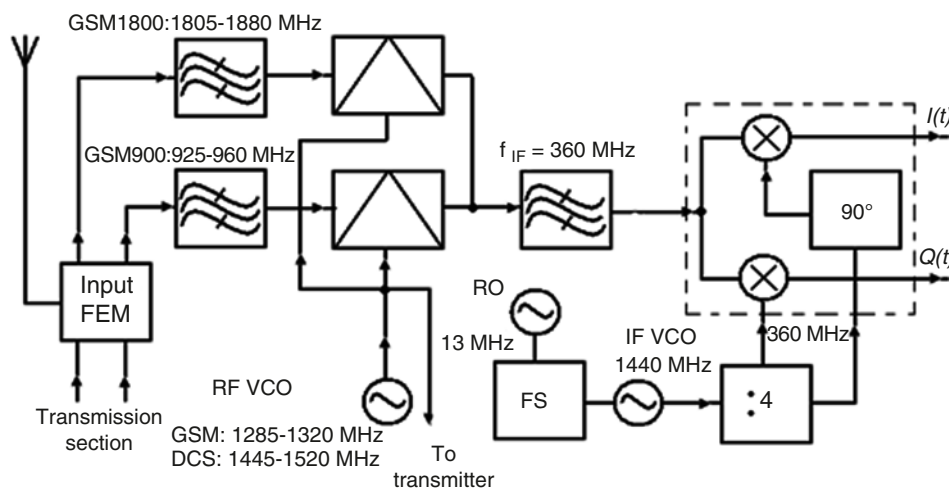


Fig. 3.39 The receiving part of the radio front-end of the mobile station of the GSM standard, which receives signals of two frequency ranges

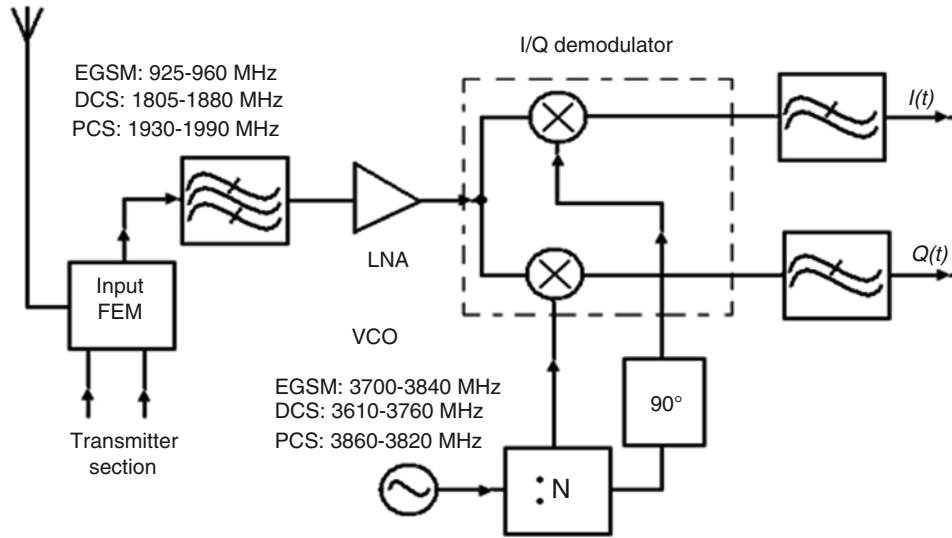


Fig. 3.40 The alternative is the transfer to the receiver structure of the direct conversion, which has the simple analog part, and all procedures of signal processing are performed in the digital section

3.3.2 The Subscriber Receiver with the CDMA Technology

3.3.2.1 Radio Interface in a System with CDMA Technology

The technology of *spectrum spreading* by the method [9, 10] of the direct sequence (DS-SS) is main in UMTS and IS-95 (CDMA2000), and it realizes by multiplication of the informational sequence with the pseudo-random sequence of bits called “chips” (elementary symbols). Resulting process has significantly higher speed compared with informational sequence, and the signal on the transmitter output looks like as a noise. When we transmit the signal formed by several subscribers on the same frequency, the digital sequence of each subscriber should be widened by its own pseudo-random sequence (PRS). On the receiving side, the informational stream of each subscriber is recovered by multiplication of all receiving signal by the known in advance (marker) sequence. At fulfillment of the condition that the rate of spread signal (the chip-rate) is essentially more than the bit rate of the informational sequence, noises created by the presence of other subscriber signals (interference noises) will be hardly noticeable at recovering of the digital stream of this subscriber. At increase of a number of simultaneously operating subscribers in the network, this leads to increase of interference noises, which, in the end, makes impossible to receive the subscriber digital sequence with required accuracy. In other words, to recover the subscriber data, it is necessary to have enough high SNR. In CDMA, to estimate the noise immunity, we use E_b/N_0 , where E_b is the power spectral density per one bit of recovered sequence and N_0 is power spectral density of the noise.

The ratio of the chip rate of spread PRS to the bit rate on the informational sequence (3.28) is called the spreading factor SF, which essentially influences on probability of the correct reception:

$$SF = \frac{\text{chip rate}}{\text{bit rate}} \quad (3.28)$$

Its growth increases the probability of correct recovering of transmitted sequence. In terms of transmission and reception, increase of SF is equivalent to the transmitter power growth. Therefore, the value of SF can be considered as some type of amplification called the benefit from the signal processing, **PG** (Processing Gain), expressed in decibels:

$$PG = 10 \lg \left(\frac{\text{chip rate}}{\text{bit rate}} \right) \quad (3.29)$$

In some cases, this quantity can achieve the large values and helps to overcome the interference influence in the receiver input, which are created by other subscribers. For instance, at $PG = 20$ dB and required value of $E_b/N_0 = 5$ dB, the receiving signal can be still recovered, if the ratio signal/interference noise is less than -15 dB. Evidently, at the constant chip rate, PG

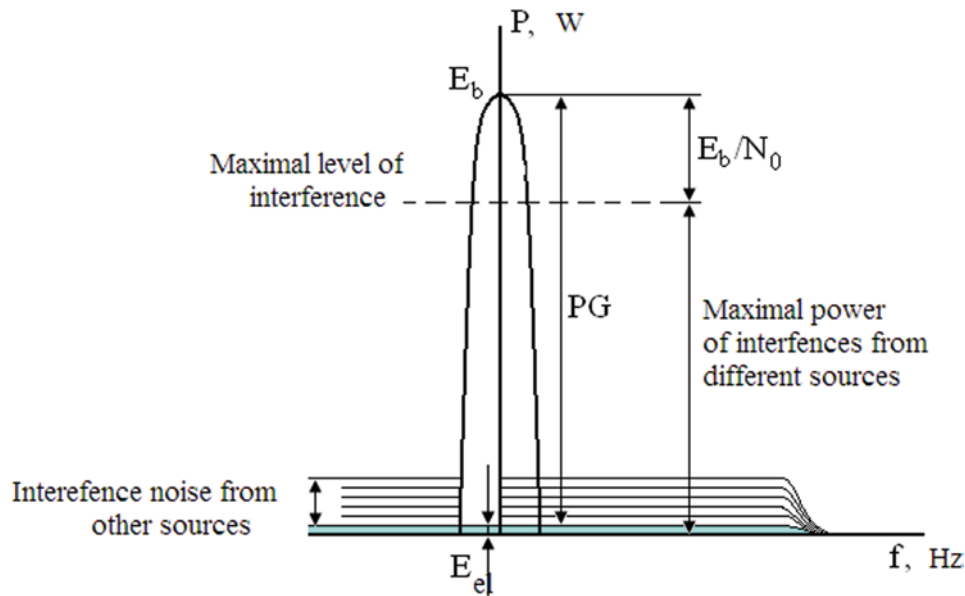


Fig. 3.41 Dependence of power versus frequency

will be more at low rate of the informational sequence and will decrease with its growth. For WCDMA systems, the chip rate is 3.84 Mchip/s, and at the speech stream rate of 12.2 kbit/s, the benefit in the gain (Fig. 3.41) is $PG = 25$ dB ($SF = 314.75$).

When the PRS of the useful signal acts in the receiver input, the power spectral density of the chip (elementary symbol) is E_{el} . All active AT of a separate cell create broadband interference at the input of all receivers, the level of which, like the useful signal, is close to the level of natural electromagnetic noise of the environment. The limit value of the interference power created in the receiver input by the sources of different nature (thermal noises, space noises, oscillators of electromagnetic oscillations, other sources) is defined by the required value of E_b/N_0 (the ratio of single bit energy to the noise spectral density). Main requirements to the equipment, in accordance with standards to the equipment for the mobile station of WCDMA systems, are presented in [9].

Maximal possible value of the spreading coefficient $SF = 314.75$ is not supported by the system, at least, in the UL direction. Possible values of spreading coefficients are 4, 8, 16, 32, 64, 128, and 256. The highest value of $SF = 256$ is used usually in the control channels, although the lesser values may be chosen.

The air interface of the WCDMA system has the nominal bandwidth of 5 MHz. At nominal offset of the carrier frequencies through 5 MHz, the grid of subcarrier frequencies in the radio access network UTRAN is 200 kHz, and in some frequency ranges, it is possible to use the subcarrier offset through 100 kHz on the band boundary (4.4 MHz, ..., 5 MHz). Such offset can be used to avoid the interference, especially, if the frequency band of the adjacent cell is located on another carrier.

3.3.2.2 The CDMA System with the IS-95 Protocol

The IS-95 [80] protocol system represents the equipment complex, in which the principle of frequency-coding signal division is used in radio channels. The operating frequency range in the direction from the subscriber station to the base station (the upward line) is located within the limits of 824÷849 MHz, while in the inverse direction (the downward line) within the limits of 869÷894 MHz. The duplex offset is 45 MHz. The carrier frequencies of transmitters and receivers are defined within the limits of mentioned frequency ranges with the step of 30 kHz. The allocated frequency range in Russia (State Standard RD 45.177–2001) and the step of carrier offset provides the compatibility of the digital CDMA2000 system and the analog AMPS system. The CDMA2000 system is a wideband system. It occupies the bandwidth of 1.23 MHz, in which 41 narrow-band channels can be occupied. Within the limits of the single-frequency channel, we can arrange 64 coding channels. One of the 64 mutually orthogonal Walsh functions serves as the address sign of the coding channel in the line from the base station (downward line), while in upward line, this function is fulfilled by quasi-orthogonal lengthy pseudo-random sequences.

Utilization of any carrier frequency for arrangement of the physical channel for any subscriber provides the effectiveness coefficient equaled to 1 for this cluster (the group of cells with different carriers). If several carriers are used in the cell (CDMA2000 3x), then they are placed with the step 1.23 MHz. At the ends of the operating range allocated to this CDMA network operator, the presence of the protective bands is supposed with the width of a half of the main band, i.e., 615 kHz.

The principle of interconnection between the base station and the mobile terminal is described in the IS-95 standard, which was initially intended for the closed network of military destination (USA). The IS-95 network can be realized in the frequency ranges less than 900 MHz with the channel width 1.25 MHz and with frequency duplexing with the frequency offset of 45 MHz and in the range below 2 GHz with duplex offset of 80 MHz. The downward channel from the base station contains 64 logical channels, which are formed using the Walsh function as the spreading sequence. Each from 64 sequences represents the line from the Adamar matrix of 64th order having the line orthogonality.

Formation of the radio signal. The process of informational stream formation in the downward channel begins from the conversion of the analog signal into the digital sequence (Fig. 3.42).

The input stream (the digitized signal, data) with the rate 1.2,...,9.6 kbit/s is exposed by the convolution coding with the coefficient $R = 1/2$ for elimination or detection of the block errors. The signal on the encoder output represents the result of

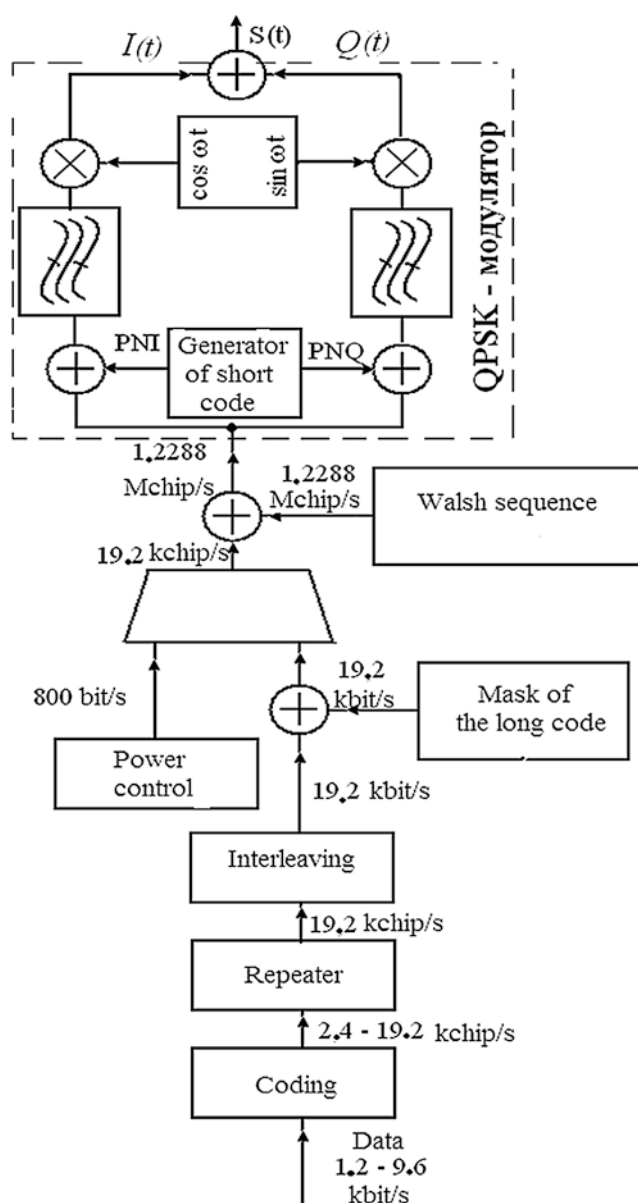


Fig. 3.42 The process of informational stream formation in the downward channel

congruence addition of two input bits and the bit containing in the memory cells. Depending on the communication conditions in the repeater, the transmission of the one data block can be repeated up to eight times providing always the output digital stream with the constant rate 19.2 kbit/s. This creates the equality of output transmission rates independently on the channel type (informational, paging). In the interleaving device that represented the matrix, in which informational bits are recorded in lines, and read in columns, one can perform the digital stream protection from the group errors.

Then the digital stream, including information and service symbols, is multiplied with 42-bit number, called the mask of the long code, which is actually the AT number. The data stream is then expanded using orthogonal Walsh function. Each bit of the original sequence is multiplied with a 64-bit Walsh sequence, forming 64 channels with a unique sequence and a very wide noise-like spectrum. This procedure allows the individuality provision of each subscriber sequence and thus provides confidence of transmitted information.

Then, each bit of the obtained 19.2 kbit/s data stream is represented by the integer period of the spreading Walsh sequence of the length of 64 symbols (the bit is the characteristic of informational sequence, the chip is the sequence element). Totally, we obtain the signal with the spread spectrum and the data stream rate $19,200 \cdot 64 = 1.2288$ Mchip/s. The 128 chips of the output sequence corresponds to each information bit of the initial stream, which corresponds to the benefit in SNR of the spread signal by $10\lg 128 = 21$ dB compared to the initial signal. Then, in each channel, the data stream is added modulo 2 with the pseudo-random sequences: in-phase **PNI** (Pseudo Noise Inphase) and quadrature **PNQ** (Pseudo Noise Quadrature). Obtained double-bit combination defines phases of the QPSK-modulator. From the output of the pass-band filters, in-phase and quadrature pulses modulate the in-phase $I(t)$ and quadrature $Q(t)$ components of the carrier oscillation of the chosen frequency. In-phase and quadrature components of the radio signal from each channel are added and form the group signal transmitted by the base station. Since Walsh sequences are mutually orthogonal, interference noises between channels of the single base station are practically absent. Transmission is performed by the packets with duration of 20 ms. The single base station can support up to 64 channels on the single transmission frequency, each of which is located through 30 kHz in the pass-band.

There are realizations of the structural diagram of the transmitter, when the spreading sequence (the Walsh sequence) is introduced not in the band of the modulating signal (Fig. 3.42), but in the IF section: after the PSK modulation.

The reception section of the transceiver of mobile communication. The processing of received signals occurs in the inversed order (Fig. 3.43).

The signal with the operating bandwidth limited by the band-pass filter passes to the input of LNA, which usually also has the BPF, which provides the image channel suppression. After radio signal conversion into the intermediate frequency band with the help of FC, the signal passes to the correlator, where it is multiplied by the etalon sequence, which is formed in the generator of pseudo-random signals. Digital correlators applied for extraction of its own signal calculate the correlation function between received and conferred sequences of the mobile station. The structure of the mobile station has usually several correlators for operation with several base stations, which is especially important at roaming and at the multipath reception. The receiver of the mobile station performs the synchronous reception due to the pilot channel, which is created by the base station. The power in the pilot channel exceeds the power in the traffic channel by 4–6 dB. This permits to the mobile station receiver to estimate the receiving signal power with the help of the RAKE receiver (correlation signal processing). Functioning of all base stations is synchronized by the system clock with the help of the global position system (GPS), which allows utilization the same spreading sequences with the individual time shift. The relative frequency instability of the base station

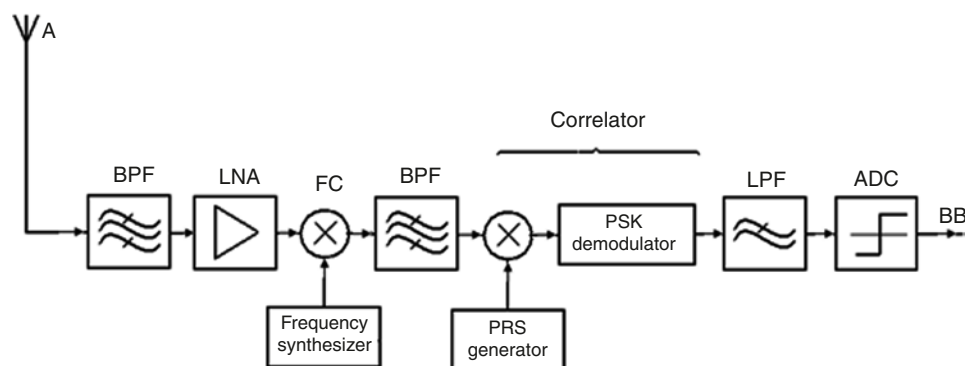


Fig. 3.43 The reception section of the transceiver of mobile communication

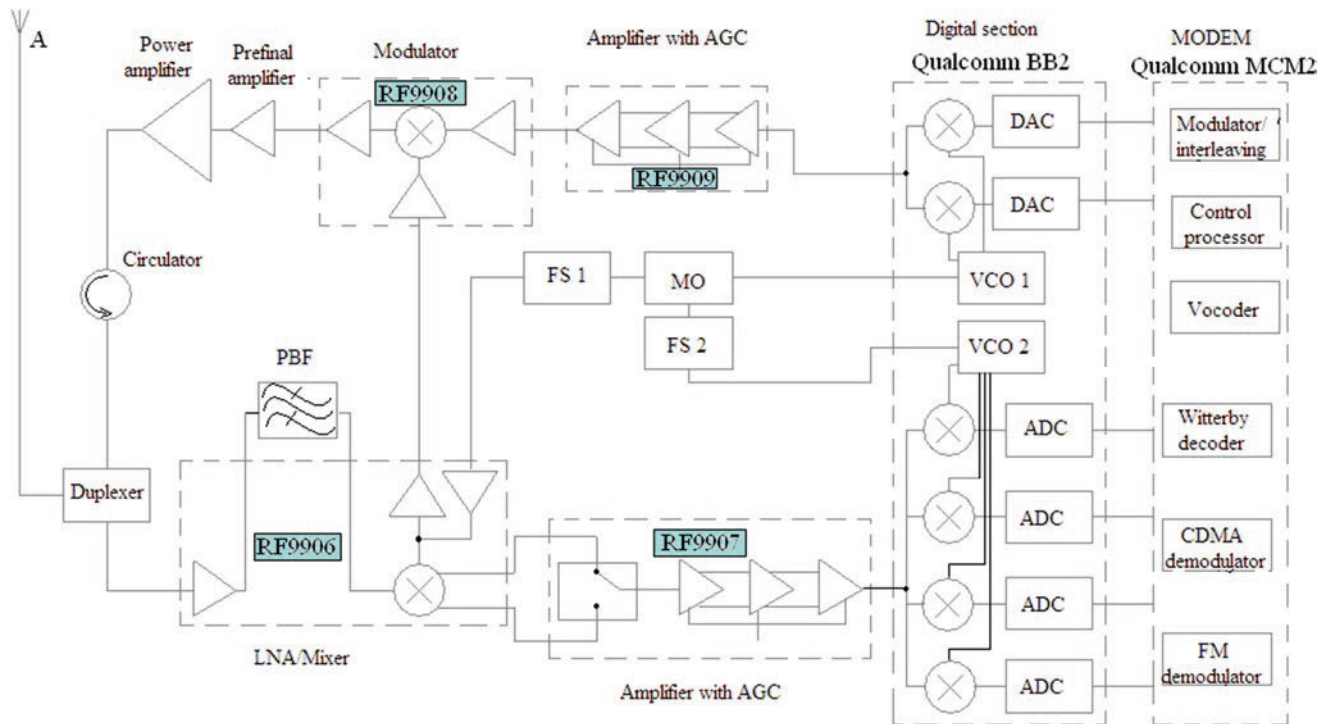


Fig. 3.44 The structural diagram of the transceiver operating in two modes: reception/transmission of analog signals with frequency modulation and digital signals with code channel division

and mobile station transmitters is $0.2 \cdot 10^{-7} \dots 1 \cdot 10^{-8}$. The cyclic shift of transmitted (received) digital sequences is settled with the step in 64 chips, which allows the roaming in 511 adjacent cells.

Within the cell, levels of signals received by the base station should be the same independently on the distance to the mobile station. For this, the transmitter power of the mobile station is adjusted in the range up to 80 dB (with the 1 dB-step) each 1.25 ms. Besides, the operation speed of the speech codec is not constant and can vary depending on the speech intensity from 8 to 1.2 kbits/s, which allows adjustment the network loading. To process the received sequence, as a rule, the Viterbi algorithm is used. It represents the in-series decoder, i.e., decision about data is made by the algorithm on the base of the whole data sequence in contrast to the symbol-by-symbol detection.

The structural diagram of the transceiver [98] operating in two modes, reception/transmission of analog signals with frequency modulation and digital signals with code channel division, are shown in Fig. 3.44.

Multifunctional microcircuits applied in the receiving and transmitting sections are realized with utilization of bipolar transistors with the hetero-junction (HBT).

The integral microcircuit RF9906 includes the low-noise amplifier and the mixer (LNA/Mixer), RF9908, the frequency up-converter (the modulator), while RF9907, RF9909 are amplifiers with the controlled gain. These ICs can be the component part of the second step of the cellular phone. The input part including the LNA and the double balanced mixer and two switching IF amplifiers with controlled gain is realized on microcircuits.

The RF9906 microcircuit (Fig. 3.44) includes the low-noise pre-amplifier, the passive double balanced mixer, two switching IF amplifiers with controlled gain, and the buffer amplifier, which allows connection of the master oscillator to two consumers. The bandwidth of LNA/Mixer at each FC output is 110 MHz by 1 dB-level. The resulting noise factor is 2,8 dB at the value of intermodulation distortions of the third order that reduces to the input IIP3 = -8 dBm for the nominal gain.

The amplifier RF9906 is covered by the series negative feedback to achieve the maximal input resistance охвачен without application of reactive matching elements. At LNA gain of 16 dB on the frequency of 800 MHz, decrease of the signal voltage level at its input due to the negative feedback is not less than 22 dB. The noise factor is less than 1.5 dB on frequencies below 1200 MHz at IIP3 = -3 dBm.

The mixer is implemented on the circuit of the double balanced converter (the universal analogous multiplier of signals) on the Gilbert cell.

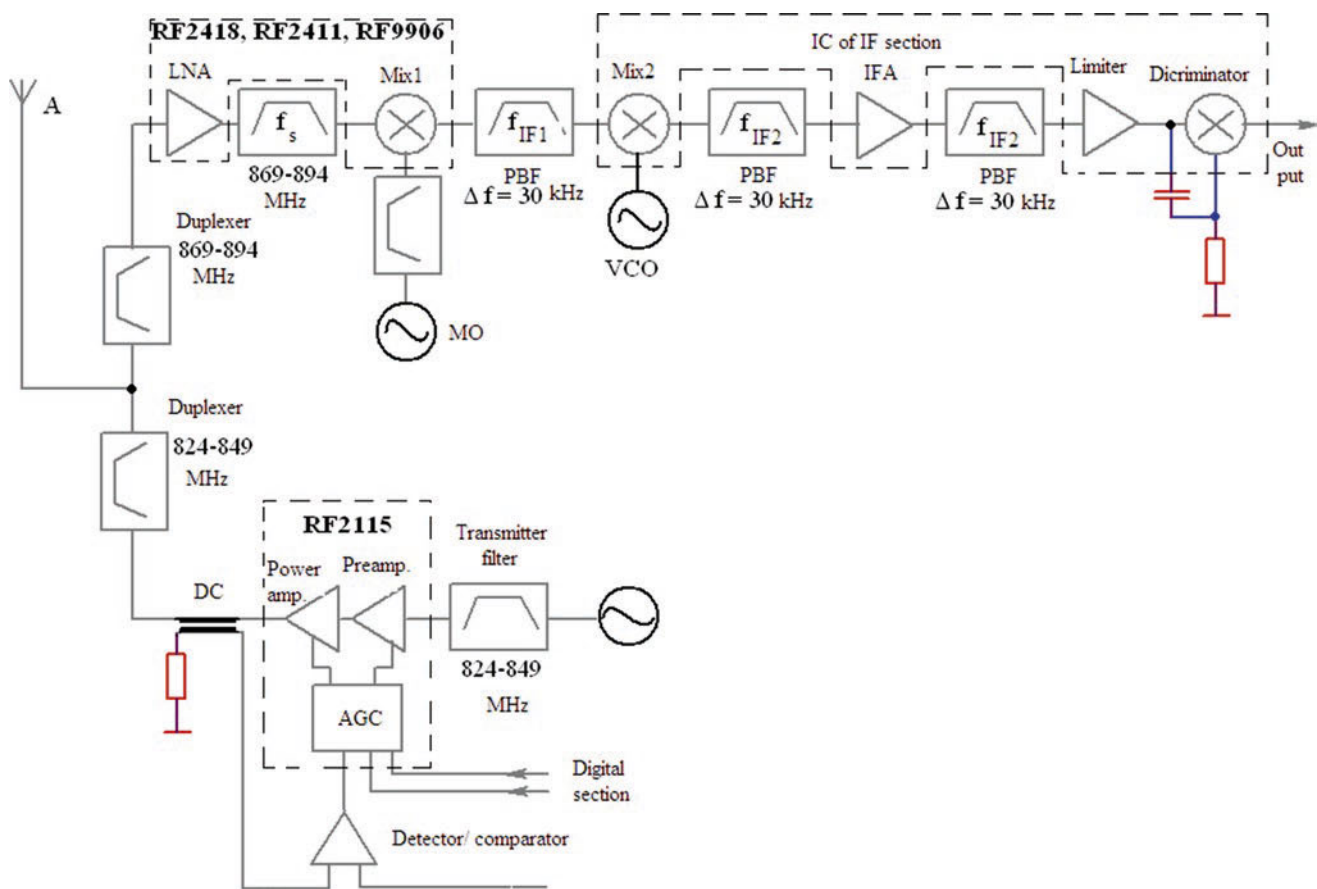


Fig. 3.45 The variant of implementation of the transceiver radio front-end at usage of the double-frequency conversion

The amplifier with AGC system in the receiver (Fig. 3.44) on IC RF9907 should have a possibility of gain controlling in the large range of input signal level but, at that, should have the low-noise factor. The dynamic range of gain control is 90 dB ($-45 \dots +45$ dB) with the analog gain control by variation of the supply voltage. The monotonous characteristic of gain control saves its linearity in the range of gain variation, which is more than 90 dB.

The variant of implementation of the transceiver radio front-end at usage of the double-frequency conversion [53] is presented in Fig. 3.45.

The duplexer factually represents two in-pair filters providing the frequency division of the band (the name from which) of receiving and transmitting frequencies. This permits to use the single antenna for operation in the two frequency ranges. The characteristic of the transmitter filter limits the interference value, which are created by the transmitter, in the input of LNA of the receiver. The filter of the transceiver receiving part protects LNA against the potential dangerous power levels at its input from the transmitter power amplifier and simplifies of the image receiving channel suppression on the first intermediate frequency.

The main LNA function is to provide the amplification enough for the noise factor exceeding in the following units, while its own contribution into the noise factor should be minimal where possible. Factually, the LNA noise factor defines the receiver sensitivity. Simultaneously with achievement of the maximal LNA gain, we try to perform the highest receiver linearity because the resulting value of IIP3, which is mainly defined by a mixer, will decrease by addition at its inputs of filters and amplifier. In other words, it is necessary to define what is the most important: the high gain of LNA, the low level of resulting IIP3, or the dynamic range. Comparison shows that the value of LNA IIP3 is significantly higher than IIP3 of the mixer without taking into account the resulting gain before it.

In a majority of receivers of the super-heterodyne type, the filter for image channel suppression is connected between LNA and the first mixer. It fulfills two close functions: providing of the image channel suppression on the first intermediate frequency in addition to the duplexer and avoiding the frequency band conversion near the image channel onto the first intermediate frequency and the SNR reduction.

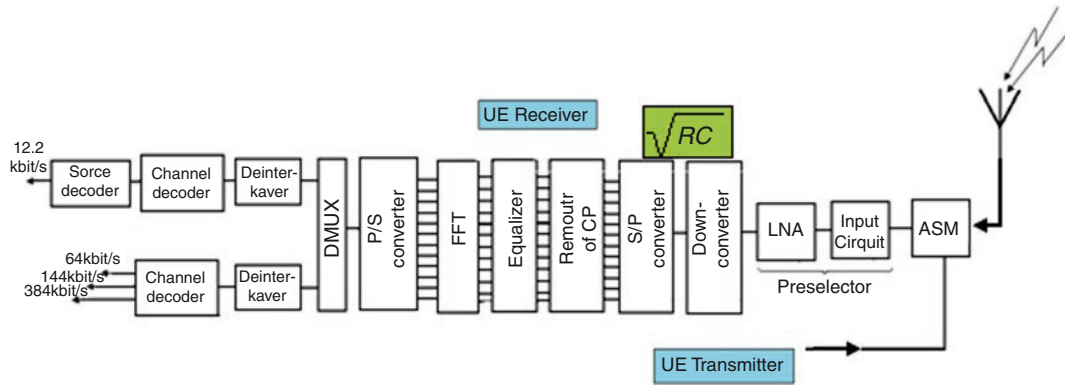


Fig. 3.46 The main fragments of the receiver structural diagram, which are used in the transmission system with the orthogonal frequency channel division

The first function of the Mix 1 is the conversion of the signal frequency f_s to the first intermediate frequency f_{IF1} . The conversion result is the input signal on the channel frequency, which is located with regard to the LO frequency by f_{IF1} . Since, in general, the mixer defines the receiver IIP3, its development is directed to obtaining of maximal IIP3 at the admissible noise factor and the specified value of LNA gain. The criterion for optimizing the LNA/MIX parameters for most multichannel receivers is to find a compromise between the LNA gain and the mixer noise figure. Thus, an increase in the LNA gain leads to decrease in noise ratio of the RF front-end (an increase the sensitivity of receiver), while simultaneously reducing the dynamic range of the LNA, i.e., leads to an increase in the risk of intermodulation distortion (reduction of intercept points IIP2, IIP3 of LNA and MIX).

A majority of receiver constructions also includes a filter between the master oscillator output and the mixer input. The purpose on connection of such a filter is to avoid the noise arrival on the first intermediate frequency due to input signal components differed from the useful signal and converting on the intermediate frequency.

The filter on the Mix1 output extracts the channel, which passes through the input circuit. In addition to extracting of the useful channel, this filter suppresses the out-of-band signals passed through the input circuit, which can lead to intermodulation distortions due to their high level caused by amplification in LNA. Since the interference signals are not created in the input circuit (duplexer), then the Mix 1 mainly defines the resulting IIP3. The additional function of the filter on the Mix 1 output is the image channel suppression on the second intermediate frequency.

After the pass-band filter extracting f_{IF1} , the second mixer (Mix 2) is usually used, which converts the useful signal into the second intermediate frequency f_{IF2} , and then the amplifier-limiter, the narrowband filters, and the frequency discriminator follow. All these functions (excluding filters) can be easily realized with the help of the single IC. The reason of signal conversion into the second intermediate frequency (usually 455 kHz) consists in the fact that in this case, the necessary filtering and the high gain (more than 100 dB) can be realized simpler. The discriminator output is connected to the digital to the base-band digital section, where the speech, control, and signaling signals and the wideband digital data are divided and processed independently.

The power amplifier (PA) of a transmitter provides the definite power level in an antenna with a possibility of its adjustment, at least, in the range of 20 dB. The key importance at power amplifier implementation is attached to its effectiveness since this defines requirements to the power source of the mobile station. Besides, the power amplifier should not create the perceptible interferences in the LNA input in the receiving bandwidth in order not to decrease the receiver sensitivity. Interferences created by a transducer define the value of minimal signal attenuation, which is inserted by a duplexer in the receiving section. In the similar manner, we can define requirements to the transmitter duplexer for the amplification mode of the harmonic signal with the help of the detector connected through the directional coupler (DC) in the power amplifier output. On the one hand, the power amplifier gain of the transmitter must be sufficient to provide the saturation mode of the active element of the output amplifier, but on the other hand, not to be very high in order not to create distortions (spurious signals). In many cases, such in-band spurious emissions cannot be completely excluded in an amplifier; therefore they must be minimized at network planning. Another key issue at transmitter designing is a requirement of PA operation with keeping of its parameters in the wide frequency range, when the reflection coefficient in the duplexer input changes rather significant.

The UMTS system of wideband access of the WCDMA standard. The structural diagrams of the receiver radio front-end for UMTS [9, 42] do not differ from applied in the CDMA2000 system and its modifications and may use the super-heterodyne construction principle (considered earlier) with the single and double-frequency conversion of the receiving signal.

3.3.3 *The Structural Diagram of the LTE System Receiver*

Before studying the architecture of the UE receiver of the LTE system, you should carefully read Sect. 2.5.3 of this book and understand the features of the formation of the baseband signal, OFDM technology, which ensures orthogonality in time and frequency of its components and OFDMA multiple access to the radio channel [24]. This will make it possible to understand the purpose of the nodes of the structural diagram of the receiver, complementary to the nodes of the transmitter.

The main fragments of the receiver structural diagram, which are used in the transmission system with the orthogonal frequency channel division, are shown in Fig. 3.46.

The signal generated by the antenna at the input of the ASM antenna switch, in the receiving mode, is fed to the input of a non-tunable preselector consisting of an input circuit and a low-noise amplifier. This structure provides the allocation of the operating frequency band of the system and suppression the image channel.

After down-conversion and allocation of the operator's bandwidth using a matched Nyquist filter at the center (intermediate) frequency, restoring the orthogonality (in frequency) of these responses, the signal arrives sequential/parallel signal transformation. After that, CP is removed from the group of 12 subcarriers that make up the OFDM symbol, which makes it possible to ensure mutual orthogonality of subcarriers over the entire symbol length (orthogonality in time). The CP, which is an effective way to deal with multipath at the receiver input, now becomes an interference in the algorithmic demodulation procedure (FFT).

Fulfillment of orthogonality condition for pulses of the modulating digital sequence of each subcarrier permits to present the transmission channel in the form the large number of independent channels with its own complex transfer function. Therefore, for correlation processing of transmitted information, we must equalize the phase shifts and values of gains, which is performed by an equalizer.

Remaining block of samples is processed with the help of the fast Fourier transform, and demodulation is performed (formation from the time sequence of the multi-frequency structure), which has the orthogonality.

In its output, the signal consisting of the sequence of responses, which have the mutual orthogonality, is formed. After parallel/sequential signal transformation, it passes to the input of the Fourier converter or to the group of parallel I/Q demodulators. The complicated circuitry structure performing the radio signal demodulation and the conversion into the pulse sequence, as in the transmitting section, can be replaced by the Fourier transform. The discrete Fourier transform allows formation of the pulse sequence by software means with the deciding device, when the conclusion is made about the reception of logic zero or unit by means of comparison of the receiving value with a threshold. This allows obtaining the bit sequence from the sequence of informational symbols, which then is exposed to deinterleaving and decoding.

The packet structure of the digital information representation begins with the preamble consisting of some number of adjusting OFDM symbols. A part of subcarriers is used as the pilot subcarriers, on which the known informational symbols are transmitted, which are used for corrections of the carrier and synchronization frequency.

When using the narrowband channels, for instance, in GSM standard, the channel bandwidth is 200 kHz, and it is impossible to increase instantaneously the transmission rate even at the presence of free channels. Another advantage of the wideband channel is a possibility for the subscriber to choose the channel parameters and the loading plot in accordance with its wishes.

3.3.4 *Reconfigurable System Receivers*

The reason for the close attention to reconfigurable systems is the growing need of users to expand access to high-speed mobile services, which is directly related to the bandwidth occupied by the operator. As indicated in Sect. 2.5.4, there are two ways to expand the operating frequency band: through the static allocation of additional potentially free spectra of other operators, performed by the scheduler, and the dynamic allocation of other operators' frequency bands in real time.

3.3.4.1 Software-Defined Radio (SDR)

Block diagram of a super-heterodyne type with digital intermediate frequency (Fig. 2.35) or direct conversion (Fig. 3.15) is the main types of architecture used in software-tunable radio implementation. The reason for this choice is the need to simplify the structure of the RF front-end.

The first step is to use a non-tunable preselector that allows you to receive signals from any operator in your system or systems of other standards. In the latter case, this can be realized by complicating the structure of the input radio frequency module (IFEM) [64] by including additional filters that allocate the frequency band of base stations of other systems (operators), adding a part of the aggregated spectrum. An alternative option is to create a parallel structure of the radio path (RF front-end), which, in addition to its own frequency band, receives signals from a system of a different standard, which complicates the AT design and increases its cost.

The shortened receiving part of the AT simplifies the structure of the radio path and reduces the complexity of filtering systems that are received side channels.

Since the position of the aggregated frequency bands in the BS and AT is known in advance, then, relying on modern ideas in choosing the architecture of receivers and advances in integrated technology makes the structure of the receiver obvious (Fig. 3.47).

This structure implements the principle of direct conversion of an analog input signal at frequency f_s to the baseband BB. The mobile node (MN) in the LTE system estimates the power arriving at the receiver input from three eNB (evolved node B) base stations. An MN selects one active eNB according to the level of the input signal and the minimum probability of error in the received sequence for constant interaction, while continuing to evaluate the parameters of other eNBs. Besides, MN receiver with an SDR structure in a dynamic mode evaluates the efficiency of using eNB of spectra allocated to it. When detecting in the spectra passive eNB of free parts of the spectrum and intense traffic in the active cell, MN forms a request for active eNB to provide it with a part of the free spectrum of neighboring cell. An E-UTRAN sends eNB to the neighboring cell to ensure joint operation of MN with two eNBs in the main and added parts of spectrum. The low- and high-frequency filters in the FEM structure select the operating frequency bands of the active BS and fed to the LNA input. A low-noise amplifier with resistor or selective load [24] provides a level of power at the Shaper of Quadratures input, sufficient to split the signal with minimal distortion into two orthogonal components: in-phase I and quadrature Q. Its inclusion reduces the risk of leakage of conversion products into the antenna. Reducing the speed of orthogonal information sequences at its output reduces the requirements for the ADC bandwidth in the DSP structure. After amplification in high-frequency amplifiers with automatic gain control (AGC), the analog components of the input signal are fed to separate DSP inputs.

DSP is a multifunctional LSI that, in addition to the sampling procedure, performs demodulation by software or circuitry. Pulse sequences formed in the demodulator in the DSP structure are converted into digital ones. Then, using a convolutional code decoding technique, a solution, such as a Viterbi algorithm, about the received code sequence is found based on the minimum code distance.

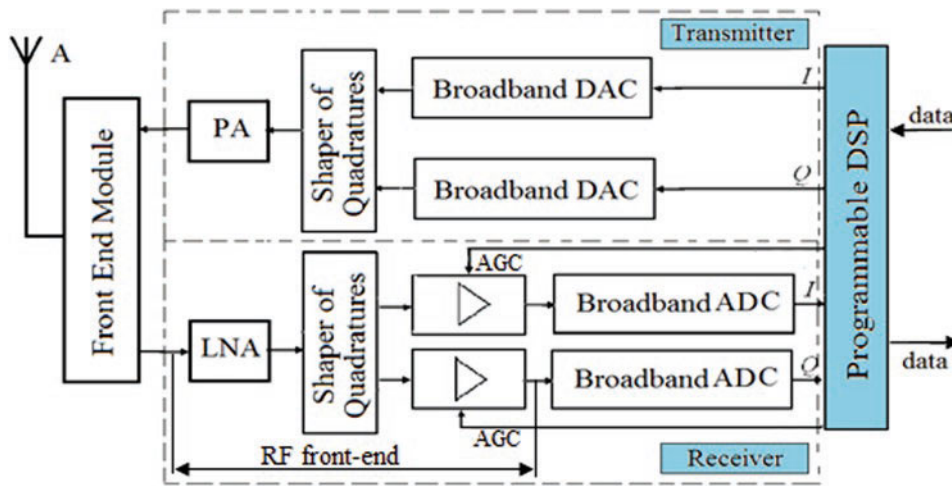


Fig. 3.47 Block diagram of a software-defined radio (SDR)

The described variant of the AT structural diagram includes one common antenna used for receiving and transmitting radio signals. From the point of view of the processes described, connecting several antennas using, for example, MIMO technology, will only change the filtering blocks in the high-frequency part of the receiver [22]. The number of antennas for the aggregated spectrum at the receiver input will be determined by the product of the number of active BSs and the number of antenna groups receiving orthogonal information streams in each allocated band.

The type of modulation of the signal received and transmitted by the AT is determined by the terrestrial radio access network.

3.3.4.2 Cognitive Radio (CR)

The concept of cognitive radio [40] is used in the technology of dynamic spectrum allocation based on IP protocols for mobile equipment DRiVE (Dynamic Radio for IP Services in Vehicular Environments). The architecture of this project combines various access technologies, increasing the efficiency of using the radio spectrum by choosing the optimal technology for providing a specific type of service. It includes a number of necessary functions and procedures, among which the main one is the function of traffic control for the duration of the entire connection session. Optimization of its indicators is based on knowledge of the parameters of the user terminal, the required parameters of the quality of service, as well as its priority and properties of the selected network. The procedure for the user to select a radio access network is performed every time the AT is connected or the state of the radio channel changes.

The concept of dynamic spectrum allocation (DSA) is based on continuous monitoring and searching for an area of potentially available radio frequency spectrum, assessing its parameters and the possibility of using a potentially free part of the spectrum at the current time.

There are two ways to reallocate the spectrum: the distribution of adjacent frequency bands and the distribution of spectrum fragments from different bands and frequency ranges. The algorithm for dynamic redistribution of the frequency spectrum of interacting radio access systems is based on knowledge of the statistics of traffic changes in them. An adaptive algorithm with long-term memory and, at the same time, quickly responding to changes in traffic was selected as the main one that controls the process of dynamic redistribution of spectra. Such properties are possessed by Kalman filters that perform fast prediction, as well as artificial neural networks and genetic algorithms.

Analysis of traffic and parameters of networks interacting with ATs is used in 3G and 4G systems for intranet and inter-network roaming and to assess the power level acting at the input of the AT receiver from neighboring BS. With a fixed allocation of the radio frequency spectrum FSA (Fixed Spectrum Allocation), this allows you to select the “main” BS, which creates at the AT input the maximum signal level compared to BS from neighboring cells. The DSA dynamic frequency selection algorithm requires the use of the DFS (Dynamic Frequency Selection) Protocol to detect devices using the same spectrum (channel) and bring the interference level to acceptable values determined by the TPC (Transmission Power Protocol) control the power of transmitters in the network.

It is obvious that the transition to a dynamic spectrum distribution complicates the algorithm for interaction of terminal devices during a data transfer session, which presupposes a preliminary scan of all potentially suitable areas of the spectrum. At the same time, in cognitive radio, the requirements for the amount of RAM and computing power of devices operating in real time are increasing.

The DFS standard defines the power level of total interference that acts on the input of the terminal device and allows you to calculate the degree of their impact when switching to another frequency area of the license range.

A simpler form of cognitive radio implementation that observes and analyzes the results of spectrum monitoring refers to a particular case of cognitive radio.

CR, depending on the status of the allocated band, can be used in the unlicensed UBCR (Unlicensed Band Cognitive Radio) or in the licensed LBCR (Licensed Band Cognitive Radio) frequency band.

General block diagram of transceiver for cognitive radio (Fig. 3.48) implemented using the structure of SDR (software-defined radio).

Figure 3.48 shows a block diagram of the transceiver together with additional blocks not used in the standard AT circuit.

The receiving path is connected via a dynamic frequency selection unit and an input front-end module (iFEM) to a broadband antenna.

The block for dynamic frequency selection is connected to the antenna in the receive mode via iFEM. Structurally, antenna can be performed as a unit of antennas optimized for operation in a certain part of the operating frequency range or as a group of antennas united by MIMO technology.

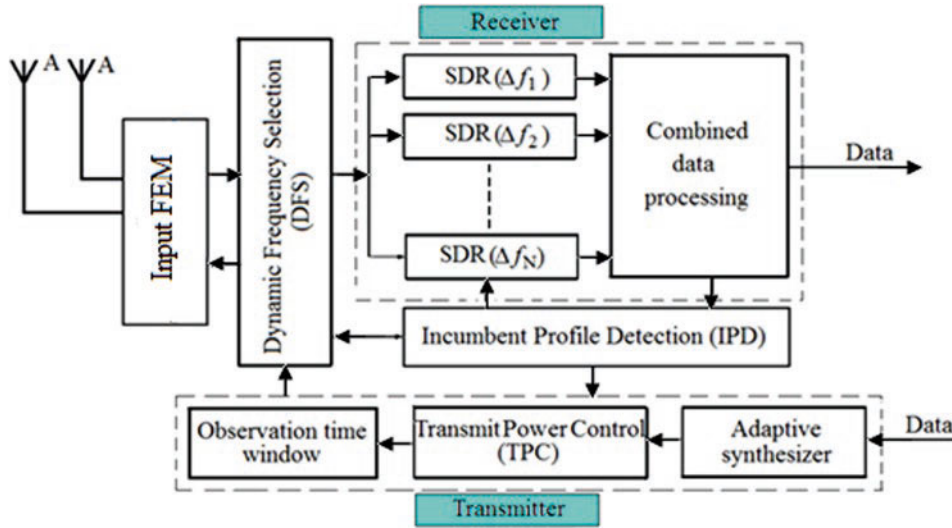


Fig. 3.48 General block diagram of a complete cognitive radio

This makes it possible to represent the information flow from the BS in the form of parallel orthogonal messages containing a part of the group signal [85]. Some complication of the processing algorithm of signals is compensated by a decrease in the level of intermodulation distortions in the radio channel and increase capacity of it.

The input FEM unit, in addition to separating the directions of reception and transmission (frequency or time duplexing), when it is an antenna switch, can divide the received frequency band into several subbands, the boundaries of which are assigned by the scheduler when building the network. The AT overview of the spectrum status in real time provided by the active BS, the traffic intensity, and the efficiency of using its operating bandwidth, as well as the presence of the spectra of other BSs interacting with the user, is performed in the incumbent profile detection (**IPD**) block. Information about the efficiency of using the path in the frequency bands ($\Delta f_1, \dots, \Delta f_N$) allocated to each SDR of the receiving unit is subjected to combined data processing, and it is sent to the IPD. The detection in the dynamic mode of free parts of the spectrum not only of neighboring cells but even of the others systems complicates the procedure for identifying the location of UE. The problem is simplified if, in some part of the frequency range belonging to another cell, the BS is already exchanging data and signaling with the UE, which means that the coordinates of BS are known. Then the use of the free part of the spectrum at the current time does not require the use of additional technologies to determine the geographic coordinates of the UE of neighboring cells or systems.

Information about the current profile of cognitive radio detection (aggregated operating frequency bands, power levels of BS transmitters participating in the session, and its distribution over the range) are stored in a special centralized data storage unit and, upon obtaining permission for spectrum sharing, in the AT transmitter unit.

Information about the distribution of frequency channels occupying the licensed/unlicensed range, the nature of the spectrum, and the change in the spectral power density in a specific geographic area is sent to the dynamic frequency selection (**DFS**) block.

The AT receives a command from the core network of the system to allocate a part of the free spectrum of the BS of its own or adjacent network for transmitting and receiving information at a specified time interval. The described method of providing a certain frequency band for temporary use is widely used in WLAN networks as a method of hopping the operating frequency FH (Frequency Hoping) according to some pseudo-random algorithm.

The functioning of a receiver with a structural diagram (Fig. 3.48) is provided using several technologies available only to cognitive radio. These are spectrum sensing (Spectrum Sensing), access to a database with geo-location (Geo Location), and transmission of signals of special beacons (Beacon Transmitter) [37].

The purpose of *sounding* is to monitor spectrum use and detect protected radio equipment in a potentially available channel. The result of this procedure is to obtain information about the occupation of the spectrum (frequency band, channel bandwidth, etc.). One of the tasks of this measurement is the detection of “white spots” (White Space). This is a portion of the spectrum available for use in a specified period of time in a given geographic area, excluding interference with the operation of radio facilities with a higher level determined by the white space devices (**WSD**) list. The interaction of the AT and

the network is provided using the cognitive pilot channel (**CPC**), which provides the device with information about the available spectrum and the conditions for its use.

Using the *geo-location* function assumes that each cognitive device determines the coordinates of its location (GPS, GLONASS, etc.) and transmits it to the database, while identifying which radio frequencies can be used in a given service area.

Logical channel beacon transmitters are designed to generate information about the presence of free channels or channels protected by licenses, which allows the IPD to make a decision on the transition of the CR to the active mode or continue monitoring the occupancy of the spectrum of devices interacting with AT.

There are four ways to provide spectrum for cognitive radio systems. The option chosen depends on the development of technology and the regulatory framework and the stage of implementation of the system and is divided into: dedicated spectrum, general spectrum, reused dedicated spectrum, and also specially allocated for cognitive systems. For more information about spectrum planning for cognitive systems, see [37].

3.3.5 User Terminals for 5G Wireless Access Systems

3.3.5.1 Technological Features of Multiplexing in the 5G System

The description of the architecture of the physical network of the 5G standard (Fig. 2.38) reflects the variety of information sources, actuators, its multi-level, and variety of guiding systems and interfaces. If at the network, transport level, information is exchanged through logical channels of the OSI model (Fig. 1.10), then at lower levels, you have to deal with real signals to be transmitted over the radio channel; you must take into account their spectral and temporal characteristics, as well as the properties of the radio channel. The existing 2G÷4G networks are optimized for the interaction of people H2H (Human-to-Human), the created M2M (Machine-to-Machine) networks are based on the paradigm of building functionally changeable for a specific consumer with limited modular expansion. This has led to the expansion of the functionality of the upper layers of the OSI open network reference model while maintaining the functionality of the lower layers.

The main difference between the receiver (transmitter) of the 5G network from other global WWAN networks is the inclusion of WPAN, WMAN, and WLAN networks in its structure, which significantly affects the technologies and interfaces used in it. For example, the inclusion of IoT networks provides machine-to-machine communications M2M (MTC, Machine-Type Communications) via the Internet. Despite the heavy use of sensor devices, now included in the 5G network, it will now require a pairing device and specific software. Unlike smartphones, the main device in 5G networks, sensor devices must have low power consumption and regular access to the Internet, regardless of human activity, weather, location, etc.

The transition of an M2M network device at a random moment in time to a mobile network from a state called *fast dormancy* will cause a jump in the allocation of mobile network resources. The transition of the M2M network device at a random moment in time to the active state, as well as the exit from it, will cause a jump in the allocation of mobile network resources. With a relatively low data transfer rate of tens of kbit/s, this does not require a noticeable expansion of the radio channel bandwidth, but when the number of sensors reaches hundreds of thousands or even millions, then access in random time to and from the network, they will overload the system control and signaling of network and even blocking it. The 5G network operator must be ready for a jump in the data flow rate, an increase in the occupied bandwidth, and the connection of corrective links. At the same time, the constancy of technical characteristics for users of heterogeneous networks should be ensured.

Cellular networks currently in operation generate traffic with approximately 50% streaming structure and a data transfer rate of the order of 100 Mbit/s, and in a 5G network, it will be 10 Gbit/s or more. Maintaining such a high network capacity for organizing low-speed messages between M2M devices via a 5G mobile network is impractical due to the large amount of overhead information (Fig. 2.15) and requires a change in the technology of their interaction.

At the same time, it is necessary to take into account the requirements of low power consumption by terminal devices and data transfer rate, the possibility of regulating the transmitter power, providing low-mobile services: connecting surveillance cameras, permissible delay time, and adjustable transmission interval.

The impetus for building 5G networks was the need for users to expand the services provided by radio technical networks and, if we are guided by the trends in the development of reconfigurable networks, then the path lies through the improvement of the LTE system. The development of the LTE-A network is the use of reconfigurable systems, including receivers with *aggregated bandwidth*: SDR and CR. The structural diagram of the receiver of the LTE system (Fig. 3.46) belongs to

the superheterodyne type, in which, convert the received signal into the modulation frequency domain, it is required to multiply each subscriber of signal with the local oscillator voltage. With a large number of such local oscillators, it is a difficult task to perform it using circuitry methods. The solution of this problem by programs using the fast Fourier transform turns out to be the best option, even with a large number of subscribers ($N = 1201$ subcarriers, $BW = 20$ MHz), but under the condition of strict orthogonality of all modulated subcarriers in time and frequency.

This nominal bandwidth is always wider than the effective bandwidth, determined by the number of radio units allocated to the respective operator. Unused subcarriers are placed at the edges of the working (effective) band, forming a *guard band* that separates the bands of working or aggregated frequencies. For example, when forming a 20 MHz radio channel bandwidth, the number of FFT points is 2048 subcarriers ($2^{11} = 2048$), and the effective one includes 1201 subcarriers located at 15 kHz intervals ($1201 \times 15 \times 10^3 = 18$ MHz). Thus, in the aggregation of two systems with a bandwidth of 20 MHz (Fig. 3.49), the allocated bandwidth will be 40 MHz, and the effective bandwidth will be only 36 MHz. Taking into account two 1 MHz guard bands and a guard gap for separating the aggregated spectra of 0.3 MHz, the entire aggregated band will be 38.3 MHz.

The second condition for achieving *orthogonality in the time domain* between the symbols of the information sequence, in addition to choosing the symbol duration, is the introduction of a cyclic prefix, the length of which is 7–10% of the length of the OFDM symbol.

The OFDM technology used in fourth-generation networks is based on the requirement of orthogonality of signals in frequency and time and, as a result, synchronous signal processing in each terminal device. In the downward direction for MTS traffic in physical channels, it is possible to use CoMP (Coordinated Multipoint) technology. This provides uniform coverage with high capacity for networks using the orthogonality and synchronization property of one of the network cells' interface. However, this requirement will be violated due to overlapping coverage areas of small cells and blocking of the network caused by the huge flow of MTS device control signals.

The technology of multiplexing information sequences orthogonal in the frequency and time domains turns out to be ineffective in the construction of 5G networks. This is caused by the overlap of service areas in short-range networks (pico-, femto-networks), which excludes the possibility of roaming due to the large number of users and a sharp increase in the amount of service information. The maintenance of M2M networks with a sporadic character of connections to the channel in the 5G structure makes it impossible for the same reason to ensure synchronization between users. Together with the requirements to increase the efficiency of radio resource use and reduce signal time delays and the heterogeneity of the connected networks, it implies the creation of a new radio access technology based on non-orthogonal signals.

A solution to this problem is possible by reducing the requirements for synchronization when transmitting MTS messages, using “rough” synchronization or by excluding the standard mode of organizing a communication session in cellular networks through the procedures for registering a subscriber, entering information about the characteristics of a subscriber in the home register, searching for a recipient, etc.

The described limitations of OFDM technology in the construction of 5G first of all require the abandonment of strict orthogonality at all stages of signal processing. The OFDM technology is based on the idea of simplifying the process of

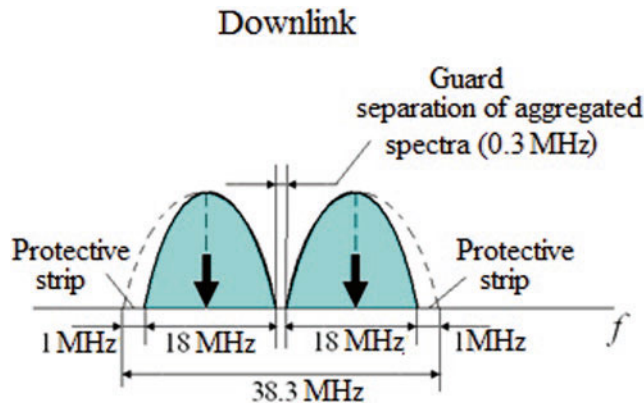


Fig. 3.49 An example of the aggregation of spectra of two frequency-neighboring operators with the bandwidth allocated to each of the 20 MHz

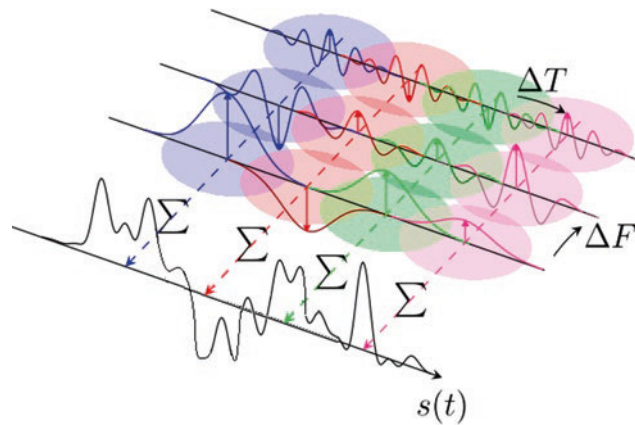


Fig. 3.50 Signal representation by the sum of scaled time-frequency offsets of the prototype window

forming a group signal with a large number of carriers, based on replacing the circuit implementation of the most complex node of the receiver's radio path with an algorithmic Fourier transform procedure. It allows you to describe the signal $s(t)$ at some point in time by a group of mutually orthogonal harmonic spectral components in the frequency domain, known as the Fourier transform. To describe the signal at another moment in time, it is required to perform a new Fourier transform, and to obtain time samples on the ensemble, multiple description is required, which, upon subsequent multiplexing, creates a huge redundancy of such a time-frequency representation of the signal $s(t)$. Gabor [79] proposed to describe the signal $s(t)$ by a linear combination of Gaussian functions shifted in time and frequency along a uniform grid (Fig. 3.50).

The Gaussian function is well localized in time and frequency, which allows the most expansion process through the product of the coefficients of the Gaussian functions. The accuracy of the description $s(t)$ depends on the distance between the grid nodes, which is determined by the product of ΔF and ΔT . If it is less than one, then such a description characterizes a specific signal, and if it is greater than one, then such a description is incorrect. The time-frequency combination of time-shifted Gabor windows is called expansion, and the short-time Fourier transform STFT is called the Gabor transform. The extension and the Gabor transformation window are dual, i.e., the coefficients used in the expansion of a particular signal in a given window are described by the Gabor transform in the dual window. If the window and its transformation are the same, then the window is called orthogonal and the Gabor transform is reduced to a Fourier series expansion. OFDM is a special case of Gabor expansion using a rectangular windowing function of length T_s and an offset of $1/T_s$ and is a discrete Fourier transform (DFT). The architecture of a 5G network should provide for the transmission of various types of signals, which, given the design features and methods of network planning, can lead to the emergence of new multiplexing and multiple access schemes:

- (a) Dynamic resource scheduling with channel adaptation to achieve high efficiency and orthogonality when transmitting standard bit pipe traffic.
- (b) Semi-static/continuous resource scheduling, regulated at the MAC level without adaptation to individual parts of the spectrum. At the same time, special attention is paid to increasing the efficiency of the multi-user mode and somewhat weakening the requirements for synchronization and orthogonality. But, at the same time, requirements for suppression of interference from neighboring cells are increasing.
- (c) "One shot transmission" with a combination of technologies of random access and sparse signal processing ("sparse") and the choice of its form for efficient data transmission over a physical random access channel (PRACH). Such messages can vary significantly in the amount of delay. Low latency can be achieved using an open feedback loop and lower timing requirements for synchronization transmitted data.

Access to the system via a physical broadband broadcast channel involves using it for rough synchronization of the user. This is achieved if part of the data and the control signal overlap in time, providing "one shot signal detection" and its alignment. Access to the system via a physical broadband broadcast channel involves using it for rough synchronization of the user. The pilot signal can be used as a reference signal. If the influence of the pilot symbol on the data is too great, it may lead to an erroneous "one shot transmission" reception decision. To avoid this, various rarefaction technologies are used: user activity, channel impulse response, spatial correlation, sparse topology, etc.

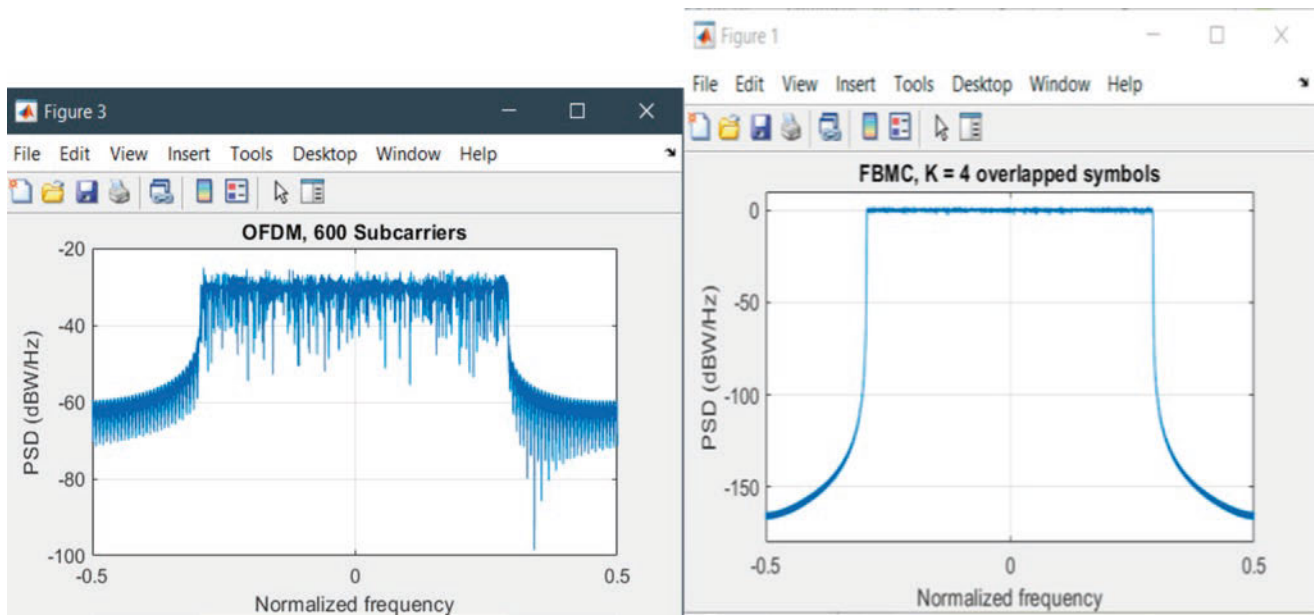


Fig. 3.51 OFDM spectrum (50 resource blocks, 12 subcarriers each) versus FBMC (600 subcarriers). At Signal-to-Noise Ratio (SNR) = 15 dB

The variety of signals used in the 5G system, which can have random access to the channel and strict orthogonality and synchronism; be very fast (several tens of Gbit/s) and very slow, requiring very low signal delays during transmission and be indifferent to their magnitude; work in a wide frequency range; etc., demanded the use of various methods of multiplexing and channel separation.

Widely used in the fourth generation systems, OFDM technology was developed in 5G as a method of **F-OFDM** (Filtered Orthogonal Division Multiplexing). This method is based on the requirement to preserve orthogonality between subcarriers grouped in a certain frequency band, united by the nature of the service. This allows the filtering of the generated subband, which keeps the requirement for the subcarrier group high while reducing the requirement for subband division in the frequency domain. This makes it possible to reduce the length of the guard intervals, which will increase the spectral efficiency of the system. Examples of energy spectra obtained by software [62] for identical conditions are shown in Fig. 3.51.

Another option for the development of OFDM technology is **FBMC** (Filter Bank Multi Carrier), when each subcarrier is filtered separately. FBMC is a multi-frequency signal shaping (modulation) technique supplemented by comb filtering each subcarrier. This technology combines hybrid frequency modulation and quadrature amplitude-shift keying: FQAM (Hybrid Modulation of FSK and QAM). The result is a reduction in the level of out-of-band emissions and an increase in immunity to interference between subcarriers. This made it possible to increase the speed of signal processing in MTS networks when using asynchronous FFT in a separate channel without providing preliminary time synchronization and correction of the subcarrier frequency.

To synchronize in time and in frequency at the same time, it is necessary to significantly complicate the receiver, but this can be avoided by performing fragmentary decoding of several sequences. Asynchronous FFT of size $K * N$ is processed in blocks of $N/2$ samples in KN points of the Fourier transform, for k samples: $k = (n + m \cdot N/2)$, where $n = 0, 1, 2, \dots, KN-1$. This sequence of KN points is stored in a memory block. The beginning of the packet will be found by the output of the FFT block based on the preliminary information obtained from the preamble. The frequency offset is first estimated using the subcarrier preamble information by calculating the phase offset product between two consecutive FBMC symbols of two consecutive pilot subcarriers. Carrier frequency offset compensation is performed using the feed forward approach technology. The channel characteristics, which are considered constant over the packet length, are calculated by interpolation over all active subcarriers and equalized in the equalizer before being fed to the FBMC prototype filter.

Signal demodulation using a log likelihood ratio algorithm and subsequent decoding provides signal recovery.

When building a 5G system network, special attention has to be paid to the interaction with the channel of MTS devices, when several users with weak mutual synchronization act in the uplink direction. This is the reason for the occurrence of inter-user interference (IUI) caused by time and frequency misalignments. For the transmission of such data, it turns out to be preferable to use **GFDM** (Generalized Frequency Division Multiplexing) modulation. Comparison of the energy spectra of signals with OFDM and GFDM modulation (Fig. 3.52) shows that GFDM has a lower level of **OOB** (of-of-band) side

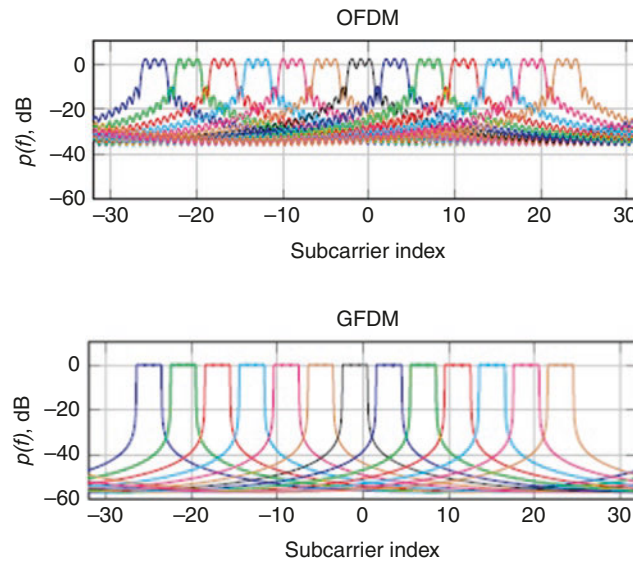


Fig. 3.52 Arrangement of subchannels with GFD and OFDM modulation

emissions. MTS devices can select random access subchannels as sporadic short bursts. Nodes (BS) use the CSMA access technology (carrier sense multiple access), when data is transmitted only in a given subchannel, if it is free, i.e., always on different subchannels.

With frequency shift, in the case of OFDM, the side lobes fall into the pass-band of the filter of the corresponding channel, leading to an increase in interference, even if it occurs within the guard gap (in frequency) due to the low attenuation of the OOB.

When using GFD modulation, the frequency offset of the adjacent subchannel within the guard interval, the direct conversion (zero frequency **ZF**) receiver has orthogonality and the MSE distortion remains the same. GFD technology has the ability to change the parameters to achieve low latency of the data block, and, therefore, to set low delays, for example, for tactile internet. This is achieved by using short frames and by reducing the number of requests for message repetition. The inclusion of filters to suppress concentrated interference in the radio path of the transceiver, as well as the use of precoding by Walsh functions, which distribute information over all available subcarriers, will significantly increase the noise immunity of transmitted messages.

The upstream CoMP (Coordinated Multipoint) data stream from a large number of sources from different cells creates a different delay at the node (BS), which requires its estimation at the receiving side. Frequency offsets generated mainly by multiple MTC devices can cause severe interference due to intersymbol distortion in multi-frequency signals that have low OOB.

3.3.5.2 5G System Transceiver Architecture with UFM Technology

The most promising technology is **UFMC** (Universal Filtered Multicarrier) – frequency shift keying with universally filtered carriers. It is also based on the use of the Fourier transform, but unlike the OFDM technology, where a block of channel symbols is subjected to an inverse Fourier transform at N points, forming a group of OFDM symbols s_i ($i \in 1, 2, \dots, N$) in the entire operating frequency band. In UFM, the total occupied bandwidth of the character sequence is divided into B subbands (Fig. 3.53).

Each of them includes a group of orthogonal multilevel n QAM sequences ($k \in 1, 2, \dots, B$) acting at the input of a separate Fourier transform. The spectra of the generated subcarrier subbands are determined by the subcarrier spacing (Fig. 3.54) and are always less than the subband width, calculated through the number of inverse Fourier transform points on the transmitting side. The redundant Fourier transform subcarriers are not modulated and placed at the boundaries of each of the subbands, forming guard gaps.

UFMC is a generalization of OFDM and FBMC technologies. If OFDM technology filters a dedicated frequency band, and FBMC filters each subcarrier, then when using UFM, groups of subbands contain groups of subcarriers. The total number of subcarriers N is divided into subbands, each of which has a fixed number of subcarriers. Each of the subcarriers from the subband is filtered with a filter of length L , and the responses of each of them are summed.

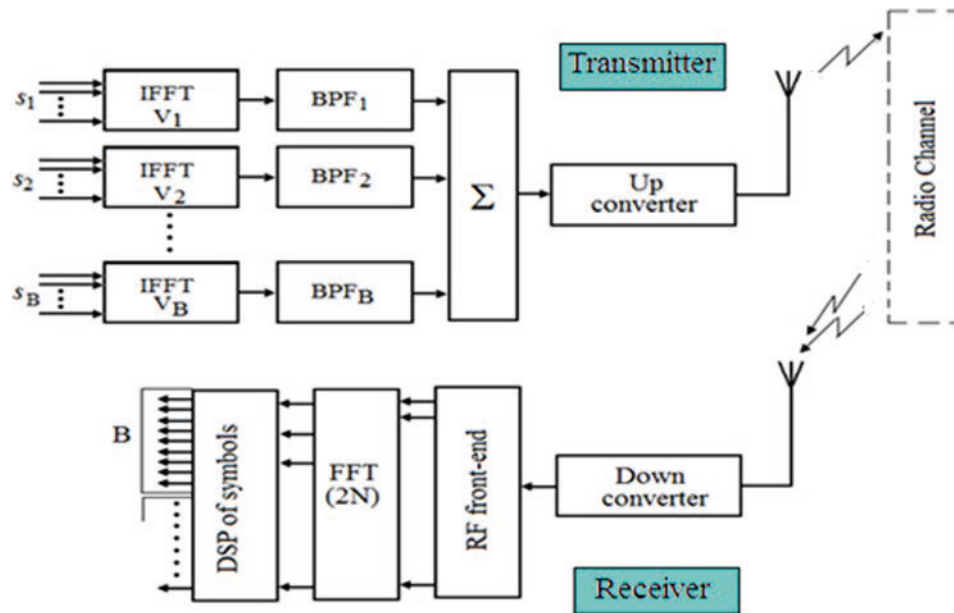


Fig. 3.53 Block diagram of a transceiver with UPMC technology

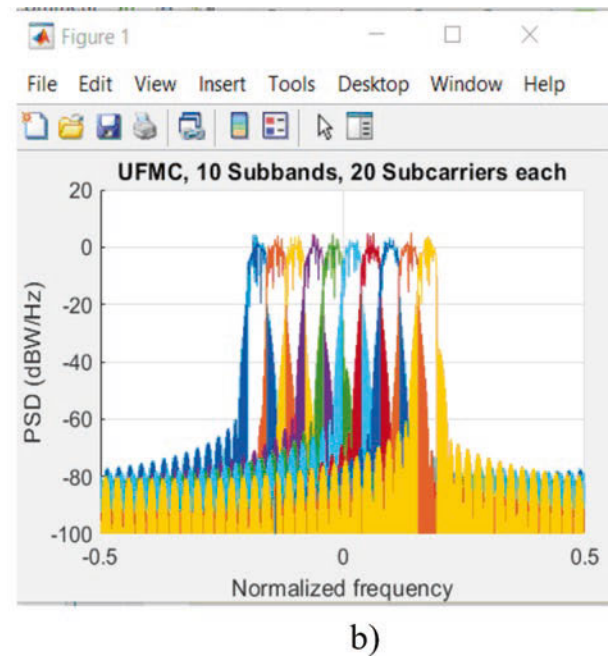
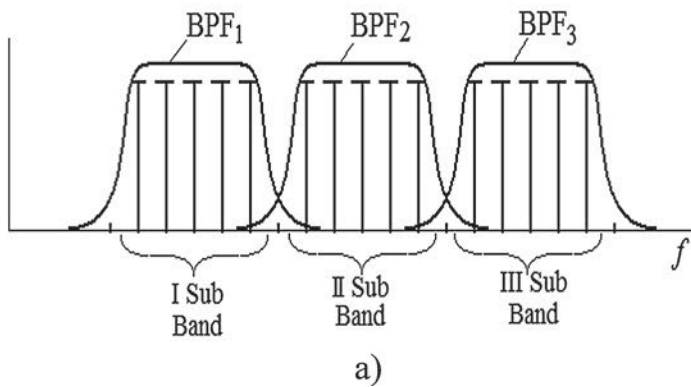


Fig. 3.54 Distribution of spectra of information streams a) at the output of the transmitter adder, b) obtained by modeling in the Matlab/Simulink environment [62] for specific conditions

This reduces the width of the occupied spectrum on the transmitting side, and, accordingly, the spectral efficiency of this technology increases.

The structure of the radio path of the receiving part of the transceiver actually repeats that used in the LTE system before the output of the Fourier transform, where the symbolic sequence of the group signal is formed (Fig. 3.46). Then it is allocated to individual subbands and DSP of symbols is processed, which results in the information sequences of the individual subchannels.

3.3.6 The Receiver for System of Mobile Multimedia Radio Broadcasting

Systems of local broadcasting with **DRM+** (digital radio mondiale) format developed the DRM concern, and the standardization RAVIS (real-time audio visual information systems) of Russian development provides the mobile audio-visual TV-broadcasting in two frequency bands of the VHF range.

Considered digital audio-video systems provide the radio broadcasting signal reception in I–II bands of VHF frequency range (65.8–74.0 and 87.5–108.0 MHz) at keeping the frequency distributions. Along with the increase spectral effectiveness compared to the analog FM-broadcasting, digital systems have the increased service reliability and a quality of the audio broadcasting, as well as a possibility to introduce new multimedia services. Digital methods of signal processing based on modern compression methods for video and audio allow provision the high reproduction quality of information reception at motion. Besides, such systems suppose a possibility of revision and service widening without revision of frequency distributions, including the arrangement of the single-frequency networks.

The advantage of such systems consists in a possibility to increase the effectiveness of VHF-range utilization, when in the single standard FM-broadcasting channel we can:

- To transmit more than ten stereo audio programs with the CD quality or several multi-channel audio programs.
- To provide the translation of video programs for the mobile subscriber at sharp reduction of TV broadcasting cost in region with low population.
- To provide the stable mobile signal reception under conditions of multipath properties (the urban building up, the uneven terrain, water areas).
- To reuse of frequencies in different inhabited localities.
- To create single-frequency networks along highway and railroad roads.
- To decrease expenses on creation the mobile TV broadcasting compared to networks implemented on the principles of the stationary networks.
- To essentially decrease (by several times) of power consumption of radio transmission devices at calculation to the single broadcasting program.
- To organize communication systems for special services and an army, as well as for warning system under emergency situations.

Main parameters of RAVIS system are presented in Table 3.3.

The RAVIS provides the transmission of three independent logical data channels: the channel of main service (MSC), low-rate channel (LRC), and reliable data channel (RDC). At that, channels LRC and RDC can be independently switched on or switched off. If channels are switched off, their carriers are used for the MSC channel.

Depending on services configurations, the different transfer parameters are possible:

1. Radio channel bandwidth is 250 kHz, modulation is 64-QAM, and the average data transmission rate is 900 kbit/s:
 - Video program with two stereo accompanied channels – 700 kbit/s.
 - Musical program – 84 kbit/s.
 - Three stereophonic programs – $32 \times 3 = 96$ kbit/s.
 - Data channel – 20 kbit/s.

Table 3.3 Main parameters of RAVIS system

Parameter	Value		
Type of modulation	OFDM		
Channel coding rate (MSC)	1/2; 2/3; 3/4		
Carrier modulation (MSC)	QPSK; 16-QAM; 64-QAM		
Radio channel bandwidth, kHz	100	200	250
Data rate (MSC), kbit/s	80–350	160–710	200–900
Number of carriers	215	439	553
Number of informational carriers	196	400	554
Distance between end carriers, kHz	95.1	194.7	245.3
Interval between carriers, Hz	444.44		
Duration of useful part of symbols, μs	2250		
Duration of the protection interval, μs	281.25		
Duration of OFDM symbols, μs	2531.25		
Duration of OFDM frame, s	103.8		

2. Radio channel bandwidth is 200 kHz, modulation is 16-QAM, and the average data transmission rate is 410 kbit/s:
 - Video program with stereophonic accompaniment – 360 kbit/s.
 - Stereophonic audio program – 32 kbit/s.
 - Data channel – 18 kbit/s.
3. Radio channel bandwidth is 100 kHz, modulation is QPSK, and the low data transmission rate is 80 kbit/s:
 - Two stereophonic audio programs $2 \times 32 \text{ kbit/s} = 64 \text{ kbit/s}$.
 - Data channel 16 kbit/s.

The structural diagram of the RAVIS at using the diversity reception is shown in Fig. 3.55.

The diversity reception (as transmission) permits noise immunity increase, especially at multipath reception, by means of some complication of the receiver architecture. The input unit is usually implemented on the single IC: the LNA and FC. The device providing synchronization of the input signal components attached to different subcarriers caused by OFDM technology is the important component of the receiver architecture. The unit performed the fast Fourier transform (FFT) that forms the continuous informational stream from the time segments obtained at processing of the separate subcarrier. In considered system, the tuning accuracy of the receiving signal is kept up to 1 Hz at providing of the input signal SNR up to 40 dB. Application of channel coding algorithms using the binary block code for error correction on Bouse-Choudhury-Hokwingam (BCH) and codes with small density of even parity provides the noise immunity growth by 2–3 dB. Utilization of diversity reception switcher, which is typical for the technology of diversity reception, reduces the probability of incorrect reception. Depending on receiving signal parameters, three types of channels (MSC, LRC, and PDC) are formed in demultiplexer, which provide the reception of the mobile TV broadcasting, channels of stereophonic broadcasting, or the data transmission. The used systems of peal suppression of the OFDM signal allow two times increase of the emitted signal power of the transmitter at keeping of its consuming power.

3.3.7 The Structural Diagram of the Receiver for IEEE 802.11 Standard

In the local computing networks (WLAN) constructed on the base of IEEE 802.11 standard and exploited in the range of 2.4 GHz, the signal formation by the method of direct sequence spread spectrum (DSSS) is used for transmission in the radio channel. This provides the signal exchange rate up to 11 Mbit/s in the operating frequency range allocated for WLAN and related to the non-licensed ISM range of 2.4,...,2.4835 GHz with the width of 83.5 MHz.

Application of DSSS anticipates the utilization of the Barker code with the length of 11 bits for *spreading of the informational sequence*, when 11-bit (11-chip sequence) formed by multiplication modulo 2 with the Barker sequence associates with 1 informational bit. Then the signal is modulated using the differential 2- or 4-position phase modulation (DBPSK or

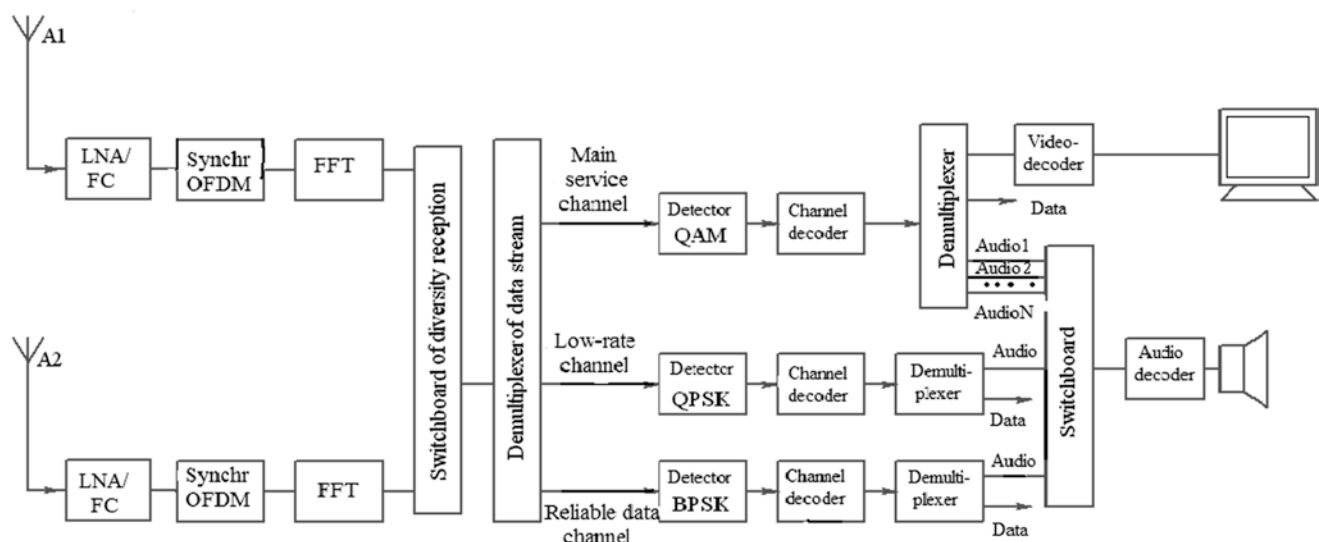


Fig. 3.55 The structural diagram of the RAVIS using the diversity reception

DQPSK) providing the transmission rate in the channel of 1 or 2 Mbit/s at modulation frequency of 11 MHz. Utilization of the additional code modulation CCK allows the growth of the transmission rate up to 5.5 Mbit/s or even up to 11 Mbit/s. The CCK method uses DQPSK modulation in the radio front-end and consists in extraction from the informational stream of groups of 8 bits. These 8 bits modulate the carrier with the frequency of 11 Mbit/s forming the CCK symbol with the phase defined by the phase of the previous chip, and its variation is defined by combination of values of the two-rate bits. If the 4 bits instead of 8 bits are used for formation of CCK symbols, the transmission rate will reduce from 11 Mbit/s to 5.5 Mbit/s.

On the receiving side, at synchronous reception, the decoding device recovers the phase value of the CCK symbol, and on them it recovers the value of informational bits. The recover accuracy is given by properties of the formation algorithm of the CCK symbol defining by the Walsh-Hadamard matrix. The typical feature of the Walsh-Hadamard sequence is high correlation properties allowing easily extraction (among receiving sequences) of sequences, which are obtained from its own subscriber, but arrived by different ways (different phase shifts) from any other.

CCK modulation is a variant of the block code, but together with simplicity of realization and the noise immunity, it has a high distribution irregularity in the phase space, which decreases the effectiveness of error minimization at decoding. Therefore, together with CCK coding, the optional method of convolution coding PBCC has got the wide distribution. At the rate of 11 Mbit/s, the informational stream passes into 6-bit register with the adder providing the replacement of 1 bit of the initial sequence by two bits of the coding sequence, which define the one 4-position QPSK symbol. For the rate of 5.5 Mbit/s, for transmission of 2 bits of the sequence, we can use BPSK. Application of 8-PSK allows replacement of 3 bits by the one symbol of the modulated signal providing the transmission with the rate of 22 Mbit/s.

The IEEE 802.11b standard describes requirements to system parameters, which are acted in the non-licensed frequency ranges (in the USA initially 2.400...2.4835 GHz). The spectral mask of the single channel specified by the standard and defined by peaks of the amplitude spectrum of the type $(\sin x)/x$, at carrier modulation by rectangular pulses, is 22 MHz and has sufficient attenuation of the adjacent channels (-30 dB), which permits to place only three channels in the band of 83.5 MHz.

The sets of IC specially intended for the definite standard (chipsets) are produced by the different manufacturers: Texas Instruments, Intersil, etc. Together with other standard ICs (the low-noise amplifier, a mixer, etc.), they are included into the whole device (a transceiver).

For instance, in the structure of chipset [99] PRISM2 (Fig. 3.56), the four ICs are included: the MAC-controller HFA3841 providing an access to the transmission medium, the digital processor HFA3861B, the quadratic modulator/demodulator on the intermediate frequency HFA3783, and the RF/IF converter HFA3683A.

Usually, the high-frequency antenna amplifier HFA3963 is added to the mentioned ICs. The chipset structure (Fig. 3.56) uses the main and additional antennas; a choice between them is fulfilled by the MAC-controller, which is defined by the operating frequency range and the parameter of the service quality (QoS). The antenna switcher connects the antenna block to the receiver input or to transmitter output. The output power amplifier in the transmitter has (HFA3963) the pass-band filter, which is implemented on the integral technology, and the special device for the transmitter output level monitoring. In the receiving module, the receiving signal passes to the input of the low-noise amplifier with the controlled gain. The gain control is provided from the main digital section (Base Band, HFA3861B) with the help of AGC system. The band of receiver frequencies is extracted by the discrete pass-band filter and passes in the balanced converter, in which there is a transfer of radio frequencies onto the intermediate frequency with the further extraction of useful conversion products by the pass-band ceramic or SAW filter. The receiving part of the integral circuit HFA3783 contains an amplifier with the controlled gain providing the operation of I/Q-demodulator with low level of the spurious products. LPF connected with demodulator outputs implemented as digital filters extract the modulating frequency band forming two digital streams passing to ADC of the digital section.

The further chipset PRISM3 development by the company was based on the architecture of the direct conversion **ZIF** (zero intermediate frequency). ICs of other manufacturers are developed on the base of the same principle, for example, the single-chip transceiver SA2400 from Philips Semiconductors company for the range of 2.4 GHz that contains RF block of the receiver and transmitter, quadratic modulators/demodulators, the master oscillator on the base of VCO, and fractional-N synthesizer. The completely integral pass-band channel filters, the device of automatic gain control, the control bus by the main device blocks, the interface for linking with the base-band processor, etc. are included in this chipset.

In the version of IEEE 802.11a standard and the latest, the OFDM technology is used, and at used bandwidth of 20 MHz, for OFDM symbol formation, 52 subcarriers are required, although the nominal value at calculation of Fourier transform should be a multiple of 2^m (the nearest value is 64), which is specified the subcarrier offset: $\Delta f = 20 \cdot 10^6 / 64 = 312,5$ KHz. Subcarriers are modulated by BPSK, QPSK, 16-QAM or 64-QAM. The formed OFDM symbol is the variety of all subcar-

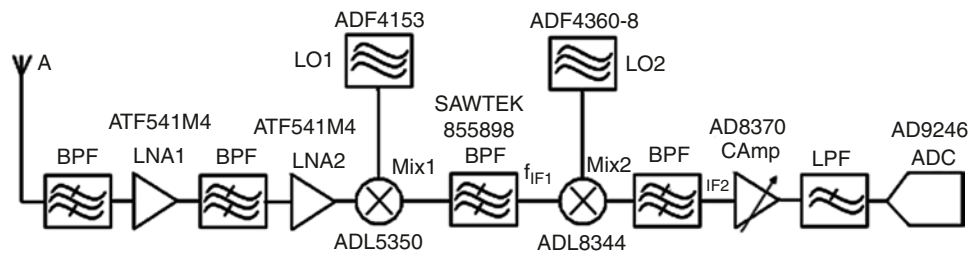


Fig. 3.57 The architecture of the WiMAX receiver with the double conversion and the component base of Analog Device company

The double-frequency conversion allows (chosen the high value of the first intermediate frequency) provision of the effective suppression of the image channel by comparably simple measures of technical implementation: the filter of the strip construction or on the discrete components.

The value of the first intermediate frequency f_{IF1} should be chosen taking into account that at frequency of received signals of 2.3,...,2.4 GHz and at lower tuning of the local oscillator, the conversion products on the output of the first mixer will not be the spurious channels for the second mixer. Possible values of f_{IF1} , starting from the mentioned condition, are calculated with the help of the nomograms или the software Mixer_Spur Calculator. exeVer.2.1. For a given operating range, the first intermediate frequency can be chosen in the interval of 210...400 MHz. The important condition of f_{IF1} is the existence of the pass-band filter produced by IC manufacturers with required characteristics (the bandwidth, the attenuation in the pass-band, the image channel suppression on the second intermediate frequency f_{IF2}) for the chosen first intermediate frequency f_{IF1} . The filter on the surface acoustic waves (SAW) produced by the Sawtek company with the central frequency of 374 MHz was chosen as the pass-band filter.

The second intermediate frequency can be chosen from the recommended frequency ranges of 10...70 MHz and 107...140 MHz. Thereof that the operating band of the receiver of the WiMAX standard does not exceed 20 MHz, we can choose $f_{IF2} = 70$ MHz, which reduces the danger of the local oscillator's harmonic leakage into an antenna, and besides, for this central frequency, there is the wide SAW filter choice. As an alternative, we can choose filters on the limped LC components.

Pass-band filters with the low loss level in the pass-band of 2.3,...,2.5 GHz and the required attenuation in the rejection band are produced by various companies, Anatech Electronics Inc., K&L Microwave Inc., and Digital Communication Inc., which allows easy selection of required parameters. SAW and ceramic filters has found the most distribution. At that, SAW filters having the insertion losses of 2...3 dB and abrupt edges of frequency responses have found the widest distribution. The amplifier, implemented on the HEMT transistor, provides a wide dynamic range ($IIP3 = 18$ dBm), low-noise figure ($NF = 0.9$ dB), which reduced the resulting noise factor of the RF front-end receiver. The noise properties of input blocks and LNA1 mainly define the relevant noise factor of the receiver. The pass-band filter on discrete components included in the LNA1 input provides the image channel suppression by 60 dB simultaneously reducing the noise factor on the receiver, which is caused by external sources. Connection of the rejection filter on the LNA1 output, which is tuned on the image channel, allows not only increase of its suppression but also decrease of the wideband noise level passing on the first mixer.

The passive mixer is implemented on ADL5350 IC having besides the embedded buffer amplifier of the LO signal. On the LO input of the first mixer, there is the additional LPF to suppress the second harmonic of LO, which causes the input signal second-harmonic conversion in the mixer and increases the component number of conversion products in the output spectrum.

The first LO (VCO1) is realized in IC VCO190-1960 T, included in the structure of IC ADF4153 containing, besides VCO, the frequency synthesized as well. Typical variants of connection of used ICs together with matching and auxiliary circuits are discussed in descriptions of the appropriate blocks.

To extract the useful conversion product, the SAW filter with the central frequency 374 MHz is connected in the Mix1 output through the matching circuits. The matching circuits are made on the external LC components providing the standard 50-Ohm resistance of the input and output of the filter.

Conversion onto the second intermediate f_{IF2} is performed in the active Mix 2 on IC AD8344. The choice of the LO2 voltage level (VCO2) provides not only high conversion coefficient of the converter, but the low level of spurious conversion products. Since the local oscillator frequency varies insignificantly ($f_{IF1} = 374$ MHz), this simplifies its implementation. The LO voltage with the frequency 304 MHz passes to the Mix2 through the external matching transformer providing the standard value of resistance of the connecting local oscillator.

The useful conversion product from the Mix2 output on the frequency of 70 MHz passes through the discrete LPF of fourth order to the amplifier with the controlled gain. The microcircuit AD8370 provides the large value of the interception point of intermodulation distortions of the third order on input (IIP3) and the control range more than 40 dB. Programmed control of gain allows mating of the radio front-end dynamic range and limits of power variation of the input signal. The elliptic filter, which attenuates interferences of higher frequencies caused by the leakage of LO frequency and reducing the sensitivity level of ADC, is chosen as a load of the controlled amplifier. The elliptic filter provides the protection against aliasing suppressing the wideband noise creating by amplifiers.

The analog-to-digital converter AD9246 has a wide dynamic range and the large bandwidth on the analog input, which allows its utilization up to frequencies exceeding f_{IF2} by almost the order.

3.3.8.2 The Receiver of Direct Conversion

A volume of rendering services in wideband access WiMAX networks of the IEEE 802.16 standard is mainly defined by technological possibilities of the companies-manufacturers of the element base. Companies Intel, Runcom, Fujitsu, Sequans Communications, and others can be concerned to the number of such manufacturers of chipsets with support of the “mobile WiMAX,” i.e., the OFDMA technology [89]. Presented by Intel, the set of microcircuits for the subscriber equipment WiMAX Connection includes the telecommunication processor and three-range radio module without analog mating or the double-range full-functional module. This chip-set is intended for combination of IEEE 802.16e a IEEE 802.11 technologies (the mobile WiMAX in the range of 2.5...2.7 GHz and Wi-Fi in ranges of 2.4...2.48 GHz and 5.1...5.8 GHz).

The SQN1130 processor developed by the Sequans company contains the integrated module of analog signal processing; therefore, for its complete application in the mobile station, the only RF transceiver is necessary. This VLSIC is distinguished by the very low power consumption: 280 mW in the operating mode and 10 mW in the waiting mode. Realized OFDMA technology with 512 and 1024 subcarriers provides the capacity higher than 30 Mbit/s and time duplexing. As the RF module for SQN1130, the company offers IC of double-channel RF transceiver SQN1140 of direct conversion intended for operation in the range of 2.3,...,2.7 GHz.

From the large enough number of proposals of analog devices for WiMAX systems, we consider the most typical method of construction of RF transceivers of direct conversion of the NXP company (Fig. 3.58).

The specialized IC UXA23466 realizes the direct conversion transceiver, which covers the frequency range of 2.3,...,2.7 GHz and has the minimal number of connected external components. The direct conversion of the input signal allows elimination of filter utilization in the IF section, which simplifies and makes low prices the transceiver as a whole. The signal power on the transmitter output is 77..0.87 dBm (depending on the IC type) at the control range of 74 dB; the level of the receiver noises is 2.5,...,3.5 dB. The IC of the transceiver UXA23466 supports two receiving and transmitting channels using the MIMO technology. The diversification transmission means that several antenna channels are used for transmission of the single data stream. This technique is directed to the improvement of the signal transmission quality, to the struggle with the one of the very strong influencing factor in the radio channel – fading. At the same time, this allows indirectly increasing the signal transmission rate.

Improvement of the network possibilities constructed with application of the IEEE 802.16 m standard and later versions allows the support of MIMO modes. The minimal configuration in the downward channel anticipates the application of two transmitting antennas on the base station (in the sector) and two receiving antennas on the mobile station. Totally, we can use up to eight transmitting and the same number of receiving antennas. The multi-user MIMO mode is introduced, when information translation to different subscribers is possible on the same frequencies.

Introduction of variations in the frame structure allows provision of compatibility of earlier versions of the IEEE 802.16 standard and other standards of the wideband transmission, for instance, LTE. For this purpose, we introduce variations in the time zone formation order, when during the duration of several sub-frames, the traffic is transmitted exclusively for devices of the single standard, for example, IEEE 802.16e or IEEE 802.16 m.

Implementation of the IEEE 802.16e standard called the “mobile WiMAX” permits to widely use possibilities of mobile applications, which makes the wideband access system the competitor of cellular communication technologies of third and fourth generation. In this case, issues of the relay subscriber transfer become relevant, which was not typical for subscribers of the WLAN network and for supporting of stable communication with subscribers being in motion. In this case, subscribers should have the access to the network directly in the process of motion (e.g., the subscriber in the car).

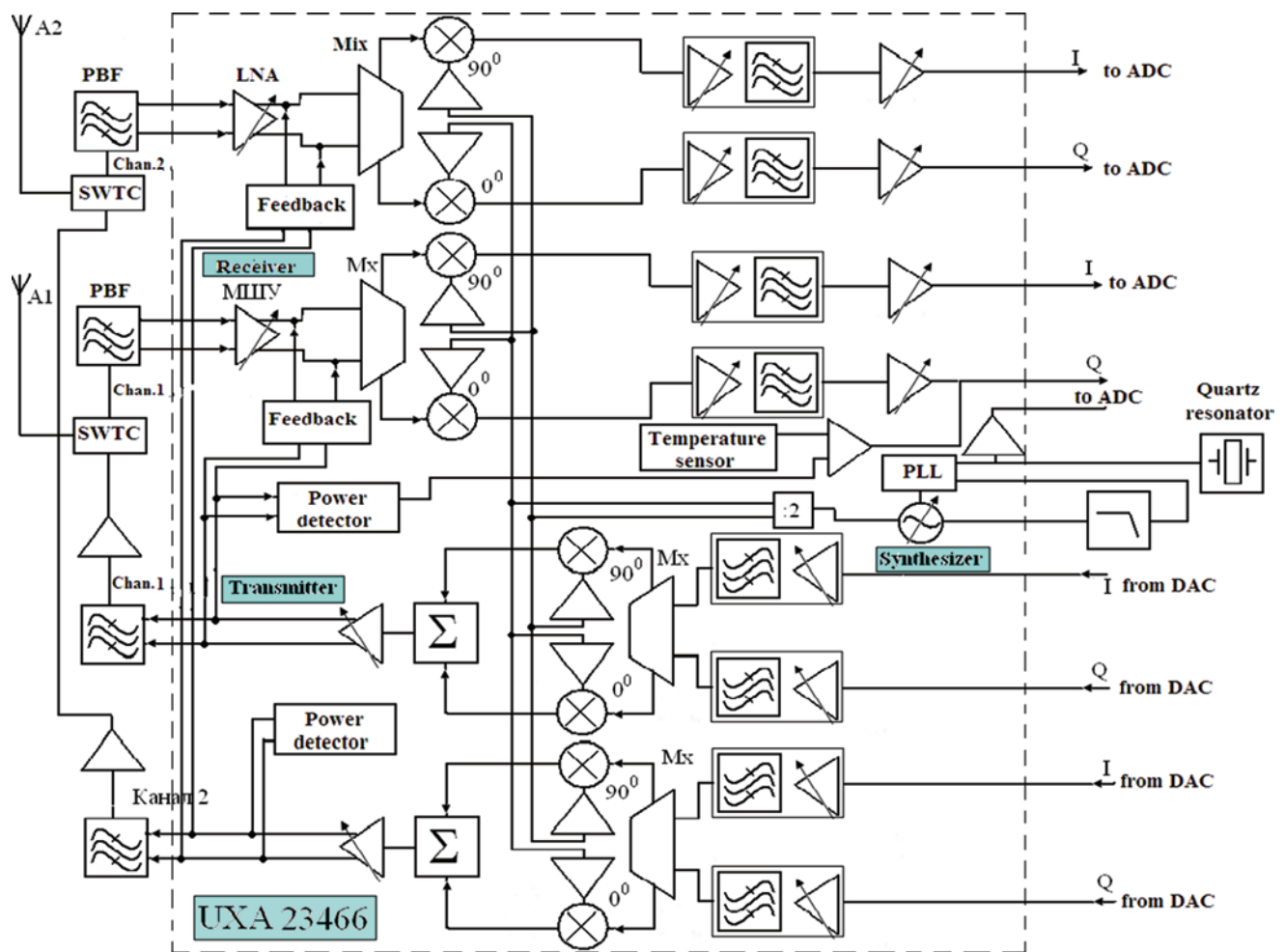


Fig. 3.58 The most typical method of construction of RF transceivers of direct conversion of the NXP company

3.3.9 The Transceiver Architecture of the DECT Technology

The main criterion at development of integrated circuits used in subscriber terminals is their low cost and large volumes of productions. At that, at realization of the radio front-end, both specialized ICs and ICs with high degree of integration with limited quantity of add-on elements can be used, which usually are the matching circuits with connected external units.

Due to features of the DECT technology [69], the applied ICs must satisfy the following requirements:

- To provide the coding of the carrier frequency (the range 900 MHz) by the sequence of digital signals (the technology of signal spectrum spreading by the direct sequence DSSS).
- The power amplifier of modulated signal must provide 100 mW on the transmitter output.
- To provide switching between the receiver and transmitter.
- To provide amplification of received signal by the amplifier with low own noises.
- To switch on of the I/Q-demodulator.

The IC set of the subscriber terminal contains two ICs: the specialized IC of the antenna interface and IC of the radio front-end (the signal frequency section and the intermediate frequency section). To obtain the optimal ratio price/quality, IC of the antenna interface, which has the high technical indices, can be realized on GaAs FET with the Schottky barrier (MESFET – Metalized Semiconductor Field-Effect Transistor), and in IC of the radio front-end, the silicon bipolar transistors are used (Fig. 3.59).

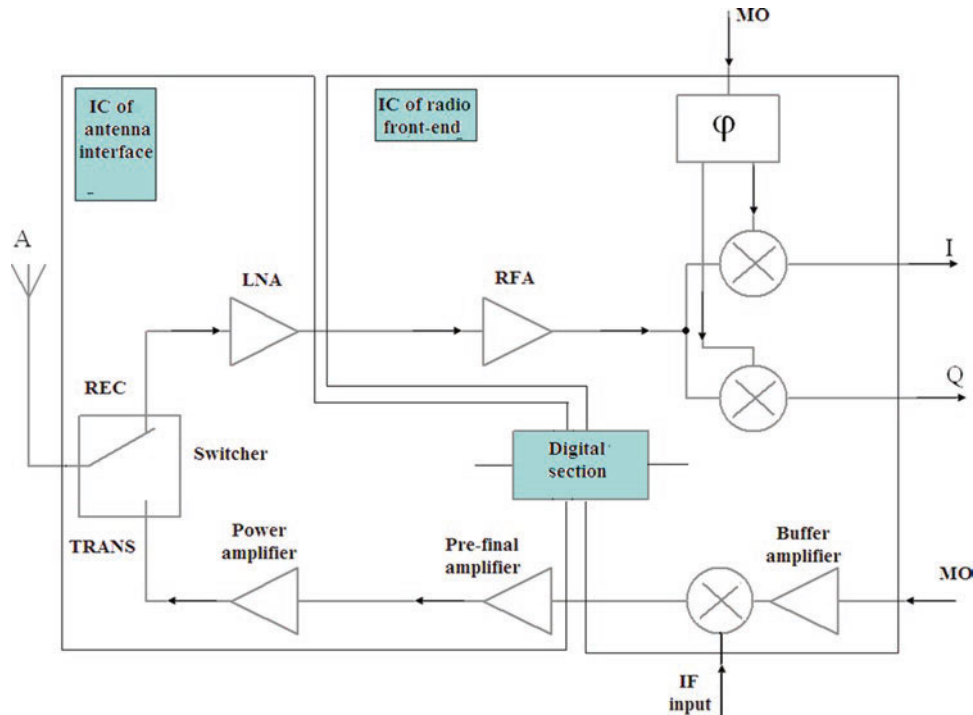


Fig. 3.59 The IC set of the subscriber terminal contains two ICs: the specialized IC of the antenna interface and IC of the radio front-end

Specialized IC for the antenna interface realized of FET with the Schottky barrier includes the switcher reception/transmission, the power amplifier, and the low-noise amplifier. The unipolar power source with 5 V voltage is used as the IC power source. The power amplifier in the transmission section provides the amplification of 23 dB and the power level in the transmission mode of 100 mW, and in the reception mode, LNA provides the gain of 13 dB at the noise factor less than 3 dB. Control of the antenna switch operation is performed from the digital section by the output voltage of the standard circuit on TTL or the CMOS structure.

The power amplifier is implemented as the three-stage amplifier on FET with the Schottky barrier with auto-bias. The final stage is realized on the field-effect MOS transistor with depletion, which operates in the AB mode at cutoff voltage ≈ 0.9 V and the consuming current of ≈ 60 mA and has the gate width of 1.2 mm. In the pre-final stages, the enriched FET are used, which permit not to pass in the cutoff mode, when the transmission mode is switched-off, at supply from the single source.

The IC architecture [98], which provides the antenna interface (Fig. 3.59), is presented in Fig. 3.60.

The signal from the duplexer output passes into the IC antenna input (in reception mode), and with the help of antenna switch, which is controlled by the signal from the digital section (Control commands), it connected to LNA. The low-noise amplifier provides the amplification of 12 dB at the noise factor of ~ 2 dB, at connection in input and output of matching resistances of 50 Ohm. Its output is connected through RFA to the input of the I/Q-demodulator of the specialized IC (Fig. 3.61).

In the transmission mode, the signal from the output of power amplifier is fed through the matching circuits to the IC point Switch Input and the antenna switch. The signal from the PA output may additionally pass through the filter (the resonance circuit), which restricts the frequency band on the transmitter. The RF amplifier of the transmitter consists of the stage of the pre-amplifier, the stage of pre-final amplifier, and the final power amplifier.

The specialized IC including components of the radio-frequency and intermediate frequencies (Fig. 3.62) is realized on the integrated bipolar transistors [98] and includes the bi-phase modulator of the transmitter together with buffer amplifiers of the reference (master) oscillator and also the quadratic demodulator of the receiver.

The power level of MO in IC input is -10 dBm, which is amplified to 0 dBm with the help of MO amplifiers and passes on the transmitter modulator and to the I/Q demodulator. In the transceiver, the direct double-position modulation and the direct quadrature demodulation are used; for this, the signal from the single external master oscillator (the synthesizer) is used. The device, which shifts voltages in branches of the demodulator by 90° , is embedded in IC, which simplifies requirements to the interface. In the input of the double-position phase modulator, we should provide the pulse voltage (a meander)

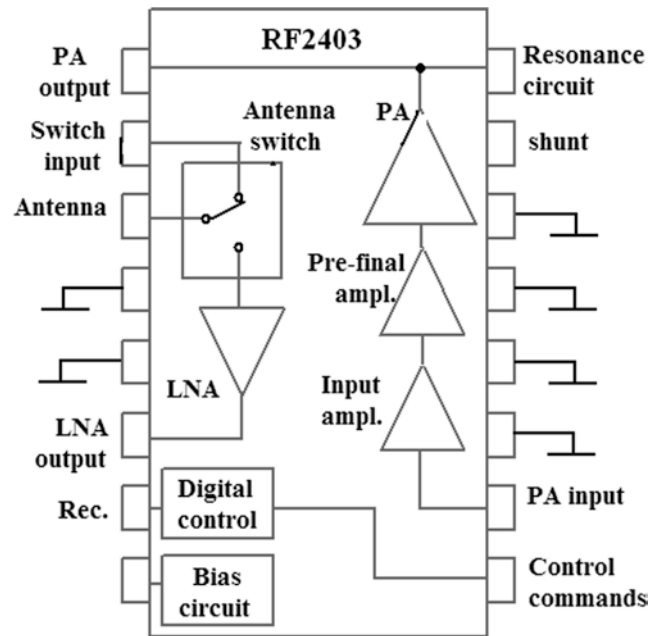


Fig. 3.60 The IC architecture, which provides the antenna interface

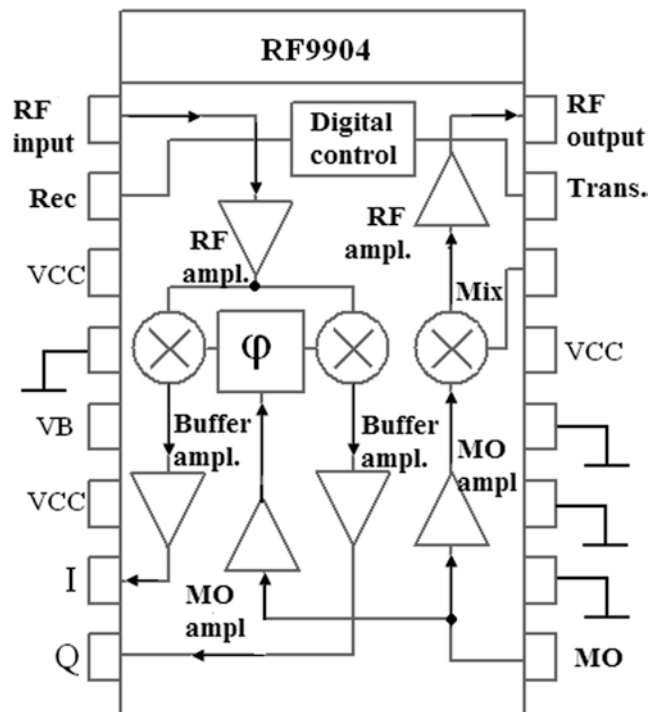


Fig. 3.61 The specialized IC including components of the radio-frequency and intermediate frequencies

with the amplitude more than 0.5 V and the repetition frequency of 0...3 MHz. The carrier suppression is performed not less than by 25 dB with regard to the carrier level at maximal signal level on the IF input. The maximal level on the IC output is -10...-15 dBm.

The quadrature demodulator converts signals directly on the frequency of 915 MHz. The input signal can act in the frequency range of 910...930 MHz creating the signal on the output of the I/Q demodulator in the range 0...3 MHz. Asymmetry

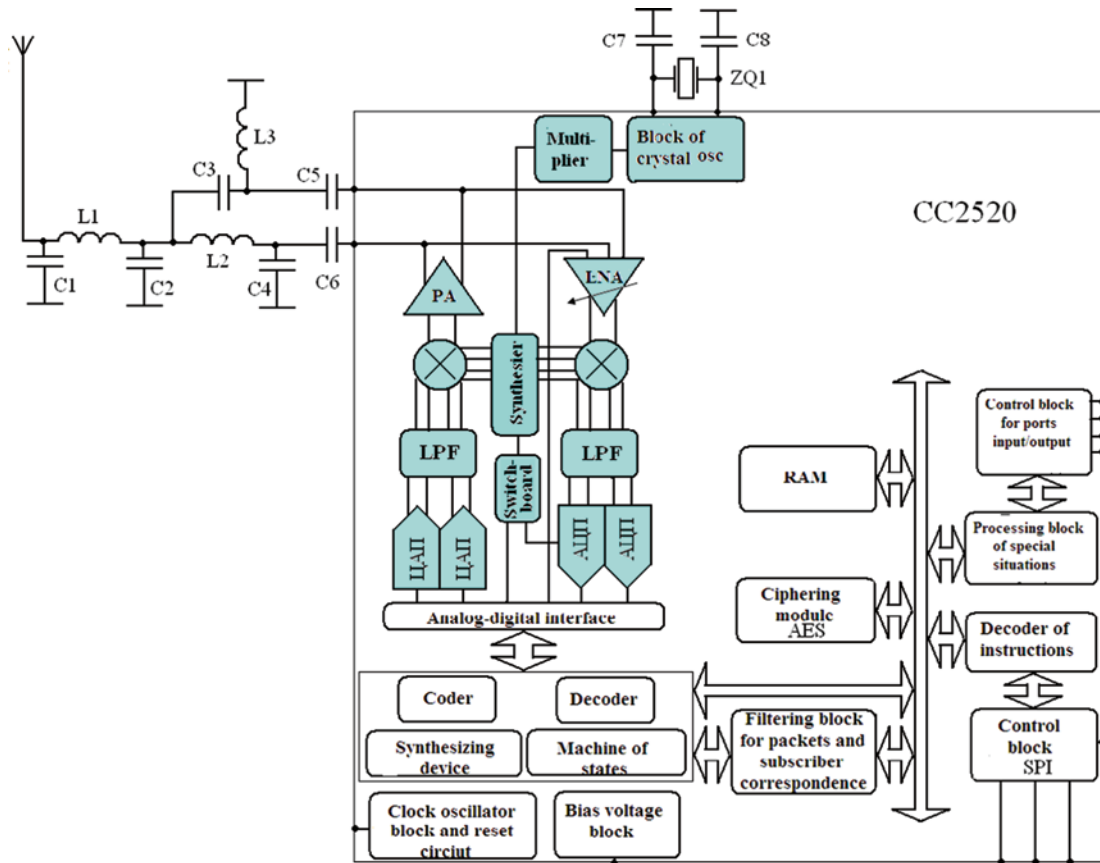


Fig. 3.62 The transceiver architecture and the IC CC2520 structure

of demodulator branches should be less than 1 dB in amplitude and $\pm 5^\circ$ in phase. The amplitude level of the pulse signal on the IC output must be not less than 1.4 V in the point of 1 dB-compression.

As a power source, we can use the single power source with the DC voltage 5 V.

3.3.10 The Receiver of the IEEE 802.15.4 Standard with ZigBee Protocol

Transceivers on IC CC2520 and CC2530 from Texas Instrument are the typical representatives of receiving-transmitting devices, which are realized the ZigBee/IEEE804.15.4 [44] protocol and have the best technical indices compared to variants of other manufacturers. The transceiver CC2520 from Texas Instruments with the ZigBee /IEEE 802.15.4 protocol is used in the non-licensed frequency range of 2.4 GHz; contains the expanded hardware support to process packets, buffering, ciphering, and authentication of data; and performs of channel parameter estimation and identification of the communication quality. The structure of creating networks can realize the configuration of point-to-point, the star, or, for systems with preliminary defined stack ZigBee Pro, the mesh type. Developing networks provide the support of all solutions based on the IEEE 802.15.4 standard: ZigBee, ZigBee Pro, ZigBee RF4CE, and GloWPAN. When they are applied in the frequency range of 2394...2507 MHz, they provide the receiver sensitivity of -98 dBm at signal bandwidth of 5 MHz at the transmission rate of 250 kbit/s. The support of low-level packet functions allows simplification of the controller structure providing compatibility with Bluetooth and Wi-Fi systems. Using the low-voltage supply with the voltage of 2.0,...,3.6 V and at current consuming from the power supply in the transmission mode of 34 mA, we ensure the nominal transmitter output power of 4.5 dBm. In the reception mode, current consumption is 25 mA. The nominal value of receiver sensitivity is -97 dBm (at its best value -103 dBm), and the communication distance is not less than 400 m (within the limits of direct visibility) and the adjacent channel suppression (at detuning by ± 5 MHz) of 49 dB. The current consumption in the sleeping mode is less than

one microampere. The low-energy consumption allows network structure, which has the battery supply and the presence of sleeping blocks.

The transceiver architecture and the IC CC2520 [100] structure are shown in Fig. 3.62.

The integral circuit is controlled by the micro-controller through the interface bus of the system programming **SPI** (System Programming Interface) and pins of the control unit of ports input/output **GPIO** (General Purpose Input/Output).

Execution of instruction or any external events (for instance, reception of the data packet) creates special situations, which can be resolved in the processing block of special situations, which provide the network flexibility and the reaction sharpness on the event. The micro-controller can be connected to one or several ports of input/output. The control of each of them is fulfilled independently by the block of port control by means of recording in the appropriate registers.

The RAM block is used for temporary storage of received and transmitted data, as well as intermediate results of signal processing and current values of monitored parameters. An access to RAM memory and the **FIFO** (First In, First Out) buffer are controlled by instructions stored in the ciphering module of the data processing block.

The block of data processing provides an execution of the most complicate instructions in data processing: memory control, provision of connection safety (switch on the ciphering module), and error correction **AES** (Automatic Error Correction).

The analog part of a transceiver includes the filter system extracting the frequency band of the operating range (L1, C1, C2), and a possibility of circuit utilization for operation in the range of 2.5 GHz or below 1 GHz is provided by connection of two filters: LPF (L2, C4) and HPF (C3, L3). The input filter inductances can be realized in the discrete technology or as segments of non-symmetric strip lines. The signal of the one from sub-ranges passes into the LNA input with controlled gain. The signal of LNA gain control is produced by the AGC system maintaining the signal level constancy in the demodulator input. The radio front-end is fulfilled as the receiver of direct conversion, which reference signal for the converter (which is synchronous and in-phase with regard to the input signal) is produced in the VCO, which parameters are corrected from the digital section. The demodulator is usually implemented as the circuit of quadratic balanced frequency converter on the Gilbert cell.

The frequency synthesizer forms the frequency grid, which stability is defined by the frequency of the quartz oscillator, which is at the same time the source of the clock pulses used in the digital section and in the analog section for VCOs adjustment on the operating frequency.

The quartz oscillator block with the external resonator (ZQ1) provides the synchronization in the digital part of the transceiver and performs the operating frequency adjustment of a receiver and a transmitter.

The passive low-frequency filter at the demodulator output extracts the frequency band, which is defined by the modulation type and the pass-band of coordinator LPF suppressing the collateral conversion products. The in-phase and quadrature components on the LPF output are converted in ADC into two digital streams, which further form the informational stream.

In the digital part of the data processing section, besides the control signal of LNA gain, the majority of the control signals for the analog part of section is formed in the machine of states (reset/maintaining the power supply, for FIFO buffers). On the digital level, the machine of states provides the correction of event sequence (calibration of the frequency synthesizer before the transfer to the active mode) and fulfills the step-by-step processing of received frames (check of a number of received bytes, check the packet numeration, confirmation of reception, etc.).

The bias voltage block provides the required values of supply voltages for analog and digital parts of IC and corrects the operation mode of transistors eliminating the spurious DC component on the demodulator input.

3.3.11 The RF Front-End of the Trunking Communication Receiver

Trunking systems of the mobile radio communication, which were the “polygon” for development of many ideas implemented later at realization of cellular systems, are distinguished by the great variety of user frequency ranges, modulation types, and companies that manufactured an equipment for these systems [7, 8]. At present, both analog and digital trunking systems are exploited, which are used for the receiver architecture of the super-heterodyne type and direct conversion type.

The transceiver structural diagram of the carrying radio station DJ-180 of Alinco company is shown in Fig. 3.63.

The carrying radio station in the Smartrunk2 system is used in the frequency range of 137.000–173.995 MHz/144.000–145.995 MHz with the varying frequency grid step 5; 10; 12.5; 15; 20; and 25 kHz. The transmitter carrier frequency (upper frequency range) is modulated in phase (F3E) with maximal frequency deviation of ± 5 kHz. The transmitter output power, at operation on the non-symmetric 50 Ω -antenna, is 2 W for power supply with the voltage 7.2 V. The receiver sensitivity cannot be less than -10 dB μ V at 12 dB SINAD.

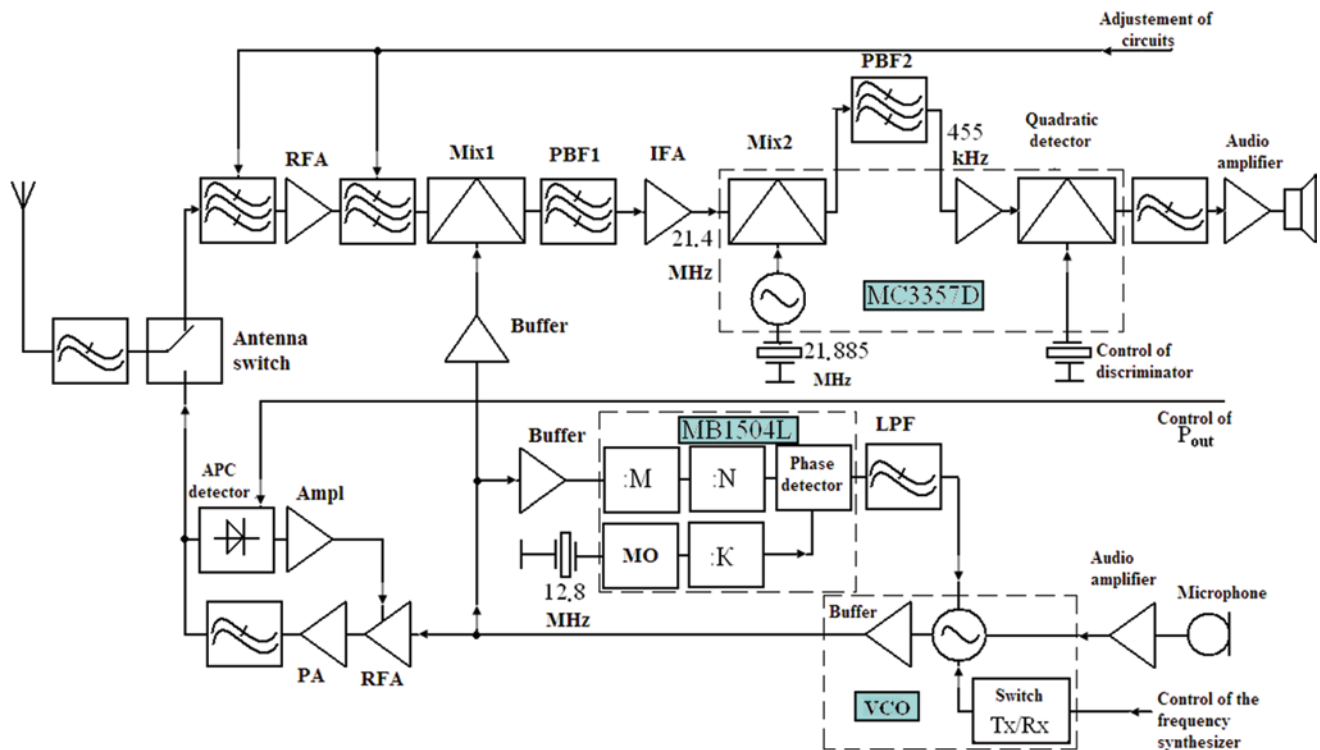


Fig. 3.63 The transceiver structural diagram of the carrying radio station DJ-180 from Alinco company

The transceiver receiving part is constructed according to the super-heterodyne structure with double-frequency conversion of the input signal. The signal from an antenna passes through LPF and the antenna switches into the preselector, which contains two pass-band filters with adjustment of the central frequency on the middle frequency of the operating range, and then it is amplified by the low-noise RF amplifier and passes on the first mixer (Mix1).

The voltage of the first LO is formed with the help of VCO, which frequency varies by varicaps in the VCO circuit. VCO adjustment is performed with the help of DC voltage of the LPF output in the feedback loop, which is produced by the phase detector. The voltage of the PLL system output caused by the frequency difference of VCO and produced by the master oscillator stabilized by quartz resonator is fed simultaneously on the varicaps in the receiving section providing adjustment of PBFs at frequency variation of the operation channel.

The useful component of the converted signal on the Mix1 output with the frequency of 21.4 MHz is extracted with the help of the monolith quartz filter. After its amplification in the IFA1, the signal passes to the Mix2 containing IC MC3357D, where it is multiplied by the signal of the second LO with the frequency 212.885 MHz producing the oscillator stabilized by the quartz resonator. The signal of the second intermediate frequency $f_{IF2} = 455$ kHz is extracted by the ceramic filter, which is in the Mix2 output. For elimination of the spurious AM, the signal from the filter output is applied to the amplifier-limiter, and then it is detected with the help of quadrature detector. The reference frequency of the frequency discriminator is defined by the quartz resonator.

The demodulated signal passes through LPF extracting the voltage of the audio frequency, which then is amplified in audio amplifier and passes to a loudspeaker. To exclude the noise influence in the pause period, the special circuit of noise suppression (Squelch Circuit) is included in the audio section, which operation threshold is specified by the variable resistor at the stage of circuit adjustment.

The transmitting part of the transceiver begins from the internal or external microphone, which output signal passes to the preliminary distortion circuit and through LPF to the circuit of frequency control of VCO (by variation of varicap capacitances in the oscillator circuit). The modulated signal, after amplification in the radio front-end, passes through LPF to the antenna switch, which provides the switching of transmission/reception (T_x/R_x) mode.

The applied system of automatic power control (APC) of the transmitter provides the constant value of the output power, eliminating the overload of the power amplifier.

3.3.12 Receivers of Analog Signals in Fixed Communication Systems

At present, receivers of analog signals widely used in systems of fixed radio communication or radio broadcasting have some features at their development and operation. Let us consider it on the example of radio broadcasting receivers.

Non-prescribable feature of the broadcasting systems is its organization according to the principle “point many points” and also organization in the frequency band allocated to the transmitter for realization of the single broadcasting channel and the single physical and accordingly the single logical channel. At subscriber movement in the reception zone and at exit from it, the relay subscriber transfer into the service zone of another transmitter has not occurred, as it happens in the systems of mobile communication.

Radio broadcasting receivers can be divided to stationary, carrying, and for the motor cars, while according to their electric-acoustic properties and electric parameters, it can be divided into four groups of complexity: zero (highest), 1, 2, and 3. They must provide the reception of mono- and stereo signals in ranges LW, MW, SW, USW, and decimeter waves, and receive signal with different modulation types, as well as they have the high reliability and the minimal cost. Practically, they are realized as receivers of the super-heterodyne type with the number of frequency conversions of the input signal, which is defined by the group of complexity, but usually not more than three. The value of lowest intermediate frequency at reception of AM signals is always 465 kHz, and at reception of FM signals, it is 10.7 MHz. If it is necessary to receive signal with AM and FM by the single receiver, its radio front-end is divided into two separate channels: each channel for signal reception with its own type of modulation.

The subscriber need in a choice of interesting programs leads to a necessity of retuning in the range of receiving frequencies.

Modern tendencies of improvement of radio broadcasting receivers are expressed in improvement of the main technical indices, in refuse from mechanical and electromechanical units and particles, and in application of frequency synthesizers, micro-processors, and digital control systems.

Improvement of technical indices of receivers is performed in direction of the modern element base utilization (transistors and ICs on its base, which have the lesser noise, the greater gain). We can achieve the better filtering of collateral channels of reception and conversion by application of piezo-electric devices: piezo-ceramic, monolithic quartz and SAW filters having small dimensions, and low cost and mass.

Elimination or replacement of mechanical blocks (range switchers, tuning blocks, etc.) essentially increases a reliability, improves mass-dimension indices, and decreases the receiver cost as a whole. Electronic switchers or sensors instead of mechanical ones and varicap matrices instead of the variable capacitors (or variconds) are the necessary components of the car receivers, which are exposed by sharp temperature variations and vibrations and are operating under condition of high humidity. The smooth tuning of the receiver to the required frequency in ranges LF, MF, and HF must be preserved when receiving analog signals, when it is necessary to change the characteristics of the filters of the input circuit and the heterodyne simultaneously and in the same way. To reduce the effect of external interference, which is very intense in these ranges, it is required to narrow the received frequency band up to the mixer input and thus increase the receiver sensitivity. A cardinal solution in improving the reception quality and simplifying the structure of the RF front-end was the use of digital type modulation, frequency synthesizers instead of local oscillator. The use of coherent detection methods and signal processing technology (OFDM) have also played a huge role. Utilization in high-quality receivers (zero group of complexity) of micro-processors allows provision of direct determination (a choice) of the necessary frequency from the keyset, to store in the memory of the set or operating frequencies and information about the current operating mode of the receiver, to control of the operating mode, etc. For this purpose, the multi-functional highly integrated chips (HIC) are developer, for example, KA1508XJ11, the frequency synthesizer; KP1507IE1, RF divider with variable division coefficient; etc.

The structural diagram of the rigorous receiver for AM signals of the third group of complexity and a part of the FM signal receiver is implemented on IC K174XA10 (Fig. 3.64).

The signal from external antenna passes on the tuned by the capacitance variation of the variable capacitor of the input filter (IN) matching with the LO circuit. The adjacent channel suppression by 35 dB at offset on ± 9 kHz at reception of the AM signal converted into the intermediate frequency $f_{IF} = 465$ kHz is performed in the piezo-ceramic filter PF1P-024 with the pass-band of 8...11.5 kHz with following its amplification in the IF section. After detection and low-frequency filtering and amplification in the modulation frequency amplifier (MFA), the received signal is reproduced by a loudspeaker.

Integrated circuit IC K174XA15 provides reception of AM signals in the range of LW and MW. To receive an FM signal, its preliminary processing is used in a high-frequency unit, which is implemented, for example, on the K174XA15 IC. This block, with the help of external components, provides preliminary filtering and conversion of the received signal to an inter-

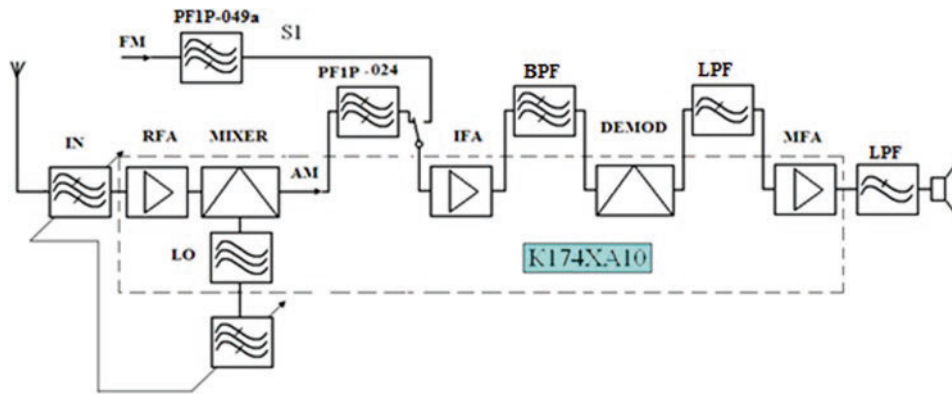


Fig. 3.64 The structural diagram of the rigorous receiver for AM signals of the third group of complexity and a part of the FM signal receiver implemented on IC K174XA10

mediate frequency of $f_{IF} = 10.7$ MHz. The useful signal is extracted by BPF type PF1P-049a with a bandwidth of $\Delta f = 200$ kHz implemented on piezoelectric ceramics. At that, the high-frequency part of the radio front-end of the receiver on the K174XA10, including the local oscillator, is switched off, and the signal is fed directly to the IFA input. After demodulation and following filtering (parameters of the FM filter at reception of FM signal are varied), the signal is amplified and converted by the loud-speaker into the acoustic form.

The integrated circuit K174XA15 represents the multi-functional circuit intended for UHF blocks of receivers of any category of complexity, which has the typical structural diagram (Fig. 3.65).

High indices of IC are caused by application of the mixer-multiplier (Mix) with the deep feedback, which has the high input resistance and great gain. The balanced local oscillator with the buffer amplification stage decreases the influence on it from the input signals and increases the UHF block stability to formation of the spurious channels of conversion. Application of the embedded voltage regulator increases the LO frequency stability at power supply voltage variations. Realization of the receiver with the architecture in Fig. 3.66 on IC K174XA15 with connected discrete components is presented in Fig. 3.66.

The signal from the antenna passes through the tuned single oscillating circuit (SOC) in the RFA1 input via the pin 10, and amplified voltage from the pin 8 is fed through the tuned SOC via pins 3 and 4 in the input of an amplifier-multiplier. The IF signal from multiplier collectors passes to low-frequency filters on RC structures and can be used at pins 13, 14.

The UHF block on the base of IC has the electronic tuning. The variable resistor R1 controls the tuning frequency. Adjustment potentiometers R2...R5 serve for accurate matching of circuits. Main parameters of UHF block are as follows: the intermediate frequency is 10.7 MHz; the pass-band in RF is 1.7 MHz; the image channel suppression is 80 dB; the direct transfer channel suppression is 100 dB; and the noise factor is 6 dB.

3.3.13 The Structural Diagram of the LTE System Receiver

The main fragments of the receiver structural diagram, which are used in the transmission system with the orthogonal frequency channel division, are shown in Fig. 3.67.

The signal generated by the antenna at the input of the ASM antenna switch, in the receiving mode, is fed to the input of a non-tunable preselector consisting of an input circuit and a low-noise amplifier. This structure provides the allocation of the operating frequency band of the system and suppression the image channel.

After down-conversion and allocation of the operator's bandwidth using a matched Nyquist filter at the center (intermediate) frequency, restoring the orthogonality (in frequency) of these responses, the signal arrives sequential/parallel signal transformation. After that, CP is removed from the group of 12 subcarriers that make up the OFDM symbol, which makes it possible to ensure mutual orthogonality of subcarriers over the entire symbol length (orthogonality in time). The CP, which is an effective way to deal with multipath at the receiver input, now becomes an interference in the algorithmic demodulation procedure (FFT).

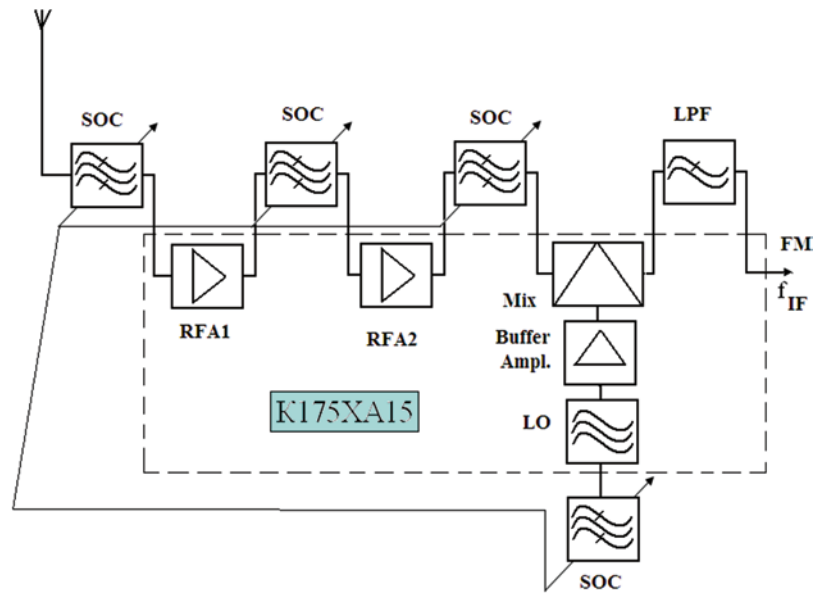


Fig. 3.65 The typical structural diagram of the integrated circuit K174XA15

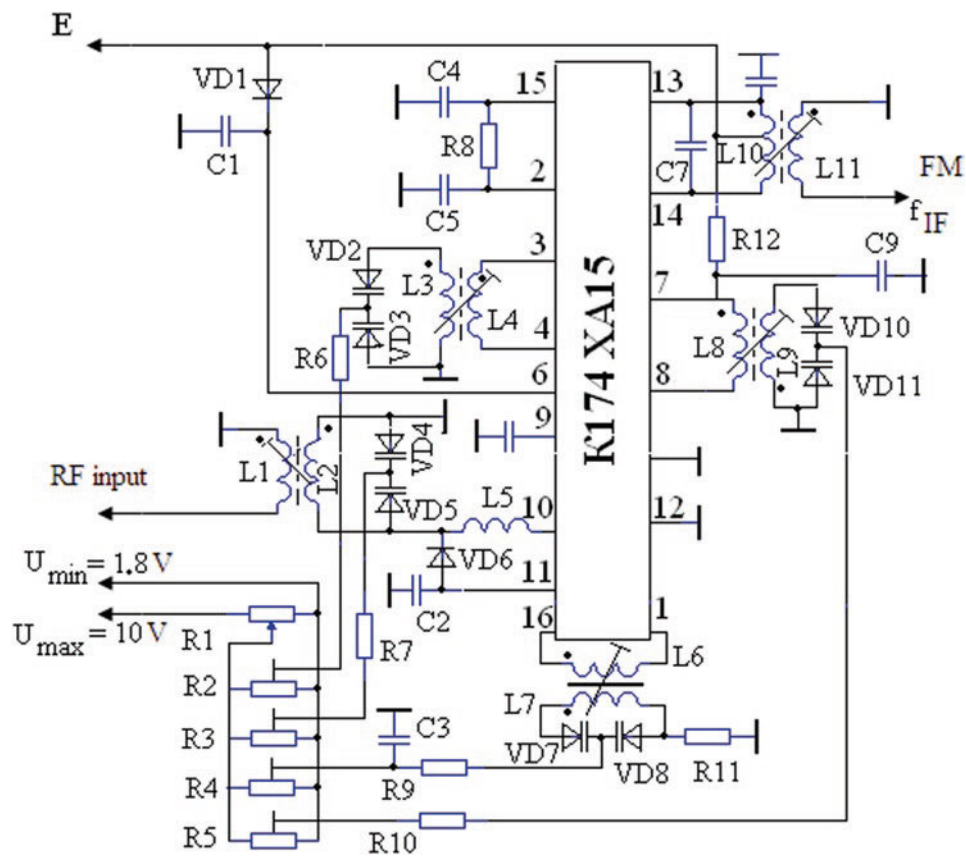


Fig. 3.66 Realization of the receiver with the architecture in Fig. 3.56 on IC K174XA15 with connected discrete components

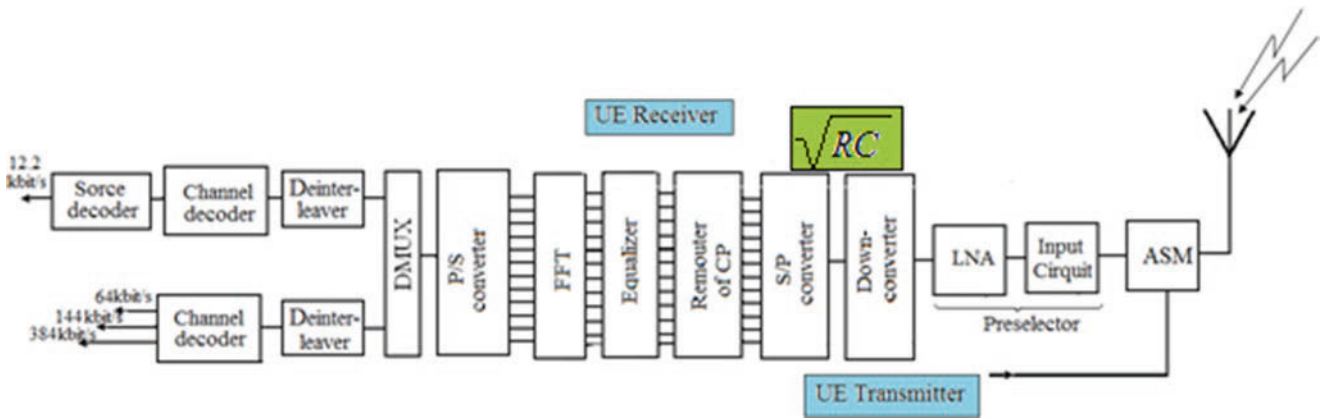


Fig. 3.67 The main fragments of the receiver structural diagram, which are used in the transmission system with the orthogonal frequency channel division

Fulfillment of orthogonality condition for pulses of the modulating digital sequence of each subcarrier permits to present the transmission channel in the form the large number of independent channels with its own complex transfer function. Therefore, for correlation processing of transmitted information, we must equalize the phase shifts and values of gains, which is performed by an equalizer.

Remaining block of samples is processed with the help of the fast Fourier transform, and demodulation is performed (formation from the time sequence of the multi-frequency structure), which has the orthogonality.

In its output, the signal consisting of the sequence of responses, which have the mutual orthogonality, is formed. After parallel/sequential signal transformation, it passes to the input of the Fourier converter or to the group of parallel I/Q demodulators. The complicated circuitry structure performing the radio signal demodulation and the conversion into the pulse sequence, as in the transmitting section, can be replaced by the Fourier transform. The discrete Fourier transform allows formation of the pulse sequence by software means with the deciding device, when the conclusion is made about the reception of logic zero or unit by means of comparison of the receiving value with a threshold. This allows obtaining the bit sequence from the sequence of informational symbols, which then is exposed to deinterleaving and decoding.

The packet structure of the digital information representation begins with the preamble consisting of some number of adjusting OFDM symbols. A part of subcarriers is used as the pilot subcarriers, on which the known informational symbols are transmitted, which are used for corrections of the carrier and synchronization frequency.

When using the narrowband channels, for instance, in GSM standard. The channel bandwidth is 200 kHz and it is impossible to increase instantaneously the transmission rate even at the presence of free channels. Another advantage of the wideband channel is a possibility for the subscriber to choose the channel parameters and the loading plot in accordance with its wishes.

3.3.14 The Structural Diagram of the Wideband System Receiver

3.3.14.1 The Non-coherent Receiver of the UWB System

Modern ultra-wideband technologies attract the greater attention for solution such problems as local systems of data transmission, the detection systems, and biometric sensors up to digital communication systems. To UWB systems, we usually refer systems occupying the range from 3 GHz to 10 GHz or its part. The most difficult for developers is the application of the low-frequency part of the range 3...5 GHz with realization of the detection and secrecy function **DAA** (**Data Access Arrangement**) in commercial purposes. Besides the lowest part, the frequency range above 6 GHz, which has the reduction of the operation radius due to loss increase at radio wave propagation, is also free from licensing. The significant impact of the operation of UWB system near the frequency of 5 GHz are exerted by Wi-Fi, WCDMA, and other systems, which causes a necessity of its suppression to increase the quality indices of the UWB receivers. The strongest interference is caused by the Wi-Fi signal, which requires the application of rejection filters in the preselector.

The simplified architecture of the UWB receiver realized with usage of CMOS technology for the range 6...10 GHz [101] is presented in Fig. 3.68.

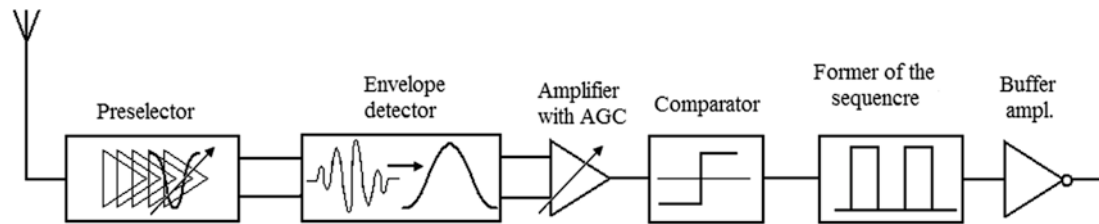


Fig. 3.68 The simplified architecture of the UWB receiver realized with usage of CMOS technology for the range 6...10 GHz

The receiving part is connected via a transmit/receive switch to a slit antenna of a planar type. It is implemented in the form of a hexagonal plume inside a circle, free from metallization. By choosing the size of the rectangle, circle and stab, it is possible to create the value of reflection coefficient $|S_{11}|$ return loss [73] is not less than 15 dB in frequency bands of the DSC, PSC, UMTS, and WLAN system. That is, such an antenna has filtering properties.

The receiver preselector realized as the single chip-set includes the five stages of the low-noise amplifier; one of them contains the active rejection filter with adjustable frequency. The first amplifier stage is performed on the differential structure and covered by the negative feedback to achieve the highest broadbandness. The second amplifier stage with non-symmetric input and symmetric output is realized on the differential stage, which transistors are connected on the cascode circuit (general gate, general source). The load of the second stage should (besides the necessary amplification and the bandwidth) provide the AFC growth in the high-frequency region by insertion of the LR chain in drain circuits of transistors.

To decrease the distortions caused by an action at the input of the UWB receiver of signals of the wideband access Wi-Fi system, we can include at preselector input the rejection filter tuned on the frequency of 5 GHz. In case of insufficient suppression, for example, of frequency band of the system WiFi, an additional filter can be turned on at the preselector, using it as the load of the last stage of the amplifier. The amplifying stages are fulfilled on the integrated technology including the passive components of the rejection filters: the inductance of the planar construction and the capacitor matrix. Variation of the rejection frequency of the circuit is provided by the discrete variation of the matrix capacitance and the bias voltage of amplifier transistors. Decrease of the stage gain in the frequency region near 5 GHz at connection of such a filter has the value of 10÷20 dB compared to its absence.

The envelope detector of the input signal realized as the autodyne (generating) mixer converts the input impact spectrum into the low-frequency region, and the passive RC filter at its output forms the signal repeating the variation law of its amplitude. The amplitude of the envelope voltage for achievement of the required receiver sensitivity should be a value of about 10 mV in amplitude.

The signal level at the comparator input (the device comparing the envelope value in the middle of the interval of radio pulse existence with the threshold value) is provided by an amplifier spanned the deep negative feedback, which creates the constant output level. The comparator forms in its output the sequence of narrow pulses with the amplitude enough for sequence former operation. The buffer inverse amplifier provides the input signal, which level and polarity are necessary for the digital section of the receiver.

The main receiver characteristic is the sensitivity, and it is measured at the frequency of error bit appearance (BER) of 10^{-3} , and it is $-70...-75$ dBm depending on the operating frequency.

Dimensions of the receiver chip are $1.25 \text{ mm} \times 2.6 \text{ mm}$ at realization of UWB receiver units on the unified substrate and application of CMOS transistors with the gate length of $0.18 \mu\text{m}$.

3.3.14.2 The Receiver of UWB System on Differential Structures

The necessity of transmission rate increase and decrease of a cost and dimensions of wireless communication system units led to appearance of six-port receivers of direct conversion. Receivers of such a configuration implemented on a single chip provide signal processing in ultra-wideband systems (6...9 GHz) at the rate of 2 Gbit/s. Utilization of the differential structure at realization of such a receiver allows increasing of construction compactness and loss decreasing compared to application of units with single-ended realization. The double-sided assembling with usage of parallel strip lines on the upper substrate side with the similar topology on the lower part (the virtual ground) provides the level equality and signal antiphase condition in each branch of the differential structure. To achieve the symmetry on in-phase and quadrature inputs (outputs) of the double-sided structure, if necessary, the grounded metal plate is placed in the substrate.

Completely, the differential structure of such a receiver is shown, in Fig. 3.69.

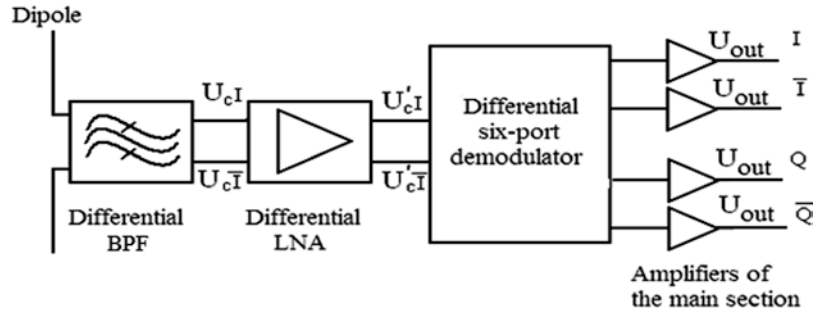


Fig. 3.69 The differential structure of the UWB receiver

Main functional blocks of the structural diagram are an antenna of the dipole type that provides the reception of the differential signal passing to the low-noise amplifier, which then is converted into the low frequency region with the following amplification of differential signals. The balanced antenna of the dipole type is implemented on the printed circuit board and is embedded constructively in the structure of the RF differential filter. Then the signal amplified in the differential low-noise amplifier passes in the input of the differential six-port demodulator. Application of differential circuits only at realization of separate stages allows elimination of matching and symmetric transformers that increases the resulting effectiveness and decreases losses and the receiver area occupied on the chip. Each transition from the non-symmetric circuit to symmetric one (and vice versa), which is typical for the structure using the single-ended amplifiers and balanced frequency converters as well as quadratic circuits, requires the application of transformers. Transformers having different construction (discrete, strip-type, quasi-lumped) cause a necessity of the add-on assembling at their connection and the increase of an area occupied of such components. Transformer absence and utilization of the unified integral technology at realization of the receiver radio front-end also decreases its cost at achievement of higher broadbandness of the device.

3.3.15 Navigational Consumer Receiver

Signals and frequency ranges. The intensive development of near-earth space and the placement of onboard stations for the creation of system for various purposes required the International Telecommunication Union (ITU) to empower global navigation systems (GNSS) with special status approved by the international community. For this purpose, for existing and prospective GNSS, several sub-band are allocated for users of navigation services in the L-band, and for the Galileo sub-system, there is also an additional sub-band in the C-band (Table 1.2). The operating frequency bands of all GNSS (GLONASS, GPS, GALILEO) are located rather compact, and when using the code channel division, spectra of different systems may even partially overlap. This allows the reception of all navigation system signals by the single multi-system receiver, and the large number of channels with combination of broadbandness of used signals guarantees the high-noise immunity of systems and enables to take into consideration an influence of the ionosphere errors on the accuracy of position determination.

In *GLONASS system*, the frequency method of signal division from different navigational space apparatuses (NSA) is adopted, and since even at full structure of the orbit group (24 NSAs) a number of visible NSAs cannot be more than 12, each of them can be conferred with the one from 12 marked frequencies (frequencies having the conditional serial number). At present, the marked frequency ranges are L1: 1598.0625...1605.375 MHz and L2: 1242.9375...1248.625 MHz. The simultaneous signal reception of the signal of the one marked frequency is possible only on-board of space apparatus, which orbit attitude allows observation of the antipodal NSA (Fig. 1.19). At that, we have a possibility of signal distinguish from such NSA caused by different signs of frequency increments due to Doppler shift. Spatial signal selection due to application of directional antennas allows decrease of the antipodal signal influence. At NSA transmitter power in the L1 range of 64 W and the transmitting antenna gain of 10...12 dB, the signal power received by the isotropic linearly polarized antenna with the gain of 3 dB is not less than 161 dBW. For the L2-range, at transmitter power of 40 W and identical parameters of transmitting and receiving antennas, the input signal power in the worst case (NSA elevation is less than 5°) is -164 dBW.

To fulfill navigational measurements, the simultaneous reception of not less than 4 NSA signals is necessary, which creates the inter-system interference from the point of view of optimal signal processing of each NSA. To reduce the influence of adjacent (in spectrum) NSAs, the marked frequency offset and the signal spectrum bandwidth should be chosen in such a manner that the side lobes of adjacent sources could fall into the operating frequency band of each channel. Their level is -48 dB with regard to the main lobe.

A spectrum of the navigational signal is defined by the highest clock frequency of the modulating sequence, which is for signals of standard accuracy (SA-code) 1.022 MHz and for signals of high accuracy (HA-code) 10.22 MHz. In NSA transmitters of the GLONASS system, the relative phase (PSK) modulation with the shift by 180° between two modulating pseudo-random sequences PRS1 and PRS2 is used. Signal orthogonality is provided by application of quadratic demodulators, which insert the phase shift by 90° .

Pseudo-random sequences are formed by adding modulo 2 of three binary signals:

- Range-finding code of standard accuracy representing the sequence of the maximal length (M-sequence) with the clock frequency of 511 kHz and the period of 1 ms.
- Navigation message with the clock frequency 50 Hz and duration of 2 s.
- The signal of clock synchronization in the form of a meander with frequency 100 Hz.

Another PRS2, which is coherent to PRS1, is formed in a similar manner but using the clock frequency 5.11 MHz for formation of the digital sequence of range-finding code of high accuracy.

The conception of GNSS GLONASS development anticipates the expansion of transmitted signal list and transfer from the frequency signal division (FDMA) to the code division (CDMA). For this, the utilization of NSA is anticipated, which operates in the L3 range on frequency of 1202.025 MHz with the BPSK modulation. Orthogonal modulating signals are realized as truncated digital Kasami sequences with the clock frequency of 10.94 MHz.

In the future, the Russian Federation plans to move into operation in the L1 and L2 bands, choosing instead of FDMA, technology code division multiple access (CDMA). With the transition to a new channel access technology, pseudo-random square wave BOC (binary offset code) sequences will be used as the modulation signal. The difference of BOC from the usual PRS consists in the fact that the BOC-signal represents the segment of the meander signal, which includes the some constant number of periods. At the length of the meander modulating oscillation segment, which is the even number of meander periods, the navigation signal spectrum widens and its auto-correlation function becomes narrower, which provides the higher potential accuracy of delay measurement. Spectrum spreading (expansion) decreases the mutual error level produced by the simultaneous impact of navigational systems, which use the traditional and new types of modulating signals. Main criteria of the meander signal characteristic are the following parameters:

$k_1 = f_m/f_{cl}$ is the ratio of meander clock frequency to the clock PRS frequency.

$k_2 = f_{cl}/f_{ref}$ is the ratio of the clock PRS frequency to some reference frequency ($f_{ref} = 1.023$ MHz is chosen as the reference frequency in GALILEO and GPS navigational systems).

$N_m = \tau_{cl}/\tau_m$ is the number of meander pulses falling in duration of the single symbol of PRS of range-finding code.

φ_m is the initial meander phase with regard to PRS.

On the base of mentioned indices, the radio signals with sine and cosine meander modulation are formed, which are the base for formation of more complicate modulating signals.

In navigational *GPS system*, the code method of signal division is used, where these signals have the width of 20.46 MHz on the carrier frequencies L1 = 1572.42 MHz and L2 = 1227.6 MHz. Carrying frequencies are formed by the single reference oscillator with the nominal frequency of 10.23 MHz and inserted relativistic amendment for the ground-based observer, which is $4.5674 \cdot 10^{-3}$ Hz.

The on-board transmitter power of NSA is about 50 W at the receiver input power of the ground-based equipment equipped by the isotropic linearly polarized antenna with the gain +3 dB and the NSA elevation of 5° is equal $-166 \dots -165$ dBW. Such signal parameters as losses on modulation non-ideality and filtering non-ideality, the power of out-of-band emission, and phase noises are practically the same as parameters of the GLONASS system. The level on inter-system interferences is significantly higher, which is caused by CDMA application, and for signals of the standard accuracy (SA) forming with utilization of the Gray code with the length of 1023 elements, and it will be more by 27 dB than in GLONASS.

Signals on frequencies L1 and L2 are modulated by two orthogonal (shifted by 90°) PRS formed by summing modulo 2 of the range-finding code and the informational sequence, which includes the service information.

Navigational consumer equipment. Destination of the consumer equipment consists in signal reception from NSA, measurement of navigation parameters, extraction of the navigational message (ephemeris, time amendments, etc.), and

solution of tasks of navigation-time determination. According to results of navigational parameters relative to four and more NSAs (pseudo-range, radial pseudo-velocity), we can determine components of the consumer vector of states: three spatial consumer coordinates, three component of its velocity, amendments to the consumer time scale, and its reference oscillator.

The architecture of the modern receiving consumer equipment (Fig. 3.70) includes: the RF block, the processor of primary processing, and the navigational processor.

The micro-strip antenna is usually used as an *antenna*, which has the small mass and overall dimensions, the simplicity in manufacture, and consists of two parallel conducting layers separated by a dielectric. Lower conducting layer is the grounding plate, while upper layer is the antenna radiator. In shape, the radiator can be a rectangle, an ellipsoid, a pentagon, and so on and realizes the circular direction pattern in the horizontal plane, and in the vertical plane, it provides the reception from NSA observing under envelope angles more than $\beta \geq 5^\circ$.

The *antenna module* (A) usually represents the integrated device including (in addition to the antenna) the device for input protection, the LNA, and the pass-band filter. The device for input protection is the nonlinear element preventing an impact on the input stages of signals with the peak power more than 69 kW/m^2 during 10 ms or the continuous signal with power density of 348 W/m^2 . Signal loss in the device of input protection should not be more than 1 dB.

The *RF block* provides the signal reception from NSA, their filtering, amplification, conversion into the intermediate frequency and following signal conversion in the digital form, as well as formation of reference frequencies.

In the *primary processing* block, the signal from NSA converted into the digital form is exposed of correlation processing, and also the search of NSA signals, tracking, and measurement of their navigational parameters, as well as demodulation of the navigational message.

The received navigational message is exposed by decoding in the *navigational processor*, and the consumer vector of states is formed. This block provides the operation mode control of own equipment and other units of the consumer equipment and fulfills the execution of additional service functions.

In the input of the RF block, signals from satellites act, which are in the visibility zone. For calculation of navigational parameters, it is necessary that a number of satellites of the one of GNSS (GLONASS, GPS, GALILEO) would be not less than 4, but theoretically it can achieve 12, and at system combination up to 48. The bandwidth of intermediate frequency filters is chosen taking into account the NSA signal bandwidth at operation in the single or double GNSS. So, for NSA sig-

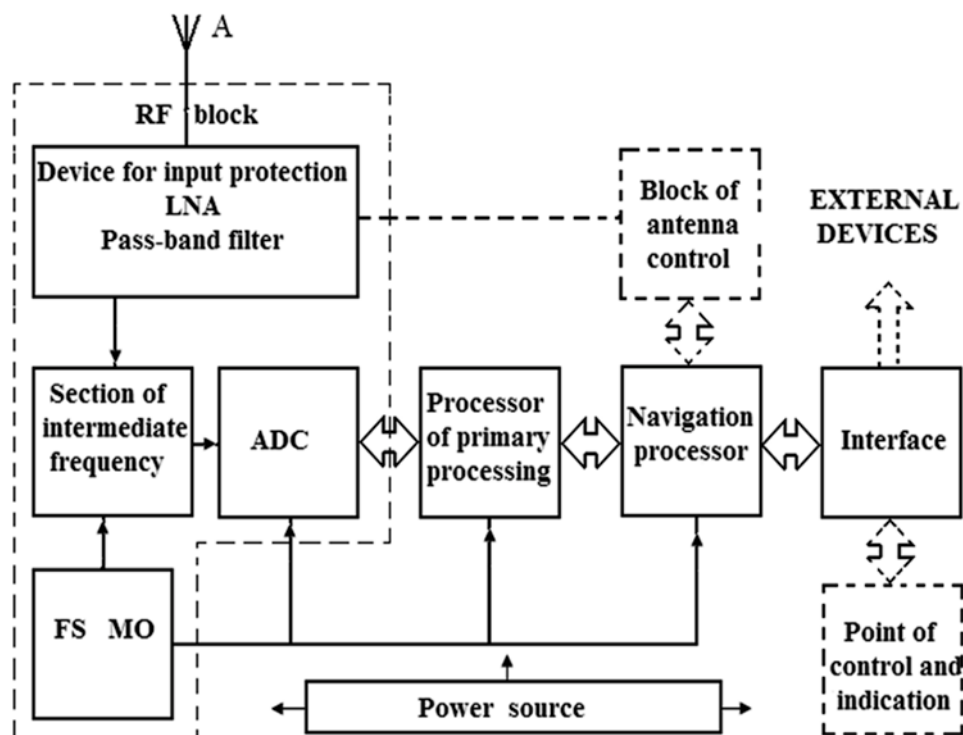


Fig. 3.70 The architecture of the modern receiving consumer navigational equipment

nals in GLONASS, the spectrum width is 14 MHz (according to first zeroes of edging signals); therefore, the width of the IF section is chosen not less than 15 MHz.

For real noise factor created by the radio front-end and reduced to the receiver input, $NF_p = -2.5$ dB, the power spectral density of the noise is -201.5 dBW/Hz. The noise power in the pass-band of 15 MHz is -131 dBW, and the SNR (coefficient of distinguishability) $D_d = 10\lg(P/P_n) = -31$ dB, i.e., in the radio front-end, the useful signal is approximately by 1000 times lower in the own noise power.

ADC connected on the output of the radio front-end forms the digital one-bit or two-bit [102] signal. In the first case, only a sign of the input signal is registered: $U_{out} = \text{sign}(U_{in})$, while at two-bit ADC three threshold levels are used (Fig. 3.71): zero and symmetric, $-L$ and L , which optimal position is defined by the actuation voltage of the ADC least bit and usually is $U_{ADC} \approx 0.1 \dots 0.5 U_{ADC} \approx 0.1 \dots 0.5$ V. This value is close to the value of rms noise deviation σ_n at the binary quantization of the Gaussian noise. If to consider $\sigma_n = 0.25$ V, then at standard load resistance $R_l = 50 \Omega$, the noise power on the ADC input is

$$P_n = \sigma_n^2 / R_n \approx 31 \text{ dBW}.$$

At known noise level on the radio front-end input of -131 dBW, the required gain in the linear section is about 100 dB.

Modern radio navigation receivers are devices combining of analog and digital signal processing. The analog part of the consumer equipment is built exclusively on the super-heterodyne architecture with double or triple frequency conversions. The first intermediate frequency can be chosen in the range 100...200 MHz, while the second IF is 10...40 MHz. At a choice of intermediate frequencies, we must take into consideration requirements to combination interference suppression, the external interference suppression, which influence on the frequency of image and other collateral channels of conversions and other factors. The architecture (Fig. 3.72) realized by Zarlink (Gec Plessey) GP1010 and GP 2010 is the example of the radio front-end construction of the navigation GPS system receiver.

Input signal is exposed to triple frequency conversion with following filtering with the help of external (with central frequency of 175 and 35 MHz) or embedded (4.3 MHz) pass-band filters. LO frequencies (1.4 GHz, 140 MHz and 31 MHz) are formed with the help of frequency divider on the output of the frequency synthesizer spanned by PLL system with the reference oscillator frequency of 10 MHz. Since the frequency grid are realized with application of digital circuits of MO frequency multiplication and division, this leads to phase noise appearance.

The discretization frequency is 6.7 MHz. The output ADX operates in the frequency conversion mode; the frequency of the ADC input signal is about 4.3 MHz; therefore, on the output, we have the signal of difference frequency ≈ 2.4 MHz. The microcircuit is intended for GPS signal reception; however, it can be used for GLONASS as well. For this, it is necessary to

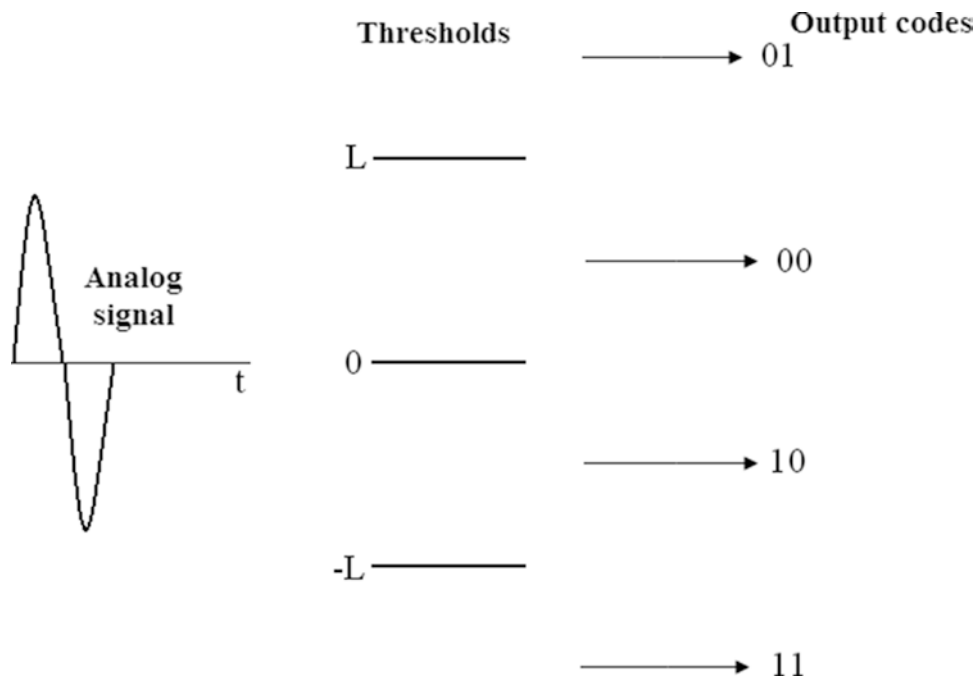


Fig. 3.71 The digital one-bit or two-bit signals at ADC output

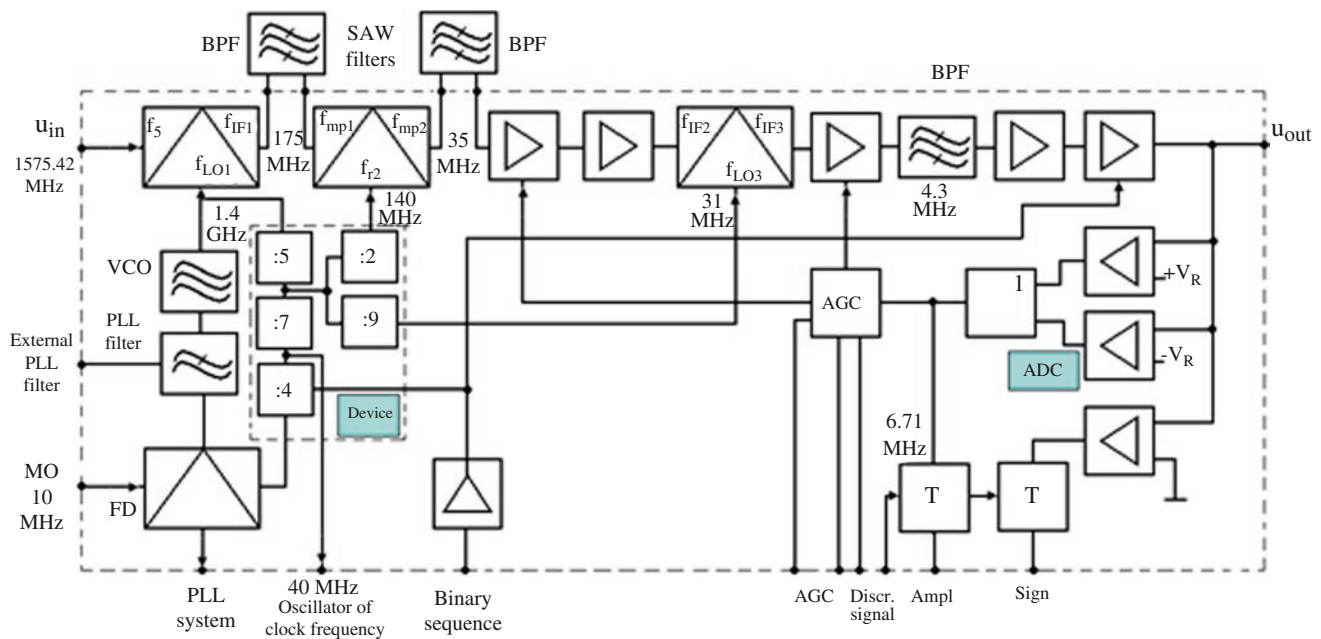


Fig. 3.72 The architecture realized by Zarlink (Gec Plessey) GP1010 and GP 2010 is the example of the radio front-end construction of the navigation GPS system receiver

break a wide GLONASS spectrum into several bands choosing values of oscillator frequencies of the mixer reference voltage for each bands of the total spectrum in such a manner that all marked frequencies would present in signal spectra, which are passed through the output filter. At that, measurements of the pseudo-range and pseudo-phase are performed with great error, which is explained by differences of phase distortions in the radio signal, received from different NSAs (of different frequency sub-ranges).

The structure permitting to provide processing of received signals of different navigational systems in separate radio front-ends (Fig. 3.69) is the alternative of above-described structure (Fig. 3.73).

The signal received from NSA of GLONASS and GPS systems by the combined antenna occupies the frequency range 1570.0–1616 MHz (the range L1, Fig. 3.74).

After amplification in the low-noise amplifier (LNA) and extraction of the operating band by PBF, the signal passes on the Mix1, which converts the signal of the navigation system GPS into the frequency 39 MHz, and the signal of GLONASS in the range of 71–85 MHz. Filters on the output of Mix1 provide the separation of radio front-ends of GPS and GLONASS, and the filter on the input of Mix3 with the cutoff frequency of 90 MHz provides the suppression of the collateral channel of the image frequency.

The GPS signal with the frequency of 39 MHz entering in the input of Mix2 is converted (after multiplication with the VCO signal with the frequency of 34.98 MHz) into the intermediate frequency of 4 MHz, which is extracted by LPF. In ADC, the signal discretization with the frequency of 34.98 MHz is performed with following processing in the correlator.

The channel structure of the radio front-end at signal processing of the GLONASS system includes additional Mix3, which simplifies the struggle with the image channel, as the LPF in its input. The spectrum of converted signal on the output of Mix3 converted in the frequency band of 50.4..0.36.4 MHz passes in the input of Mix4, where it multiplies by the MO voltage with the frequency 34.98 MHz. After extraction of the useful signal band by LPF occupying the band of 15.5...1.5 MHz (inverted spectrum), the signal passes to ADC. Here the signal is exposed to discretization with the frequency of 34.98 MHz, which essentially exceeds the requirements to the discretization frequency choice according to the Kotelnikov theorem. The navigational signal exposed to discretization passes to the device performing the correlation processing and forming the information about object location.

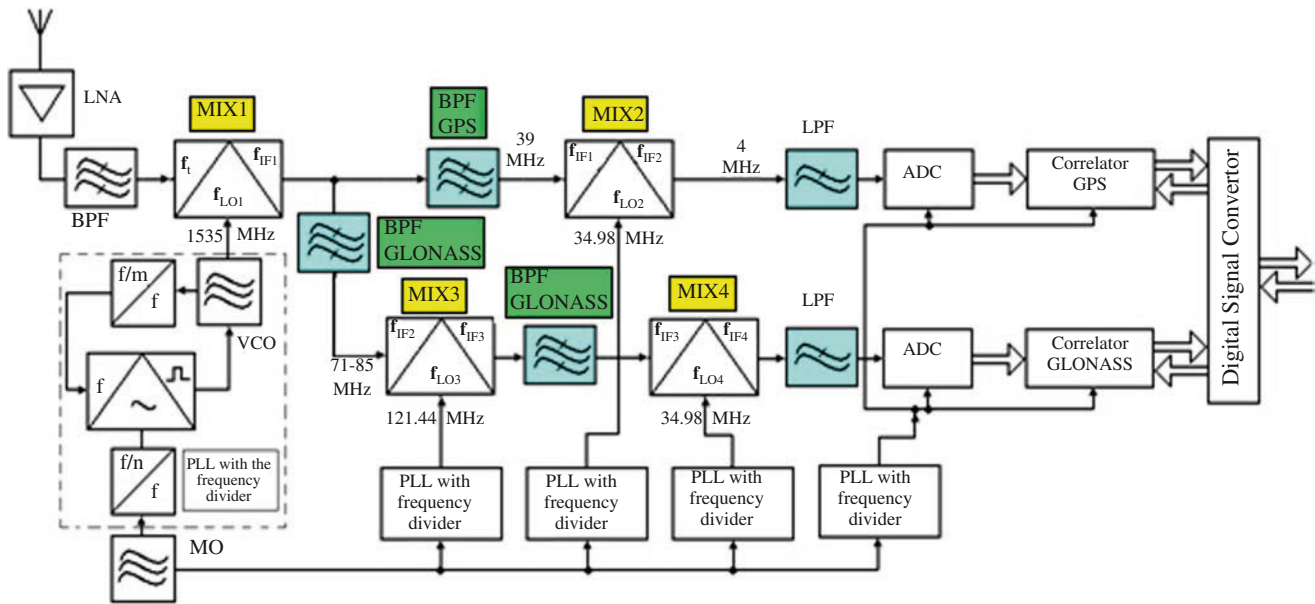


Fig. 3.73 The structure permitting to provide processing of received signals of different navigational systems in separate radio front-ends

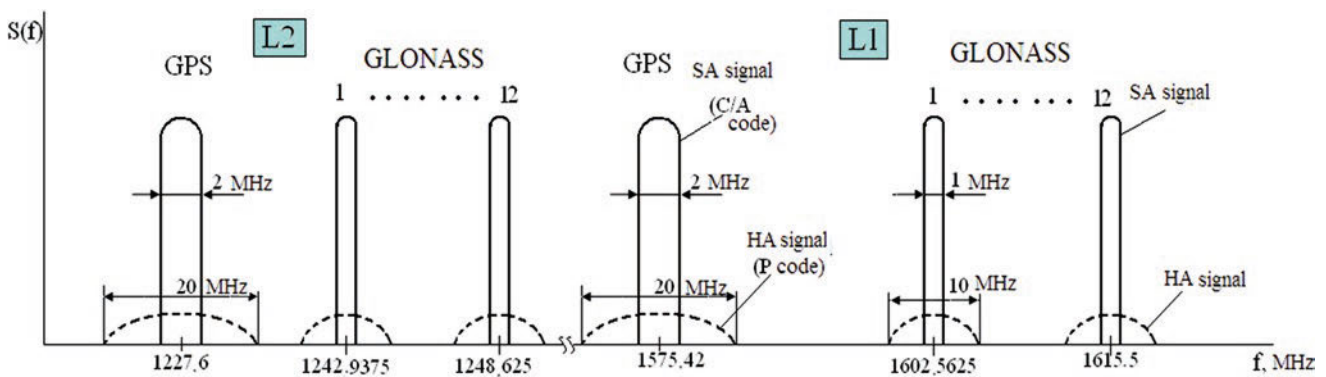


Fig. 3.74 The signal received from NSA of GLONASS and GPS systems by the combined antenna occupies the frequency range 1570.0–1616 MHz

3.3.16 The Receiver for Biomedical Applications

The necessity of long-term and continuous connection to the base station of the wireless radio communication system located on the several meters distance is the feature of devices, which are used in the systems of health monitoring. Besides very low consuming power, these devices, which can be implanted or carried, must be extremely simple and support the operation in the standard excluding the interference with other devices in the hospital or at home and should provide exploitation during long time without replacement and, as a result, the low cost of medical service.

We know methods of construction of medical destination receivers, which have the ultra-low consuming power, according to the ultra-regenerative structure [103], which provides reception of signals on the single frequency. The comparably low proper selectivity and impossibility to arrange operation in several channels are restrictions to utilization of such architecture. Application of the direct conversion architecture is limited by the relative narrowband channel property (100–300 kHz), which results in significant influence of flicker noises and the appearance of bias voltage. The architecture [104] of the super-heterodyne receiver (Fig. 3.75) with low intermediate frequency ($f_{IF} = 400$ kHz) allows provision with higher selectivity, to decrease the consuming power and to solve problems arising in receivers of direct conversion (flicker-noises, the spurious bias voltage).

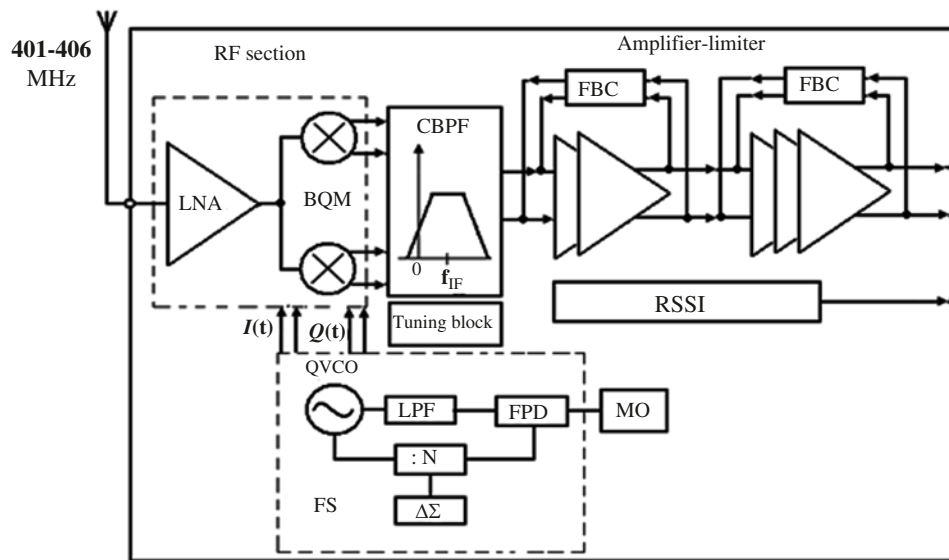


Fig. 3.75 The architecture of the super-heterodyne receiver with low intermediate frequency for biomedical applications

The low-noise amplifier (LNA) and balanced quadrature mixers (BQM), the complex band-pass filter (CBPF) of eighth order, and the amplifier with gain control RSSI (**R**ecive **S**ignal **S**trength **I**ndicator) are connected in the receiver input. LNA is realized on the single-ended structure, which permits to decrease an influence of the spurious bias at low signal level on the mixer input, and the applied circuit of power current reuse in all stages allows the reduction of consuming current. In addition, in the IF section, transistors with low temperature potential φ_T are used, which allows its utilization at low-voltage power source.

Application of MOS transistors with low temperature potential allows utilization the 1 V-power sources for realization of capacitors on its base. The central frequency of CBPF can be calibrated with the help of the tuning block (Fig. 3.75), which contains the relaxation oscillator realized with usage the identical CBPF elements, which allows simultaneous elimination of the bias current dependence of separate blocks, as well as their variations at change of the power supply voltage and the operation mode.

A choice of the local oscillator structure is the very important condition to achieve of the low consuming power. Additionally, it is required to maintain the high accuracy of in-phase ($I(t)$) and quadrature ($Q(t)$) signals, which allows effectively suppress the image channel in the complex BPF (CBPF). Among possible solutions of the structure, one can choose the quadratic VCO (QVCO) with connection to mixers without additional buffer stages, which provides the minimal consuming power and conversion of operating band signals of 401...406 MHz onto the fixed intermediate frequency 400 kHz.

The power supply of receiver stages is based on application of the circuit of etalon conductance g_m (Sect. 3.2.4, [73]), included in IC structure and maintaining the constancy of IC block parameters at variation of the power source voltage, modes of active elements, and environment temperature.

The radio frequency section contains LNA with non-symmetric input and output and the balanced mixer with the non-symmetric input. At that, noise characteristics of the amplifier and its linearity are improved due to application of the current-reuse technology compared to the widely known cascode circuit with an inductor in the negative feedback at the same consuming power and the complexity of designing.

In the mixer realized at the circuit of current-reused in transistors, entertainment the source of constant EMF of small value (1V), which provides the small level of the flicker-noise of converter. Relatively small value of the power source current is used in the switching section, while the greater part of DC of the mixer is used to control the mixer slope, which improves its characteristics. The total power consuming in the RF section is less than $370 \mu W$.

The complex band-pass filter (CBPF) is used for suppression of collateral conversion channels. The Gm-C topology realized on the active RC-structural [105] forms the complex filter consuming small powers from the power source. The structural diagram of the Butterworth filter of 8th order is shown in Fig. 3.76.

Since information about the modulation law is defined by changing of the instantaneous frequency of the received signal (but not by an amplitude), this allows utilization of the amplifier-limiter at the output of the complex filter. Such an amplifier

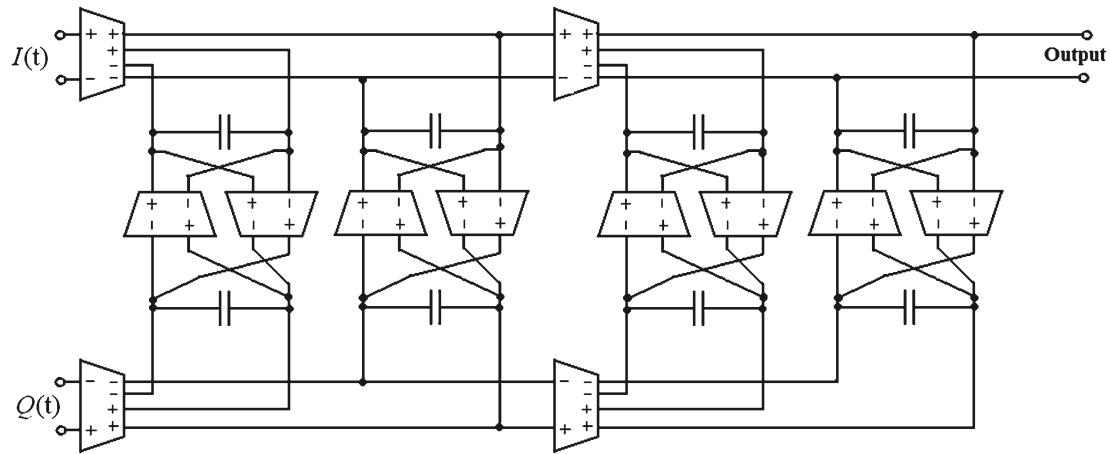


Fig. 3.76 The structural diagram of the Butterworth filter of eighth order

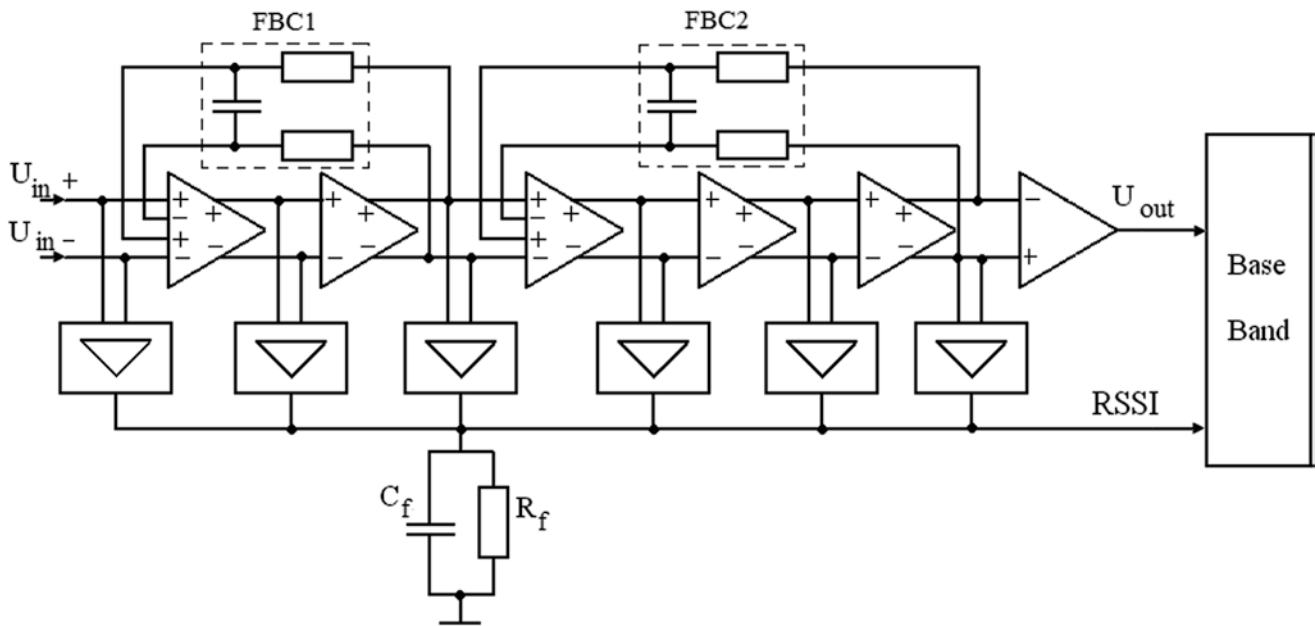


Fig. 3.77 The amplifier-limiter at the output of the complex filter

(Fig. 3.77) with a function of level estimation of the received signal (RSSI, Fig. 3.75) happens to be simpler and consuming the lesser power compared to the conventional circuit of the amplifier-ADC structure.

The amplifier-limiter contains five identical stages differential amplification cascades on the field-effect PMOS transistors connected as the circuit with the mutual source. The output stage is implemented on op-amp with the non-symmetric output, which provides the mode of two-sided limitation (rail-to-rail swing). The RSSI circuit forming the control voltage for IF section gain control contains the full-wave rectifier, which is connected to the output of each stage of the amplifying cascade with following filtering (by $R_f C_f$) of the rectified voltage for extraction of DC component applied to the base-band section.

To prevent a transfer in the saturation mode of differential stages due to the arising spurious bias voltage, we can use the feedback circuit on DC for its elimination. Two loops of a feedback covered two input differential stages and three last stages provide deeper feedback and increase the amplifier stability compared to utilization of the unified feedback for compensation of the spurious bias. Low-pass filters, which are included into each feedback loop, provide the more exact correction of the bias voltage.

The frequency synthesizer (Fig. 3.75) includes the master oscillator stabilized by the crystal resonator, the quadrature VCO spanned by the feedback loop, which includes the phase detector and the frequency divider by N . The bandwidth of PLL loop is 5 kHz, which is the main reason of phase noises and defines the operation of the delta-sigma modulator ($\Delta\Sigma$) and the accuracy of LO frequency values. The applied balance QVCO circuit allows effective struggle with the image channel providing its suppression up to 32 dB under condition of symmetry of lines, connecting the power to the mixer input from QVCO and identity of load connected to them. The value of the spectral density of phase noises at offset of 100 kHz is -106 dBc/Hz at total consuming power 1.2 mW and for 1 V-power source of the receiver. The noise factor of the receiver reduced to its input is 14 dB at radio front-end gain of 82 dB. The area occupied by the receiver is 1.8 mm^2 and allows its utilization in the multi-channel mode for medical monitoring on wireless technology, which is realized for implantation or as the carrying device.

3.3.17 Differential Quadratic Receiver on Mobile Systems of Detection

The differential quadratic receiver with single-frequency conversion. The interest to receiver construction with application of the direct conversion architecture is caused by the less cost of its implementation, a possibility to perform the RF front-end units, and the digital section on unified technology. Nevertheless, such a phenomenon as the local oscillator signal conversion, which is leaked in the mixer input and leading to appearance of non-controlling bias voltage on the mixer signal input, is peculiar to such architecture. The bias voltage appearance is caused by insufficient decoupling between the signal and LO ports of the mixer due to the spurious coupling capacitance and/or through a substrate at frequency equality of the signal and LO. Owing to multiplication of the LO power leaked in the signal input and the LO signal passing in the LO input of the mixer, the component of zero frequency arises, which is identical to the bias voltage. Existence of such a voltage leads not only to gain limitation in the analog circuits of the informational section but also to SNR decrease. Reduction of this problem sharpness can be achieved by application of mixers on LO sub-harmonics [106], when, for instance, the LO frequency is equal to a half of input signal frequency.

The structural diagram of the direct conversion receiver for the non-licensed range of 24 GHz intended for wideband systems such as the mobile detection systems with the phased antenna arrays is shown in Fig. 3.78.

Such receivers is usually performed on the hybrid technology due to a necessity of formation [106] of the quadratic (I/Q) and differential signals. Nevertheless, advantages in the modern integral technology permit the transceiver realization together with hybrid structures providing the phase shifts of $90^\circ/180^\circ$ on the single chip [107]. The receiving part of the transceiver includes the low-noise amplifier (LNA), two mixers on sub-harmonics (SHMix), three quadrature couples (QC) and two intermediate frequency paths.

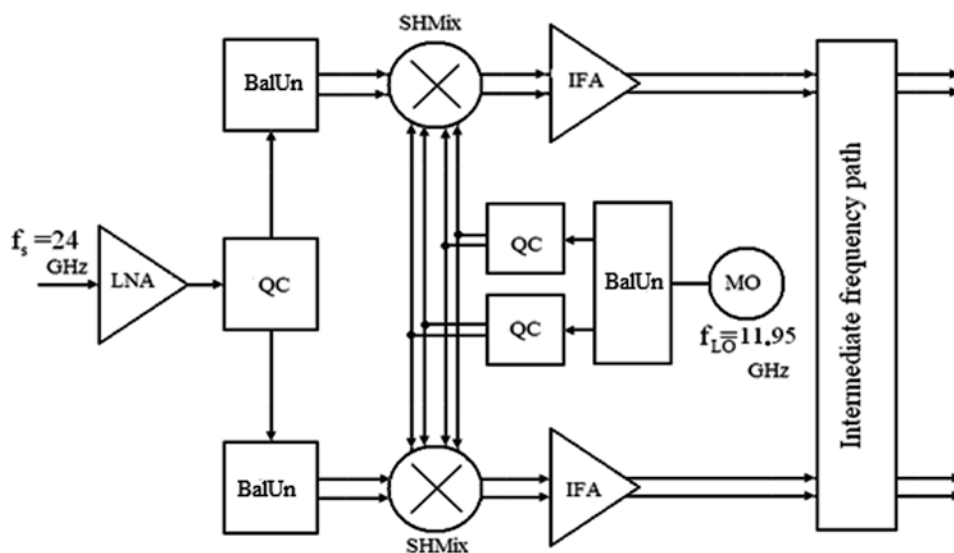


Fig. 3.78 The structural diagram of the direct conversion receiver for the non-licensed range of 24 GHz intended for wideband mobile detection systems with the phased antenna arrays

(directional couplers), three symmetric (matching) transformers MT (**BalUn**) with medium point permitting to connect consequently the single-ended circuits with non-symmetric input (output) and push-pull circuits (with symmetric input-output), and two IFAs. To obtain in the range of 12 GHz the signals of equal powers and having phase shifts by 90° ($90^\circ/180^\circ/270^\circ/360^\circ$), the signal from the non-symmetric output of the master oscillator (MO) with the frequency of 11.95 GHz is separated using the symmetric BalUn and the QC implemented on the single chip.

Supposed variant of receiver construction intended for operation in the non-licensed frequency range of 10.6–31.6 GHz can be implemented with the receiving signal conversion onto the zero intermediate frequency (the receiver of direct conversion) or with conversion to the “low” intermediate frequency (100 MHz). In the last case, the differential IFA is included in the input of differential mixer on sub-harmonics.

A refuse from application of polyphase filters for this purpose is explained by the essential signal damping in the pass-band and by increased level of phase noises caused by resistors, which are included in the RC-polyphase filters (Fig. 3.77). We should also note that the simple phase splitter of the LO sub-harmonic signal can be used only on the differential output of the mixer.

Application in the radio front-end of quadratic couplers allows avoidance a necessity to form radio signals with the phase shift of 45° , as it is required in receivers implemented with application of the conversion on the main harmonic. The LNA connected with the receiver input with operating frequency of 24 GHz splits the input signals (with the help of QC) into two quadratic signals equaled in power, which then are converted into two equaled in power differential I/Q signals by the symmetric transformer connected in the input of the sub-harmonic mixer (SHMix). These differential I/Q signals acting in the input of the balance mixer are multiplied with the MO sub-harmonic providing the direct conversion. The main component of the output signal, at conversion into the low frequency, can be further amplified in IFA, and the out-of-band can be effectively suppressed by the pass-band filters.

Checking Questions

1. What are differences of the structural diagrams of direct conversion receivers and receivers of super-heterodyne type?
2. What are differences of the tunable input circuits in the super-heterodyne receivers and receivers of direct conversion?
3. What are features of the structural diagram of the receiver with digital intermediate frequency?
4. Which collateral reception channel does exist in receivers of any architecture?
5. What are the advantages of non-coherent (nonlinear) detectors compared to synchronous ones?
6. Why do the synchronous detectors have got the wide distribution in digital system of radio communication?
7. Why can we use the tunable input circuits for analog types of modulation in LF, MF, HF ranges, while in the VHF range the non-tunable circuits are used?
8. How can we provide the constancy of the intermediate frequency f_{IF} at receiver operation in the frequency range and tunable input circuits?
9. How does the reception of the necessary channel (of the transmitter) provide at non-tunable preselector in the receiver of super-heterodyne type?
10. Which distortions are related to features of radio signal propagation and are arisen in WWAN and WLAN network?
11. What is the reason of appearance of additional collateral reception channels in super-heterodyne receivers? List them.
12. In which part of the radio receiver front-end, which is realized on the super-heterodyne architecture, the suppression of additional collateral reception channel is provided?
13. What is the reason of utilization sometimes of input signal conversion to the low intermediate frequency?
14. What is the principal difference of the interface in the physical channel of networks of GSM standard and the later generations?
15. In which cases can we used the double and even triple frequency conversion of the input signal in receiver of the super-heterodyne type?
16. Calculate the image channel frequency $f_{im, ch}$ for following conditions: the frequency of the receiving signal $f_s = 1000$ MHz, the intermediate frequency $f_{IF} = 100$ MHz ($f_s < f_{LO}$). Which is the value of image channel frequency $f_{im, ch}$ if the receiving signal frequency $f_s = 1000$ MHz at lower tuning of the local oscillator ($f_{LO} < f_s$)?
17. Calculate the value of image channel frequency $f_{im, ch, 2}$ on the second intermediate frequency in the receiver of the super-heterodyne type with the double-frequency conversion for the following conditions: the signal frequency $f_s = 1000$ MHz, the frequency of the first local oscillator $f_{LO1} = 700$ MHz, and the second intermediate frequency $f_{LO2} = 100$ MHz. In which stages is the image channel $f_{im, ch, 2}$ suppression performed?
18. The impact of concentrated interferences in the mixer input (see Sect. 3.1.1.2) of the super-heterodyne receiver with the single-frequency converter of the input signal, which spaced from the frequency f_{LO} and its harmonics by quantity of f_{IF} ,

i.e., $f_{LO} \pm f_{IF}$, $2f_{LO} \pm f_{IF}$, $3f_{LO} \pm f_{IF}$, etc. (3.4), leads to formation of the component with the frequency f_{IF} in its output. Under condition $f_s < f_{LO}$, the component of f_{IF} , which is produced by f_s , will be useful, while all others, by the collateral reception channel (Fig. 3.5), i.e., by the image channel, which the most influencing on the reception quality, the interference will be in the mixer input with the frequency $f_{LO} + f_{IF}$. For conditions mentioned in the specifications: $f_s = 1000$ MHz, the LO frequency will be $f_{LO} = f_s + f_{IF} = (1000 + 100) \cdot 10^6 = 1100$ MHz.

If the useful signal frequency will be more than the LO frequency ($f_s > f_{LO}$), the image channel will be on the frequency $f_{LO} - f_{IF} = (1000 - 100) \cdot 10^6 = 900$ MHz.

Appendix 3.1 Calculate the Value of the Image Channel Frequency of Super-Heterodyne Receivers and Method of Channel Aggregation in the LTE-A System

Application 3.1

Calculate the value of the image channel frequency $f_{im, ch, 2}$ on the second intermediate frequency in the receiver of the super-heterodyne type with the double-frequency conversion for the following conditions: the signal frequency $f_{s, min} = 935$ MHz, $f_{s, max} = 960$ MHz. In which stages is the image channel $f_{im, ch, 2}$ suppression performed?

The mentioned range corresponds to the operation range of receiving frequencies for the GSM standard. At application of the super-heterodyne type receiver structure with the double-frequency conversion (Fig. 3.7), the value of the first intermediate frequency can be chosen as $f_{IF1} = 300$ MHz, which is determined by requirements on the image channel suppression on the second intermediate frequency and by simplicity of technical realization of the F1 filter (Fig. 3.7).

The average frequency of the reception range is

$$f_{avr, freq} = \frac{960 + 935}{2} = 947.5 \text{ MHz};$$

$$f_{LO1} = f_s + f_{IF1} = (947.5 + 300) \cdot 10^6 = 1247.5 \text{ MHz};$$

The second intermediate frequency should be chosen from the condition (3.9):

$$f_{IF2} > \frac{f_{s, max} - f_{s, min}}{2} = 12.5 \text{ MHz} \text{ (we choose the standard value of } f_{IF2} = 70 \text{ MHz).}$$

The frequency of the second local oscillator (3.10) is:

$$f_{LO2} > \frac{f_{LO1} - f_s + f_{IF1}}{2} = \frac{(1247.5 - 947.5 + 300) \cdot 10^6}{2} > 300 \text{ MHz (we choose } f_{LO2} = 300 \text{ MHz).}$$

For calculated and chosen values $f_{LO2} = 300$ MHz, $f_{IF2} = 70$ MHz, the frequency of the useful signal converted to f_{IF2} will be $f_{IF1} = 230$ MHz (at upper LO tuning (Fig. 3.8), and the image channel frequency is $f_{im, ch, 2} = 370$ MHz.

Application 3.2

Calculate which option is preferable to choose to form an aggregated 100 MHz bandwidth, consisting of 20 components each 5 MHz wide (512 Fourier transform points) or 5 components with a 20 MHz bandwidth (2048 Fourier transform points)?

In the first case, the frequency band effectively used in each aggregated component is 4.5 MHz (301×15 kHz), and in the second 18 MHz (1201×15 kHz), which makes it irrational to choose the second option for constructing an aggregated bandwidth with a width of more than 60 MHz. Application 3.1

A compromise can be a variant of building an aggregated band, selecting the number of aggregated components by the number of radio blocks used in each band.

Chapter 4

Noise Immunity of Radio Receivers

In this chapter, we discuss problems of radio receiver noise immunity. Section 4.1 is devoted to intra-system interferences and describes the structure of covering zones of fixed and mobile cellular communication systems. The channel blocking, the intra-channel and inter-channel interferences, and nonlinear distortions are discussed. The multipath effect is considered in mobile and fixed communication systems in detail. Intra-symbol interferences and interferences from near-located transmitter are described. Methods for improvement of the channel noise immunity are considered in Sect. 4.2 for invariant signal reception, at adaptive system structures, with channel diversity. We discuss the adaptive compensator, adaptive antenna systems, adaptive demodulators, and robust algorithms. Section 4.3 is devoted to MIMO technology systems including the mobile communication systems for different important standards. The chapter contains two appendices with examples of numerical calculations.

4.1 Intra-system Interferences

In spite of a great variety of mobile communication systems (MCSs) (Table 1.2), practically all modern MCSs operate in the microwave range and use the digital modulation of the carrier oscillation and the cellular principle of network construction; therefore, distortions, which occur at the organization of the information channel, are rather similar to their reasons and methods of their elimination.

Let us consider distortions arising in the most distributed digital systems of mobile communications.

The signal transmission in the radio communication systems is related to conversion in an antenna of electrical signal created by a transmitter into electromagnetic waves and their propagation in the free space (in the radio channel; Fig. 1.3) and the inverse conversion in the receiving side into the electrical signal. The channel properties of MCS depend on properties of the physical medium of radio waves' propagation and on characteristics of the terminal devices (antennas, parameters of a receiver and a transmitter), as well as on the speed of mobile station motion.

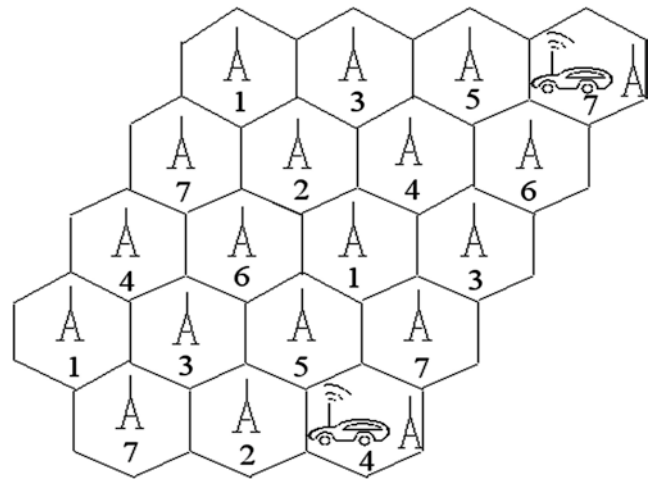
The structure of mobile communication system can be divided into two parts: mobile and fixed. The mobile component provides the interaction of a mobile originated providing bidirectional communication over a radio channel with a base station serving this cell and already belonging to the fixed part of the system. The fixed part of the system performs a network connection between the base station of the calling subscriber and the base station of the called subscriber.

Mobile terminal of the called subscriber, as well as the calling one, can be in motion, within the coverage area of the corresponding base station, transmitting and receiving messages over the radio channel.

The fixed part provides control of digital streams, transforming them into logical channels and optimizing their parths of transmitting. The covering zone of the cellular communication system is divided into sub-zones, which are served by the base station, which is usually located in its centers. The covering zone of the single base stations can be symbolically designated by the regular hexagon, which is usually called a *cell* (Fig. 4.1).

A necessity to provide subscribers by communication in the service zone at relatively small transmitter power in the limited frequency range and for the rather large number of subscribers is a reason of dividing the full zone into cells. A parameter, which takes into account an influence of the mentioned factors, is called the *network capacity* and is characterized by the *maximal number of subscriber radio stations maintained by a system fitted per 1 Hz for 1 km²*. If in the narrowband mobile communication system, for instance, GSM, the frequency channel division (FDMA) is used, then, at the number of mobile stations M and the frequency band allocated to the single station of B Hz, the frequency range of $M \cdot B$ (Hz) is necessary

Fig. 4.1 The covering zone of the single base stations designated by the regular hexagons



for the system. If the full service zone will be divided to sub-zones, we can repeatedly arrange M channels with the bandwidth of B Hz distributing them to different types of cells. The cell group using all available carrier frequencies is called the *cell cluster*. Seven types of cells using repeatedly the same frequencies are presented in Fig. 4.1, which enables a growth of a number of simultaneously supported subscriber radio stations (the capacity of the communication system) by four to six times compared to a system built with the utilization of the single base station. An area of the single cell is much less than the service zone of the whole system; therefore, the power produced by the base station of the single cell is significantly less than the power of the base station, which would cover the whole service zone of the system.

The reuse is defined by the limited distance depending on the permissible level of intra-channel interference. An offset between adjacent channels and distribution of frequency channels within the limits of specified geographic zones are chosen with such an account so as to avoid interferences between adjacent channels. We can distinguish the other types of interferences caused by an influence of the closely located transmitter, channel blocking, inter-symbol interferences, and intermodulation interferences.

The main characteristic, which defines the interference influence in the channel, is the error probability per bit of transmitting information. Among impacts in digital systems, we can distinguish the interference influence mainly caused by intermodulation distortions and by noises:

- The adjacent channel interference (ACI)
- Interference in the main channel (CCI – co-channel interference)
- External interference or the noise in the main channel (ECI)

Influencing impact of interference is defined not only by its absolute power but the following relations acting in appropriate points:

- Current/interference (C/I)
- Current/noise (C/N)

External interference in the frequency band of radio channel can occur in individual cells (clusters), when exposed to sources that generate electromagnetic radiation in the operating frequency band with a discreted or width band spectrum. For example, the ECI interference level in the range of 2.4–2.485 GHz can be -40 to -60 dBm, which is caused by operation of the microwave ovens in this range.

In the well-regulated frequency ranges, the ACI and CCI are the main types of interferences in the cellular communication systems.

4.1.1 Channel Blocking

Owing to the fact that not all channels of the base station of each cell are occupied at the same time, with the growth of the traffic intensity, the temporarily free channel is entered in the operating mode increasing the system's capacity. A network is usually built taking into consideration the expected intensity of its utilization.

At channel distribution, we must take into account the expected intensity of the load (the total time of occupation of links and channels of the phone communication during the time interval). The *load intensity* is the measure of channel utilization expressed in the form of its average relative loading. It is measured in erlangs. For example, the average half-hour channel occupation during 1 h is equivalent to the traffic intensity of 0.5 erl. Since the number of calls is the random quantity, we may introduce a concept of the *maximal load hour* (MLH), i.e., the hour time interval, in which the traffic has the maximal intensity.

It is known that in the typical phone network, only a small part of subscribers use the network service at the same time. At that, an intensity of its utilization can be different depending on the individual subscriber need. This permits to use the idea of trunking consisting in the utilization of the restricted number of channels by essentially the larger number of subscribers. Impossibility to obtain a connection due to temporarily absence of free channels is equivalent to channel blocking for a consumer. This leads to the loss of the connection request (*system with losses*) and to necessity of the new attempt usage to have a connection.

The *grade of service* (GoS) or the *quality of service* (QoS) is a measure of channel access in the system with load concentration in hours of the maximal load. The quality of service is the qualitative measure, which must be taken into account at the development of cellular systems so as to provide the required system capacity and channel distribution between cells. The grade of service is usually expressed in the form of *blocking probability*, i.e., a probability that the consumer, who wishes to settle connection, will have no free channel or the waiting time of the free channel will exceed the settle limit.

In systems with losses, the blocking probability is expressed by the Erlang formula:

$$\text{GoS} = \frac{\frac{A^C}{C!}}{\sum_{k=0}^C \frac{A^k}{k!}}, \quad (4.1)$$

where A is the total load intensity, C is the number of channels in a pull, and k is the number of the current channel.

This formula is right in the assumption that channel requests have the Poisson distribution, the time of channel utilization has the exponential distribution, the number of subscribers is infinite, and the number of available channels is finite. In complex systems – *systems with waiting* – at the absence of a possibility to instantaneous connection due to an absence of the free channel, the subscriber is put on a waiting list. Requests are satisfied as channel release or eliminated from a queue at the achievement of the definite limit of waiting time.

The measure of reducing probability of the channel blocking, which is most often used at the stage of system designing, is the multilevel structure of the cell part in a cluster, where the maximal traffic intensity is expected (stadiums, large trading or entertainment centers, etc.).

The main approach to decrease the blocking probability is the increase of the cell number in a cluster, which allows the creation of the cell of lesser size (the first-level cell or micro-cell) inside of the central cell, which is considered as reference one. Figure 4.2 shows the three-level cellular structure: the central cell is included in the zone of the largest size; then, the zone of medium size is “embedded” in it; inside of which the zone of the unit size is situated.

When the number of channels used in the cell approaches to maximum, within the plan of channel distribution, the service area of the single cell can be decreased. This allows the creation of the larger number of cells with the least physical (spatial) offset and increased recurrence of the same channel frequencies.

Such variation of the cellular network configuration allows service of the much larger number of mobile subscribers within the limits of the large number of small-size cells. Factually, the subscriber cellular network should always contain the cells of different sizes depending on the traffic density in this zone. For the zone of the maximal size, the channel number falling at the zone of the unit size is $0.625 = 10/16$, where 10 is the maximal number of free channels. For the zone of medium size, the number of free channels falling at the zone of unit size is $3.75 = 15/4$, where 15 is the maximal number of free channels in the medium zone. For the zone of unit size, the number of free channels falling at this zone is $25 = 25/1$, where 25 is the maximal number of free channels in the zone of unit size.

In this case, an influence of the base stations of the second stage on the base station of the central cell is essentially relaxed by a choice of carrier frequencies of base stations inside of the micro-cell, even in the case of application of isotropic antennas.

Motion inside a cell can lead to variations of operation conditions of the subscriber radio station, and if the receiving signal level is less than the chosen threshold value, we can arrange the communication channel with another base station and the control of subscriber radio station operation will be transferred to it. The procedure of relay transfer of connection to the new base station, including inside another cell, at the intersection of the subscriber radio station by the boundary between

Fig. 4.2 The three-level cellular structure

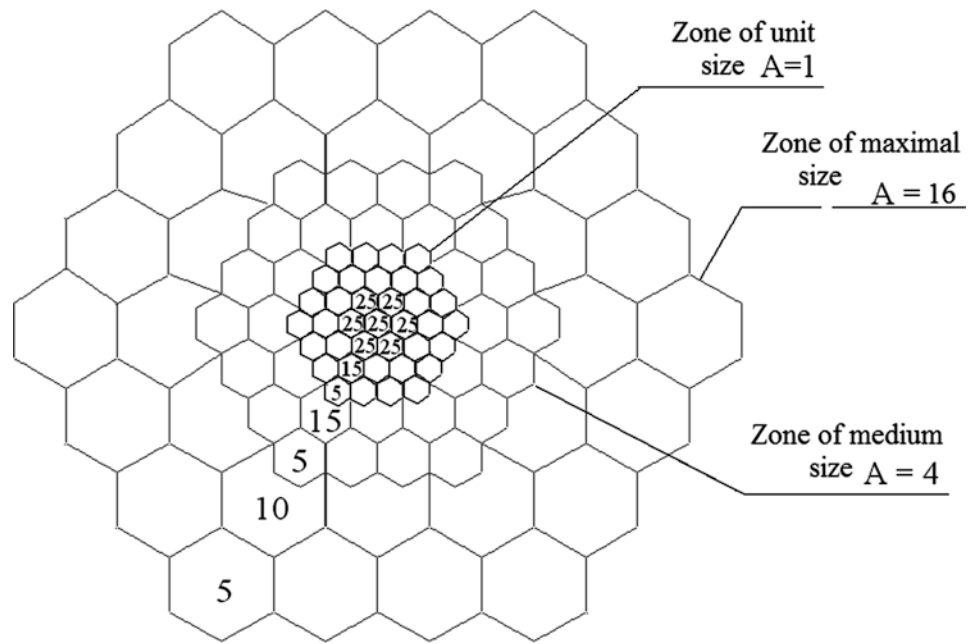
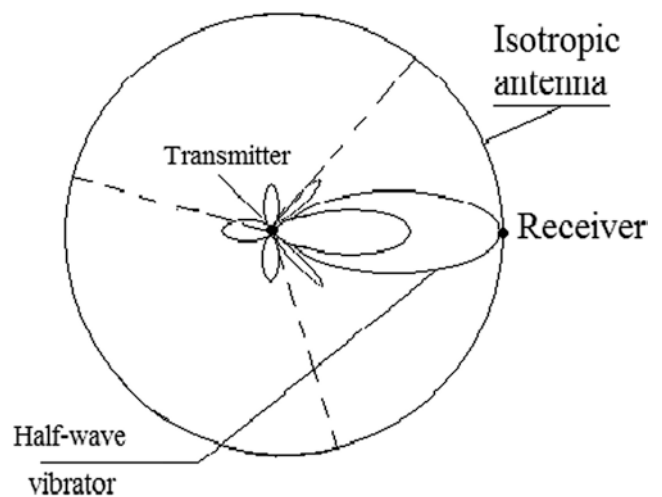


Fig. 4.3 Utilization of sector antennas with a pattern width of 120° instead of the isotropic ones



cells is called the *handover* or *handoff*. The most difficult handover option is the transmission of MS between cells belonging to different BSCs (Base Station Controller) under the control of different MSCs (Mobile Switching Center).

The radio channel is network component most susceptible to external sources of electromagnetic radiation. To assess the current parameters of the channel, the MS regularly, every 0.48 s, measures the level of the 16 strongest carriers, and transmits the results to the BS interacting with it. In turn, the BS measures the level of interference in free slots and sends the measurement results, together with those received from the MS, to the BSC controller. In inter-MSC-handover, the measurements are hierarchically transmitted to the Operations and Maintenance Center (MOC) to optimize the chosen connection route.

The measure, which decreases an impact of the intra-channel interference, is the utilization of sector antennas with a pattern width of 120° instead of the isotropic ones. In this case, each cell is split into three sectors (Fig. 4.3).

As a first approximation, we choose the free space as the propagation medium. Depending on the category of the mobile communication, an antenna can be implemented as:

- The *isotropic*, which has the circular directivity pattern, when the antenna emits a signal with the same power in all directions
- The *half-wave symmetric vibrator*

If the same power is applied to the transmitting antenna (realized as the isotropic antenna) or to the half-wave vibrator, then the electrical field strength, which is produced by a half-wave vibrator in the receiving point, will be 1.64 times (2.15 dB) more than from the undirected antenna, which can be treated as the antenna gain G_{rec} . At that, the pattern has a figure-of-eight shape in the horizontal plane with a width of 180° .

Obviously, the signal power $P_{s, in}$ in the receiver input will be defined by the gain coefficient of the transmitting antenna in the specified direction G_{tr} and by losses on propagation in the free space $L_{fr, sp}$:

$$L_{fr, sp} = \left(\frac{\lambda}{4\pi d} \right)^2, \quad (4.2)$$

where $\lambda = c/f$ is the wavelength in the free space, c is the light speed, f is the transmitted signal frequency, and d is the distance between receiving and transmitting antennas.

4.1.2 Intra-channel Interferences

An interference in the main receiving channel arises at transmission in the single frequency band of two or more independent channels. Such a situation is typical for cellular systems, when the same frequencies can be used many times, which increases a danger of the co-channel interference appearance.

As we see from Fig. 4.1, cells are numerated in such a manner so as the distance between cells with the same number, which defines the carrier (operating) frequency, would be maximal. This allows the minimization of the signal mutual influence, which is transmitted in different cells and in the single frequency channel. Quantitatively, in analog systems, they can be estimated by the ratio of the useful *signal* power in the receiver input P_s to the total power on the *interferences* P_{int} , which can be defined by a quantity of influencing cells K_0 :

$$\frac{P_s}{P_{int}} = \frac{P_s}{\sum_{k=1}^{K_0} P_{int.k}}, \quad (4.3)$$

where $P_{int.k}$ is the average power of interferences produced by the k th cell.

In systems using digital modulation, the multiple access, efficient digital speech coding and channel error correction coding, *signal-to-interference* ratio (4.3) can be reduced to 9 dB [1] compared to 18 dB in analog systems under the same conditions (cell size, base station transmitter power, number of reused frequencies). This shows that it is possible to receive a signal with a lower power (i.e., we obtain a receiver with a higher sensitivity) with an equal S/N ratio at input of the demodulator of digital or analog signals (of the same fidelity).

With the same value of the S/N ratio in digital and analogue networks and the same power of the BS transmitter, the power of useful signal, acting at input of the MS receiver of digital network, increases. This allows digital systems to increase the size of the cell (coverage area of one BS) and increase the distance between cells operating at the single frequency.

Formation of service area, the structure of which consists of hexagonal cells, the area of each of them is determined by power of the BS transmitter and the sensitivity of the MS receiver. Such a structure turns out to be much more effective than using one BS to cover the same area. The use of one (common) BS requires significantly more power than the power of the BS of one cell. The cellular structure of the coverage area has a significantly larger number of simultaneously served subscribers and high efficiency of frequency band allocated to the system and the flexibility of structure (Fig. 4.2), which allows taking into account the forecast of traffic intensity in a given territory. With a high concentration of subscribers, smaller cells are used, and the frequencies allocated by the BS are reused more times, which increases traffic per unit of serviced area. At reduction of the traffic density, the transmitter power of the base station will increase, the cell size becomes larger, and the lesser number of channels belongs to each base station.

Reduction of P_s/P_{int} ratio and a decrease in the total interference of power and the number of used operating frequencies in each cell can be achieved by replacing the omnidirectional antennas on BS with sector antennas (Fig. 4.3).

Usually, the sector served by the single antenna of the base station is 120° . The directional pattern of the antenna is defined in horizontal and vertical planes and can be estimated by the value of the angle between two points of the main lobe, in which the power will be decreased by two times (by 3 dB). Narrowing of the antenna pattern, which is realized on the half-wave vibrators, can be achieved by the growth of their number.

Isotropic antennas, made in the form of a vertical half-wave vibrator, have a circular radiation pattern in horizontal plane. The narrowing of the radiation pattern of such an antenna can be achieved by placing it in front of a screen (reflector).

To separate the receiving and transmitting channels when using one antenna in the BS transceiver, you can use different polarization of electromagnetic waves in the reception band (e.g., horizontal) and the transmission band (vertical), which additionally weaken the transmitter signal at the receiving input by at least 30 dB (Figs. 1.1 and 1.2 [73]).

At the same time, such a construction of the base station antennas for the cellular communication systems is an effective approach for a fight with lumped interferences in mobile communication systems.

The *tuned* isotropic antennas are used in the subscriber terminal of mobile stations. The development stage of the antenna and the transceiver as a whole is emergency of MIMO technology. Structurally, this is reflected by the appearance of several antennas (in BS there are four, and in the MS, usually two), which form, respectively, two or four orthogonal information streams. At the utilization of MIMO technology, several (usually two) receiving-transmitting antennas can be placed on the mobile terminal. At that, the user can orient the own carrying terminal forming polarization with a varying angle.

Evidently, the application of the handover procedure requires continuous monitoring of each subscriber radio station location, which is provided by one of the methods:

- Switched on and being in the waiting mode, the subscriber radio station must periodically update information about its location.
- A group of base stations interrogates the response of the subscriber station knowing that it is situated in the service zone of one of them.

In addition, we must take into account that the execution of the handover procedure occurs at the intersection of the subscriber radio station of a boundary between cells during conversation. For this, the subscriber station in the new cell must receive a free traffic channel from the new BS on the BSC command and only then release the channel in the old cell. The hysteresis rule is used to ensure a seamless connection with a new base station and to avoid multiple transitions to a new base station and back to the old one. This means that the subscriber radio station switches to the new base station if the signal level, which is received from it, exceeds the signal level from the current base station by the specified threshold value.

In conformity with systems of satellite communication, in which intermodulation distortions caused by multipath properties are practically absent, antennas of ground-based gateway stations of the personal satellite communication have the significant gain (40–60 dB) and the antenna aperture of several degrees of the spatial angle.

Assuming there is only one transmitter on the channel, the receiver sensitivity can only be determined by noise. If the external noise power is neglected, the receiver noise factor N , determined by the total noise power, including the thermal noise of the matched antenna and receiver, is calculated by:

$$P_{nT} = kTB_n N,$$

where $k = 1.38 \cdot 10^{-23}$ J/K, T is the absolute temperature, and B_n is the two-sided noise band of the receiver.

Usually, at the calculation of the thermal noise power, one can use the value $T_0 = 290$ K° and then the noise power spectral density is:

$$kT_0 = -174 \text{ dBm / Hz}$$

In technical indices of receivers of mobile communication systems, we usually use the ratio (E_b/N_0) of required energy per bit (E_b) to the noise power spectral density (N_0). In practice, the SNR finds the wide distribution which is the ratio of the average carrier power to the average noise power (usually the value of average rms noise voltage is measured).

Besides the thermal noise of connected antenna, at receiver input, the noises are acting, which are created by the receiver stages (input circuit, amplifiers, converters). They are recalculated to receiver input to compare the power of external noise and that created by the RF front-end (or their noise figures). Such noises created at first approximation by independent sources have the uniform power spectral density and the Gaussian distribution function of the probability density and are called the additive white Gaussian noise.

Receiver parameters under the influence on the intra-channel interferences are calculated in [Appendix](#).

4.1.3 Channel-to-Channel Interferences

This type of interference arises between signals emitted in the same cell (sector) on different carrier frequencies. An example of the appearance of this type distortion in cellular systems is the addition of two channels, that is, in the first approximation, oscillations of carrier frequencies from two subscriber stations, at the input of the base station receiver. In this case, the side lobes of the signal spectrum of the subscriber station, which is geographically closely located to the base station, fall into the band-pass filter, which extracts the adjacent channel possessing a carrier closed in frequency. This frequency is used by the remote subscriber radio station creating the *inter-channel interferences* (ICI).

Moving subscriber radio stations within one cell change their spatial position relative to the common base station, creating power levels at input of its receiver that differ by 40–79 dB (*the effect «near – far»*).

For a fight with inter-channel interferences, we can use:

- Inter-channel frequency spacing, which composes in practice $4B$, where B is the pass-band of the receiving filter
- The synthesis of transmitting and receiving filters, which effectively attenuate the side lobes of transmitting and receiving signals
- Precision regulation of signal power, which is transmitted by base and mobile stations in each channel

The last method has the large practical significance as it allows the creation of approximately similar permissible low power in the base station receiver input from different subscriber stations. At that, it provides at its simultaneous impact in the input of the base station the required ratio P_s/P_{int} (or the probability of error appearance). Obviously, a choice of channel distribution method – the system resources – can have an essential effect on the level of inter-channel interference. There are two methods of channel distribution:

- Fixed distribution, when the distribution of carrying channels is defined from the condition of inter-channel interference minimization
- Dynamic distribution, which provides the channel distribution from the condition of the traffic intensity (maximal capacity of the network)

The rigid fixation of carrier frequencies (channels) in the cellular communication system provides the minimal mutual influence of channels in the prejudice of network capacity. Requirements on the value of inter-channel spacing are saved at the utilization of the operating frequency hopping, which is used for increasing the channel noise immunity against the impact of interferences and multipath property.

In a strategy of flexible channel distribution, the allocation of additional channels is planned in advance taking into consideration the time of the day, the cell location, and the traffic statistics in the previous periods. If the system traffic is measured in the real time scale in the switching center of the mobile communication, this provides a possibility to channel redistribution at any time moment depending on its intensity.

The channel allocation method in specific system must satisfy the trade-off between the quality of service (mainly determined by blocking), the complexity of its implementation in building the network, and the efficiency of using the allocated spectrum.

The receiver parameters under the influence of inter-channel interferences are calculated in [Appendix](#).

4.1.4 Nonlinear Distortions

There are several reasons of nonlinear distortion appearance:

- Necessity to use a mode of the transmitter power amplifier, at which the active element is in the condition, closed to saturation
- Insufficient transmitter power of the mobile station, when the base station receiver transfers into the mode of maximal amplification, which causes an overload of receiver amplification stages of the radio front end at simultaneous impact of signals from other mobile stations
- Leakage of transmitter signal in an input of the transceiver receiving part due to insufficient filtering in the preselector, which causes an overload of input low-noise amplifying stages

Reduction of the nonlinear distortion effect can be achieved by:

- Application of special methods of carrier oscillation modulation (FSK, GMSK, PSK), which form the radio signal with the constant carrier amplitude.
- Introduction of predistortions in the law of modulation signal variation to achieve the constant envelope of the radio signal; this is expedient to apply in the case of the utilization of multi-position AM (for instance, 16QAM) or multi-channel modulation, for example, the OFDM technology.

In receivers of base and mobile stations, the special significance is taken by intermodulation distortions arising in the preselector LNA and in the mixer and caused by the insufficient dynamic range of input stages. If in the operating frequency band of system, concentrated with a high power level appears, when they affect in the AE on the LNA of the preselector or the mixer, this leads to appearance of harmonic components of noise in the output currents. The result of these impacts (often out-of-band concentrated interferences) is an appearance in the operating frequency band of powerful combinational components, which have the total or difference frequencies. Such distortions that occur as a result of the influence of the second harmonic of interference with the frequency f_{int1} and the first harmonic of another interference f_{int2} on the AE input are called third-harmonic distortions and are characteristic of superheterodyne receivers. In direct conversion receivers, the main ones are intermodulation distortions, which are the first harmonics of interference with the frequency f_{int1} and f_{int2} at the input of AE, which create total or difference vibrations that fall into the operating frequency band. Intermodulation distortions of different orders arising in input stages lead to the decrease of the receiver sensitivity and are parameters which must be taken into account at the estimation of real selectivity at the account of useful signal reception on the background of lumped interference action. This is considered in detail in Chap. 5.

4.1.5 The Multipath Influence in Mobile Communication Systems

Channel parameters of transmission systems have the most important significance at the development of any communication system. Channel properties, influence of interference and distortion, and its width (bandwidth of the transmitted signal) determine the maximum possible signal transmission rate for a given quality. At that, characteristics of signal transmission channel depend on the type of the communication system. So, for example, channel properties of cellular, satellite communication systems essentially differ from systems applied for communication organization inside premises.

In networks of fixed communication systems of the SW range, as practically in all networks of mobile communication systems, transmitted signals are also exposed by such impacts as multipath propagation. Additionally, in mobile communication systems, the radio wave scattering on irregularities of the propagation medium leads to this type of phenomena. The summation of the useful signal with its copies having different time delays (different phase shifts) and different intensities at receiver input, which is associated with the properties of communication radio channel, leads to a phenomenon called *fading*. The level of resulting signal even when exposed to, for example, a line-of-sight and on it, a time delayed copy. The total signal can reach values that are many times higher than sensitivity level of the receiver (when the phases of incoming signals coincide), or below the level of the receiver's own noise (with antiphase additions of the incoming influences).

Signal Propagation in the Free Space Characteristics of signals received from the base station depend not only on losses on propagation but also on the locality relief. Since the antenna of the mobile station is located low enough, the locality topography has a significant influence on the level of the received signal both on the mobile station and on the base station. Frequently, especially under conditions of the heavily uneven terrain, an antenna of the mobile station is situated outside of the visibility line, and a signal arrives to a receiver only in the form of reflected and scattered components.

For systems of cellular, personal, and naval mobile communications, as well as land mobile systems of radio communication of general destination for the signal propagation route, the existence of the short enough part of the direct visibility line (LOS – line of sight) and a zone of direct visibility absence (NLOS – non-line of sight) is typical. The signal arrives to a receiver by several ways which differ from each other (Fig. 4.4).

Near the mobile station, each signal component additionally scatters by obstacles (adjacent objects), which causes delays and phase shift appearance. During motion, the mobile station antenna can be outside the direct visibility of the base station antenna, which is typical for the strongly uneven terrain and for dense urban building up and leads to the reception of reflected and scattering components of signals. Depicted channel model illustrates the main phenomena occurring on the short time interval at the transmission of the digital signal. At that, the transmission rate of the digital signal can be up to several hundreds kbit/s at operating frequency up to 2.5 GHz. Signal propagation under these conditions can be compared

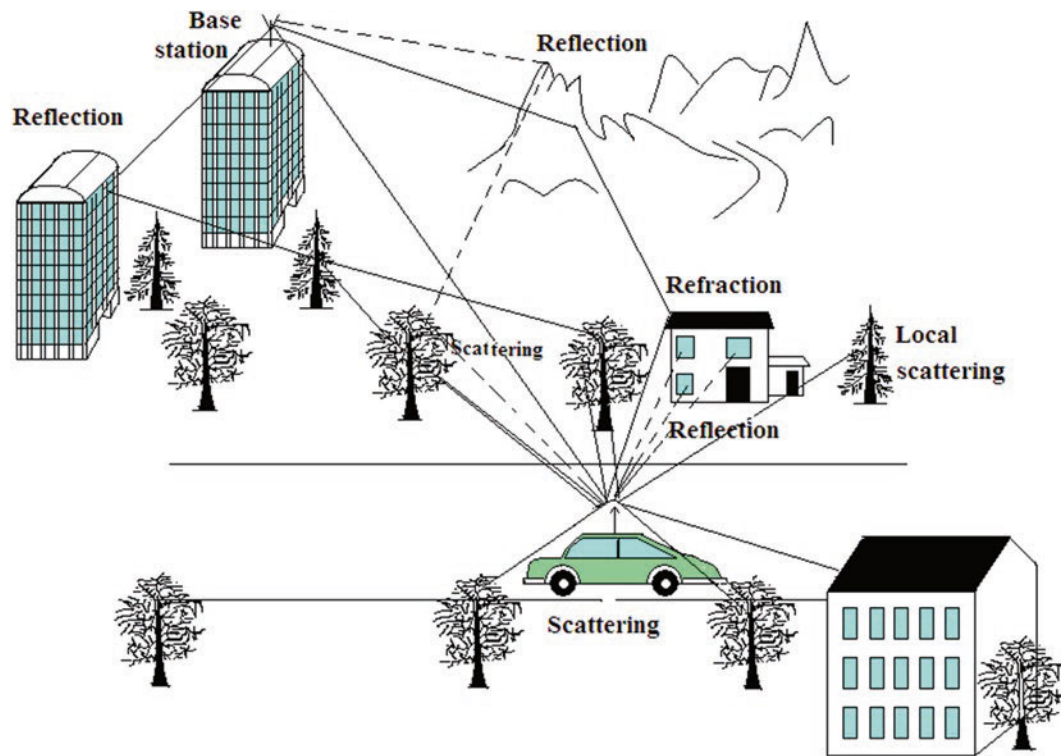


Fig. 4.4 The signal arrives to a receiver by several ways

with the movement of an object along a route with changing parameters, randomly depending on the environment, the number of participants, and their discipline when trying to arrive at a point destination in different ways. For subscriber, who is even near a noise highway, when driving, cars becoming additional dynamic “reflectors,” and the environmental parameters will change significantly under the influence of Doppler scattering, which will change characteristics of radio channel, and therefore degrade the quality of signal reception. In addition to the previously described multipath distortions, the movement of the AT randomly changes the paths of radio signals to the BS, that is, the characteristics of the radio channel. Signals reflected from object rapidly and randomly changing their position in space, relative to the mobile station, receive rapidly changing time delays (phase shifts) and power fluctuations at the input of the BS receiver. This requires the transceiver designer to apply algorithms for the formation of a radio signal and its processing on the receiving side, which provides resistance to such time-frequency distortions.

To simplify the analyses of such distortions caused by several reasons and, first of all, the mobility, can be considered as mutually independent processes arising during the propagation of radio waves and caused by propagation loss, fading caused by multipath propagation, shading or scattering. The fading related to multipath properties is described through the envelope fading that does not include the regular processes caused by modulation. Doppler scattering means a change in spectrum of harmonic carrier with frequency f_0 that has passed through a radio channel with multipath. At the output of the channel, a signal is formed with a discrete spectrum occupying approximately the frequency band between $f_0 - f_D$ and $f_0 + f_D$. This is a consequence of the appearance at the input of the receiver of several beams with different phase shifts. Phase changes cause noise due to random frequency modulation (FM). This allows Doppler scattering to be regarded as nonstationary, discrete in time (or time varying) random phase noise. A rapid change in a position of the MS in space leads to change in the lengths of all signal propagation parts from the BS and, accordingly, to the time delays of the main signal at line of sight (LOS) and its copies (non-LOS) arrive of other ways at receiver input. The receiver fluctuating signal envelop at input of the MS, in which there is often no line of signal, has Rayleigh distribution. The impact at the input of the receiver of signals with a random change in time delays (time scattering), together with the fluctuating amplitude, from time-frequency selective fading. Along with multipath and scattering, impulse interference, interference from other subscribers (within a cell and between cells), also affects the input the MS input signal. In a circuit-switched system, signals from other subscriber stations operating in the same cell, but a different carrier frequency, may occur at input of the MS, which is caused by insufficient filtering of side frequencies. The same effect is observed at the interference of the useful signal with the signal on the same carrying frequency but falling to the receiver input from another cell.

In mobile stations of cellular communication systems, omnidirectional antennas are used, which are close in characteristics to an ideal isotropic antenna. Such antennas were used in the BS of the first stages of the development of mobile communications. An isotropic antenna (ideal device) has a spherical radiation pattern, that is, at all points of spherical surface of radius r , the electromagnetic field induced by the antenna will be the same. The model of the radiator of such antennas is sphere with very small radius (close to a point). The power density ρ [W/m²] emitted by the isotropic antenna and measured at distance r [m] is:

$$\rho = P_{tr} / (4\pi r^2), \quad (4.4)$$

where P_{tr} is the power in the power amplifier output of the transmitter, which passes to an antenna.

In the real antenna, only a part of the power acting in its input is emitted; the other part is dissipated or converted into heat. The effectiveness of electrical power conversion into the electromagnetic field is estimated by the effectiveness of an antenna $\eta_A = P_{tr}/P_{PA}$, where P_{PA} is the electrical signal power in the power amplifier load of the transmitter. The receiving antenna, which possesses the A_{rec} aperture and is located at the r distance from the omnidirectional transmitting antenna, receives the power:

$$P_{rec} = \rho A_{rec} = P_{tr} A_{rec} / (4\pi r^2). \quad (4.5)$$

The effective area of an antenna relates to the A_{rec} aperture (for receiving parabolic antenna with the s diameter) by the relation:

$$A_{rec} = \frac{\pi s^2}{4} \alpha, \quad (4.6)$$

where α is the coefficient of aperture utilization (it is 0.5–0.6) and $\pi s^2/4$ is the real antenna area.

To take into consideration the energy effectiveness of the antenna with the effective aperture A_{eff} , we introduce a concept of the antenna gain:

$$G = 4\pi A_{eff} / \lambda^2, \quad (4.7)$$

where

$$\lambda = c / f. \quad (4.8)$$

Here, $c = 3 \cdot 10^8$ m/s is the speed of electromagnetic oscillation propagation, and f is the carrier frequency at transmission.

The ideal omnidirectional antennas have $G = 1$; from this, $A_{eff} = \lambda^2/4\pi$.

On the base of relations (4.5)–(4.7), we can obtain the formula for the calculation of losses in the free space. For omnidirectional antennas ($G = 1$), we have:

$$\frac{P_{rec}}{P_{tr}} = \left(\frac{\lambda}{4\pi r} \right)^2 = \left(\frac{c}{4\pi r f} \right)^2. \quad (4.9)$$

For antennas, which are spaced by r meter distance in the space and have G_{rec} and G_{tr} gains, the formula for losses takes the form:

$$\frac{P_{rec}}{P_{tr}} = \left(\frac{\lambda}{4\pi r} \right)^2 G_{rec} G_{tr}. \quad (4.10)$$

In logarithmic scale, we have:

$$\begin{aligned} L = 10 \lg \frac{P_{rec}}{P_{tr}} &= 10 \lg G_{rec} + 10 \lg G_{tr} + 10 \lg \left(\frac{\lambda}{4\pi r} \right)^2 = \\ &= 10 \lg G_{rec} + 10 \lg G_{tr} + 20 \lg \left(\frac{c}{4\pi r f} \right), [dB] \end{aligned} \quad (4.11)$$

$$L[dB] = 10 \lg G_{rec} + 10 \lg G_{tr} - 20 \lg f - 20 \lg r + 147,56 dB. \quad (4.12)$$

$$L[dB] = 10 \lg G_{rec} + 10 \lg G_{tr}$$

It follows from (4.4) the intensity per unit area is the unit of measurement of the emitted power. The electrical field strength E created near the receiving antenna, which is implemented as the single- or double-vibrator antenna, induces the following EMF at its output:

$$E_A = E \lambda / \pi. \quad (4.13)$$

The power produced at the receiver input with R_{in} (which is in the general case the complex quantity: Z_{in}) for the matching mode with the antenna resistance R_A ($R_A = R_{in}$) is:

$$P_{in} = \frac{E^2}{4R_{in}} = \frac{(E \lambda / \pi)^2}{4R_{in}} = \frac{E^2 \lambda^2}{4\pi^2 R_{in}}, [W/m^2]. \quad (4.14)$$

The power P_{in} calculated in dB relative to 1 W is:

$$_{in} [dBW] = 10 \lg E^2 [V] + 10 \lg \left(\frac{\lambda}{\pi} \right)^2 [dB] + 10 \lg \frac{1}{4R_{in}} [1/\Omega]. \quad (4.15)$$

For the standard load wave resistance $R_{in} = 50 \Omega$, the term $10 \log(1/R_{in}) = -23$ dB, then

$$_{in} [dBW] = 10 \lg E^2 [\mu V^2] - 10 \lg (10^6)^2 + 10 \lg (\lambda / \pi)^2 - 23 dB. \quad (4.16)$$

The value of input power expressed in decibels with regard to 1 mW power is:

$$\begin{aligned} _{in} [dBm] &= 10 \lg E^2 [\mu V^2] + 10 \lg 1000 - 10 \lg (10^6)^2 + 10 \lg (\lambda / \pi)^2 - 23 dB, \\ _{in} [dBm] &= [dB \mu V] - 113 dBm + 10 \lg (\lambda / \pi)^2. \end{aligned} \quad (4.17)$$

Expression (4.17) represented through the operating frequency value $\lambda = c/f$ has a form:

$$P_{in} [dBm] = E [dB \mu V] + 46.6 [dBm] - 20 \lg f [Hz], \quad (4.18)$$

$$E [dB \mu V] = 20 \lg f [Hz] - 46.6 [dBm] - P_{in} [dBm]. \quad (4.19)$$

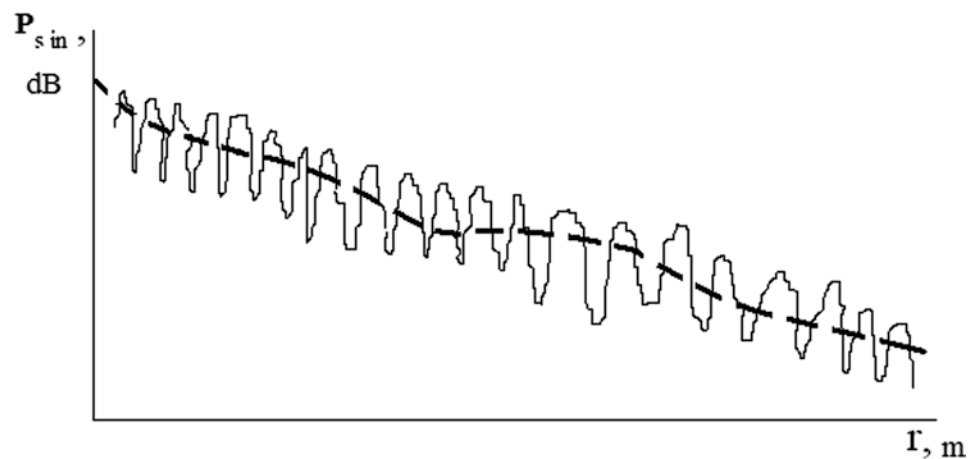
The majority of mobile communication systems (land, trunking, personal) operate under conditions of radio wave propagation at the absence of direct visibility, which increases losses on propagation compared to conditions of direct visibility. On the base of expert data, the model is obtained, which allows the calculation of average losses on propagation in the free space at simultaneous existence of propagation on the way of direct visibility (LOS) and its absence (NLOS) for the r distance between a transmitter and a receiver:

$$L(d) = L(d_0) - 10n \lg(d/d_0), \quad (4.20)$$

where d_0 is the etalon distance – the route segment length to the first obstacle (the segment of direct visibility) – and n is an index, which takes into account the NLOS existence. Typical value for n , taking into account the existence of NLOS, for cellular systems operated outside buildings 3.

We distinguish fast fading and slow fading, which differ by statistical characteristics. The slow fading is caused by relatively small variations of the terrain relief on the signal propagation way. The fast fading is caused by signal reflection both from mobile and fixed objects. Fading of such a type is called multipath fading.

Fig. 4.5 The typical plot of power decrease of received signal versus the distance to the transmitting station



Signal characteristics between the mobile station and the base station of systems of cellular and trunking communication depend mainly on multipath fading because radio waves propagate along the Earth's surface. Figure 4.5 shows the typical plot of power decrease of received signal with regard to the distance to the transmitting station.

With distance from a transmitter growth, we observe a decrease of the average power on the input of the mobile station (dotted line), around which the fast power oscillations are shown with regard to the average value (thin solid line). Variations of average signal power are called *slow fading*, and fast jumps around the average value happening on the distance of several fractions of wavelength are called *fast fading*.

Communication with the mobile stations can be arranged both in motion and without motion. At that, objects can move in different directions and with different speeds. Objects along the motion directions including moving ones lead to reflections, scattering, and energy absorption emitted by a transmitter. These phenomena are accompanied by the formation of several mutually shifted influences at the input of the receiver MS. This offset can be part of or even exceed the length of modulation interval of radio signal. Their simultaneous effect is the result of multipath propagation of radio waves and leads to fading. At that, lengthening of discrete signal elements happens, which leads to its smearing. This phenomenon has got the name of *delay widening*. At the same time, reception of two different signals but closely located in frequency, which have different delay widening, can essentially decrease the mutual correlation. The low correlation level can be provided also by a choice of carrying frequencies and spatial diversity of radio equipment.

Obviously, frequency planning, a choice of network configuration, and cell dimensions should be fulfilled not only taking into account the traffic intensity but conditions of signal propagation. In turn, this is determined by the terrain, the density of buildings, when the signal (entering the receiver) is repeatedly refracted, reflected, and also scattered by objects that absorb electromagnetic oscillations. The influence of listed factors and also a type of used base station antennas allows the formulation of requirements to the model of radio frequency channel. The radio channel model, which takes into consideration all features of signal propagation under specific conditions (the urban building up, the rural terrain, etc.), happens to be rather complicated, and therefore, there are many variants of empiric models (the models of Lee, Hata, Walfisch-Ikegami, and Okumura), applied for conditions, which takes into account the most typical features of the radio channel [4, 20].

The multipath does not have a noticeable effect in satellite systems using repeaters in geostationary orbits for organizing direct TV and radio broadcasting channels, when the angle of arrival of radio waves relative to the Earth's surface is large and the influence of the relief is weak. However, for terminals north of latitude 50° or south of latitude 50° , the elevation angle (between the horizon and the direction to the satellite) is too small. This leads to a weakening of the received signal in the upper atmosphere and additional scattering on ionized layers, causing multipath at the input of the receiving terminal. Satellite personal communication systems have a multilevel structure. At the lower level, systems with the smallest service radius (picocells) operate, then, at a higher level (microcells), wireless telephony with a high information flow rate, then cellular radio communication and radio access systems. Satellite systems are on the upper level. Personal communication satellite systems can be used to create networks in hard-to-reach areas or to connect to mobile terminals via existing cellular networks. The problem with multipath, in addition to typical for TV and radio broadcasting systems, is supplemented by well-known problem of fading in cellular systems. The construction of mobile satellite communication networks is based on the use of constellation of satellites located at different distances above the Earth's surface: in low-earth orbit (LEO), in medium-altitude orbit (MEO), and in geostationary high orbit (GEO) in the equatorial plane. The organization of personal

or corporate satellite communications is based on the concept of using of LEO with a relatively small mass of satellites. Their placement in the polar regions of planet provides coverage of its entire territory (system Iridium). Another option for building a satellite communication system is the placement of satellites with an inclination of their orbits to the equatorial plane (GLOBALSTAR, Skybridge). The service area of system is $\pm 70^\circ$. From the point of view of used structure of satellite communication systems, two parts distinguished: terrestrial and satellite. The ground part and its switching system constantly monitors the satellite part of the system and, when the probability of correct reception decreases due to multipath or satellite hitting the non-operating zone, organizes data transmission over network. The satellite part through the gateway system gains access to the terrestrial network (public or providing high-speed access to the backbone fiber-optic networks).

Signal Propagation Inside Premises Indoor signal propagation has the specific character caused by distance decrease between a transmitter and a receiver, transmitter power decrease, significant signal weakening due to reflections by internal walls and the furniture. Information about the characteristics of radio channel is especially important for system of the cordless telephony, the local computing and sensor systems. The mentioned propagation features lead to lesser time dispersion of signal delays (30–60 ns) in buildings with the small content of metallic constructions and thick walls compared to the typical channels organized in the building (≈ 300 ns) with large free spaces and a variety of metal elements.

The channel inside premises can change in time due to displacement of the mobile terminal or application of the directional antenna, as well as displacement of reflecting objects (persons, office furniture, equipment).

On the base of experimental data, losses in the radio communication channel organized inside premises can be estimated by the relation:

$$L(d) = L(d_0) - 10 \lg \left(\frac{d}{d_0} \right) + X_\sigma, \quad (4.21)$$

where X_σ is the random variable with Gaussian distribution and the dispersion of σ^2 . Such a model has the linear dependence of losses (in dB) versus the distance (d, in dB). The measured values of σ and r allow planning of the WLAN, WPAN, and ZigBee network structure. More complicated models are applied, which take into account the signal propagation in the free space, as well as losses at passing through walls and floors laying on the straight visibility between a transmitter and a receiver.

Influence of Signal Parameters The earlier-considered properties of radio channels took into account only the influence of multipath properties of signal propagation. The significant influence on its properties is exposed by characteristics of transmitter and receiver filters, as well as the digital stream rate in the radio channel (the bandwidth occupied by a signal).

In wideband code division access systems, the signal at the receiver input can be considered orthogonal along the length of the spreading sequence. This also applies to the information signal and its copies, since the time delays of its copies significantly exceed the period of the chip. Consequently, the digital information data at the output of the demodulator will be orthogonal, which allows them to be processed in correlator as independent sequences. This allows the application of correlation processing of each received beam, integrating obtained results to make a decision about received signal. CDMA technology allows each AT to choose its own sequence from a variety of information influences at the receiver input. This is done in the correlator, which multiplies each received sequence by the reference one entered into the fixed memory of the device at the moment of its first registration in the network. Extraction of LOS (or the strongest signal) and its echo-copies and their further parallel correlation processing is performed in RAKE receiver or a matched filter, which significantly reduces the probability of error when deciding on the value of the received information symbol.

Features of Ultra-wideband Signal Propagation Depending on the purpose of ultra-wideband systems, the size of the cell in which the radio communication is carried out, the AT architecture may differ significantly (for more details, see Sect. 2.1.1). Conventionally, it is possible to divide UWB systems (according to the signal propagation range) into those operating in urban environments and indoors (short distance), as well as systems providing a long information transmission range (up to several tens of kilometers). In the first case, the signal attenuation is small, but on the signal shape distortion, essential influence is affected by multipath propagation. Signal propagation over the Earth's surface on the large distance leads to its damping and to distortions of its shape, but practically the multipath influence is absent. The pulse shape of the transmitted signal (Fig. 4.6a) undergoes spreading and smoothing of the final pulse with a seriously change of shape. When using the very short pulses (less than 0.1 ns), distortions become so significant that information extraction from the time dependence becomes impossible. At such pulse duration, the essential signal absorption near the frequency of 22.2 GHz is observed caused by its absorption by the water molecules.

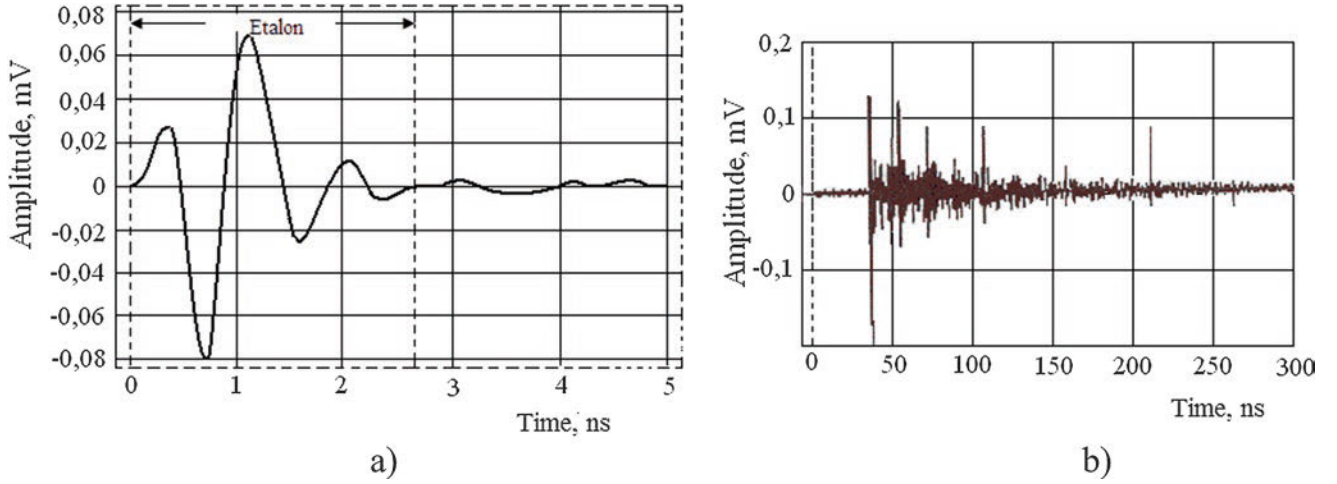


Fig. 4.6 An example of time diagram of signals [14] received on different distances from a transmitter inside premises: (a) fast time, (b) slow time

At present, UWB systems are mainly used in the construction of high-speed data transmission networks over short distances (face recognition, driving without human intervention, interaction between office equipment). When organizing an internal network, it is necessary to take into account the strong dependence of the characteristics of the radio channel on location of walls, ceilings, and floors, as well as on materials from which they are made. An example of time diagram of signals [14, 71] received on different distances from a transmitter inside premises is shown in Fig. 4.6.

The reference pulse is a pulse received at distance of 1 m (Fig. 4.6a) from the transmission antenna, the design of which has selective properties [73]. Its shape is close to that of the bipolar Gaussian monocycle, which is widely used in UWB systems. As we see from Figure 4.4b, signal shape in some premises point essentially differs from the etalon one, as its distribution of the spectral density. Variations in the received signal are caused by its multiple re-reflection and by the existence of shadowing areas, which leads to scattering of the pulse energy. Inter-room walls, which are thicker than the emission wavelength, produce the great shadowing zones, while wave scattering and reflection at propagation inside premises impart the multipath properties of the ultra-wideband signal propagation and cause its interference in the reception point. These factors restrict the pulse repetition period (no more than 300–500 ns) and, relatively, the signal transmission rate in the system. To increase the noise immunity of system in such difficult conditions of signal propagation and the difficulty of predicting radio channel characteristics in the receiving device, it is advisable to use correlation methods of signal processing (RAKE receiver, matched filtering).

The random nature of change in the signal level at the receiver input, caused by the delay and the azimuthal angle of arrival of the beam, shows that, in addition to the correlation processing of the signal, it is advisable to use space-time coding in UWB systems as an effective method for increasing the noise method immunity of signals. The received signal is the sum of the LOS and its copies caused by reflection, scattering, or deflection when the signal passes through radio-transparent structures, containing information about the properties of channel in each beam. If, in narrow-band system, the reflected signals at the input of receivers are reduced in level and differ in phase compared to the transmitted signal, then due to the wide frequency band occupied by the signal, in UWB systems there is frequency dependence of the reflection, scattering, and diffraction coefficients in different parts of its spectrum, which can lead to distortion of its shape. At the same time, the shape distortion of the receiving signal can be different beam-by-beam.

The features of signal propagation and the nature of the arising distortions have formed two types of approaches to methods for studying the properties of a channel and forming its models using a probabilistic approach or deterministic method for assessing the properties of radio channel. When using statistical models, the receiving signal can be described as:

$$s(t) = \sum_{i=1}^N a_i g(t - \tau_i) + n(t), \quad (4.22)$$

where a_i is the parameter defining the fading of the signal amplitude of the i th beam; τ_i is the parameter defining the delay of the i th beam; N is the number of influencing multipath components; and $n(t)$ is the observing noise.

If the pulse delay is less than 2 ns, then we observe the significant correlation between adjacent beams and we can neglect the shape distortions. In real propagation conditions, the impulse delay time is larger, and therefore the correlation coefficient

is sufficiently low, which makes it possible to neglect the distortion of signal shape with a single reflection of the beam. At multiple reflection of the separate beam, we can use the more complicated models for the estimation of received signal parameters, which are based on the estimation of the average energy obtained at some position of a receiver. This average energy is normalized to the full energy obtained at 1 m distance from a transmitter and for known value of the τ delay.

At deterministic estimation of channel properties, the electromagnetic field is calculated in any point of the space, which permits to take into consideration the patterns of used antenna, and channel properties are described by the frequency-dependent transfer function. Frequently, to describe the channel problems, we can use methods of the geometrical optics, which permit to divide the beams into diffraction, reflection, and direct components, as well as some coefficient considering the diffraction.

4.1.6 Inter-symbol Interference

If the data stream rate is comparable to the channel pass-band width or fadings in the channel in time occupy the significant part of the modulation period, the *inter-symbol interference* (ISI) occurs, which essentially complicates the reception and decreases the system performance.

Inter-symbol interference occurs in bandwidth-limited communication channel. Their characteristic feature is the incompleteness of transient process from the previous impulse at the moment of appearance of next one, what is called the *channel memory*. The channel response on the sequence of input signals causes the signal superposition on the channel output. The AFC of the channel with memory significantly differs from the ideal channel leading to distance decrease between signals on the channel output. At the signal synthesis and the code choice for channels, this does not take into account (the channel is considered as an ideal one), but we try to take ISI into consideration at the optimal receiver (a decoder) synthesis.

There are several approaches to fight with the inter-symbol interference:

- The adaptive channel compensation
- The sequential signal decoding with application of the Viterbi algorithm, for which it requires to have the calculated or measured pulse characteristics of the channel, which significantly complicates the signal processing.

The mentioned methods of fight with inter-symbol interference are used at digital modulation of the carrier allocated by the single channel. The mentioned methods of combating inter-symbol interference are used when modulating with a base-band digital signal of a single carrier. Manipulation of large number of carriers (multi-carrier modulation) of radio channel as an alternative to the described methods of combating inter-symbol interference has found wide application in radio communication systems, broadband local area networks (WLAN), and digital television and radio broadcasting systems. In this case, the high-rate digital data stream is divided into the large number of slower streams. Each of them modulates the separate carrier. The decrease of the digital stream rate significantly reduces the inter-symbol interference, which distorts the small part of information symbols. At that, the carrier frequencies can be chosen close enough, and the spectra of modulated oscillations can be partially overlapped, which practically do not influence on the error reception probability, when using the correlation processing of the received signal.

The signal “deformation” one seeks to eliminate by application on Viterbi decoder and its modifications with simultaneous application of the convolution coding algorithm. The basing idea of such an approach is the conversion of the channel with ISI into the totality of Gaussian channels without a memory, which differ each other by the scalar transfer function or by the SNR. Such a conversion is based on the utilization of a large number of orthogonal subcarriers with the frequency modulation – the orthogonal frequency multiplexing (OFDM). Modern digital systems use physical channels to transmit data and control information in form of packets of various lengths, which are separated from each other by guard intervals containing information about the recipient’s address and the place of the packet in the turn at the recipient’s BS. If the length of the package exceeds the allowable size, it is automatically split into two or more packages. During protective interval, the signal on channel input is equal to zero.

Re-reflection, scattering, and delay of the transmitter signal in the radio channel lead to the simultaneous action of a large number of rays at the input of receiver (the main signal and its copies). This complicates the description of the transfer function of the channel model, which takes into account the dynamics of its change, requires the use of more complex signal processing algorithms in the receiver and transmitter, the transmission of information about the channel state from the AT to the BS over opposite channel, etc.

4.1.7 Interferences from the Closely Located Transmitter

This type of interference occurs when two or more MSs operate at the same or close frequencies, which can occur, at near the boundaries of served cell, for example, at frequency planning of network. If transmitter MS2 of neighboring cell generates significantly more power at input of receiver MS1 than the signal of the active BS1, then this leads to masking the weak signal of the active BS1 by the signal of transmitter MS2. The occurrence of such co-channel interference is also possible, for example, in the UMTS with code network scheduling. In direct spread spectrum systems, when all radio channels within the same cell operate on the same frequency (or close frequencies), near-far interference occurs. If the distance between the BS and mobile terminals, one of which is located close to it, and the second is at considerable distance, then they create a different signal level at the BS input. The power level from a closely located MS can create at input of the BS receiver, exceeding the power from a remote MS transmitter by 25–30 dB. This has the effect of masking a weak (distant) MS caused by in-band interference from a nearby MS. In these conditions, if both transmitters have the same power and operate at the same time on the single frequency, then signals received by the mobile terminal from the base station will be masked by signals from the spurious transmitter.

The co-channel distortion created by neighboring cells can be reduced in GSM system by optimizing the frequency reuse in cells the network scheduling process. In the UMTS system, there is no frequency diversity of MS transmitters in cells, and the active influence of neighboring BSs, on the one hand, simplifies the success of soft handoff and, on the other hand, increases the level of system interference. Their intensity is determined by the network load factor, which characterizes the degree to which the system interference power exceeds the receiver's own noise level. An integral estimate of the degree of influence of systemic interference, which depends on the number of active MSs in cell and the "near-far" effect is the required SNR value at receiver input [18]. An effective method of dealing with the signal power scatter at the input of the active MSs. Technically, this is achieved through the use of fast power control system TPS (transmit power control) of MS signals with a frequency of 15 kHz. They are the rather effective means of the high noise immunity achievement of systems operating under conditions of the significant dynamic range of receiving signals, which are typical for mobile communication systems.

4.2 Methods of the Channel Noise Immunity Improvement

Taking into consideration the complexity of analog and especially digital information transmission organization, the interaction of two radio engineering systems is defined by the OSI etalon model of two systems [108]. Since in this textbook, we consider problems of radio receiver construction for wireless systems, we shall be limited by the physical level, which provides the radio signal transmission. The interaction of two terminals with the account of radio channel properties between them should provide the maximal transmission rate in the specified frequency band at the chosen architecture of transceivers and SNR.

Distortions arising in the radio channel and in the radio front end are peculiar to any radio communication systems: analog and digital, fixed and mobile communication independently on the approach to the radio access. The trend towards expanding the range of services provided by mobile systems required an increase in data to transfer rate, which led to an expansion of the radio channel frequency band. And with the transition to digital modulation methods and the active use of multilevel coding, this complicated the algorithms and methods for processing the received signal.

The utilization of the mutual radio channel resource in modern radio communication systems leads to a necessity of interesting channel extraction from a mixture of separate channels, which acts in the receiver input. A task of channel separation can be solved at first stage by antennas converting electromagnetic waves, which are the function of spatial coordinates and time and called *vector* signals. The electrical signal on the antenna output is the time function only and therefore is concerned to *scalar* signals. Accordingly, we can distinguish the methods for signal processing as vector and scalar.

The spatial method is concerned to a number of vector methods, which is based on signal separation according to the geographic position of terminals. Obviously, signal separation in direction is based on the application of antennas with narrow pattern and the low level of side lobes. Another variant of vector separation of the emission sources in the space, which is based on the application of MIMO technologies, includes:

- Utilization of "intellectual" or intelligent antennas permitting to form the narrow directional pattern of signal (beams) transmission
- Application of space-time coding
- Usage of polarization channel division and the polarization signal processing

The simplest variant of the MIMO technology is the method of diversity technique, which is rather an effective approach to increase the radio channel noise immunity widely used in systems of fixed decimeter communication and in the radio access networks of mobile communication systems [109]. This method assumes the use of two or more antennas in the receiver, located at a distance in space. This technology has got the further perfection in wireless local networks of, for example, IEEE802.11, and networks of wireless access of IEEE802.16e and LTE standards. The application of the frequency range near 2 GHz allowed the realization of the orthogonal tuned antennas of base stations of the planar construction for operation in the single frequency range.

Formation on the transmitting side of signals with different polarization allows you to extract a useful signal using antennas with identical polarization properties (Fig. 1.5, [73]) or use less efficient circularly polarized antennas in the MS. Such an extraction of orthogonal in polarization sources can be enough effective at keeping of signal polarization at transmission via the radio channel.

The scalar signal separation is a complicated task since the carrier parameters (an amplitude, a frequency, a phase, the position of time points) and particularly values of the transmitted message are random quantities. At that, if parameters of the carrier oscillation change enough slowly (in some cases, we can consider them as constants), then modulation parameters change on the fly.

Radio signal passage through the common medium leads to variations of all signal parameters with different weight coefficients (the group signal). Recovery of the useful signal (separation with the spurious adjacent channels) is possible if it can be unambiguously defined according to the group signal.

The task of signal separation essentially simplifies if carriers and modulation functions are orthogonal. As modulating a sequence is selected at CDMA technology that is orthogonal in time to all others used in cell, it is a standard stored in long-term memory of MS and BS interacting with it. When a large number of signals, modulated by their own unique sequence, act at the input of the MS, they are compared with the standard in the correlator. At the output of the correlator, a nonzero correlation function is formed when the received and reference sequences coincide, which is the symbol of the Walsh sequence. The correlator acts as “time filter” that only passed the correct (belonging to particular MS) sequence. This fact is decisive when choosing a modulation/demodulation algorithm that is invariant to multipath on the radio channel, which leads to signal fading at the receiver input.

In OFDM technology, the developers ensured orthogonality between baseband digital sequences by choosing the duration of their symbols, creating a space of frequency-independent spectra, the width of which is set by the requirements of the network controller. In the time-frequency domain, it is displayed into the form of radio blocks, at the structure of mutually orthogonal domains. The orthogonality of information symbols in time and frequency is provided on transmitting side and is also preserved at the output of the radio channel. This is provided thanks to the inclusion corrector with a frequency response of the form $x/\sin x$ at the input an upconverter of the transmitter and the input downconverter of the receiver of equalizing filter.

One of the methods in optimizing the operation of the radio communication system is the estimation and prediction of the emission variety of operation signals and interferences in the specified space area, the propagation medium, and receiving devices. The absence of full statistical information about interferences and signals leads to *a priori* uncertainty, and to overcome it, we can apply the simplified integral estimations.

4.2.1 Invariant Signal Reception

In conformity with the communication system, the *invariance* means the equipment stability to external influences. The indirect invariance estimations can be the noise immunity, the transmission rate, the sensitivity to the impact, or, most often, the error probability. Of all the structures of the data transmission system, the receivers connected to the radio channel are most affected. The variety of properties of sources of external electromagnetic influences leads to an unpredictable change in the characteristics of radio channel in time. An attempt to weaken their influence in design of receivers and the introduction of new algorithms and methods for signal processing has not yet allowed constructing a system that is invariant to a random combination of external and internal noise. Therefore, when designing a system, we must restrict ourselves to the fulfillment of a certain preselected invariance condition, for example, only under the action of concentrated interference or Gaussian noise in channel.

The invariance to the *lumped* interferences can be achieved by the application of wideband combined signals with the base $B = 2B_s T_{el} \gg 1$, where B_s is the bandwidth occupied by a signal and T_{el} is the duration of the elementary symbol. The integral estimate of the invariance to lumped interference is the value of the error probability caused by interference and depending on the signal excess of h_s^2 :

$$h_s^2 = P_{in} / P_{int}, \quad (4.23)$$

where P_{in} is the signal power in the receiver input and P_{int} is the power of lumped interferences.

Evidently, the increase of the signal base leads to the decrease of the error probability, making it closer to the invariant value $p_0(h_s^2, 0)$. The interference impact influence can be estimated by the loss in noise immunity $\eta_s = Q_{s0} / (Q_{s0} - \Delta Q_s)$, where Q_{s0} is the signal energy, which is necessary to provide the required value of error probability $p_0(h_s^2, 0)$ at interference absence, and ΔQ_s is the decrease of signal energy in the receiver input caused by the interference.

For real systems of mobile communication and systems of satellite and personal communication, the construction of the receiver, which is invariant, for example, to the frequency *instability* of the master oscillator caused by the Doppler shift, is one of the most important tasks. If to consider that ξ is the random frequency deviation of receiving oscillations with regard to the average value of ω_s , then at application of the phase modulation of the BPSK type in the transmitter, the signal acting at the receiver input is $r(t) = U \cos [(\omega_s + \xi)t + \varphi]$. The signal reception at the absence of the exactly known value of its frequency is possible at the utilization of the noncoherent processing algorithms of the input impact.

At the autocorrelation signal reception with the relative phase modulation (OQPSK) and the decision making according to two consequent received messages of $r_n(t)$ and $r_{n-1}(t)$ with duration of T_{el} , the result of the autocorrelation processing of $r_n(t)$ and $r_{n-1}(t)$ is the transmitted information element:

$$U(r(t)) = \int_0^{T_b} r_n(t) r_{n-1}(t) dt, \quad (4.24)$$

where $U[\bullet] = \pm 1$, ω_s is the frequency of the received signal,

$$r_n(t) = U \cos [(\omega_s + \xi)t + \varphi_n], \quad nT_{el} \leq t \leq (n+1)T_{el}$$

$$r_{n-1}(t) = U \cos [(\omega_s + \xi)t + \varphi_{n-1}], \quad (n-1)T_{el} \leq t \leq nT_{el}.$$

The output voltage is defined according to (4.24) by the multiplication of two sequential messages with the introduction of the time offset per one chip (elementary symbol) T_{el} with the help of the delay line and following integration $U[\bullet] \approx 0.5U \cos [(\omega_s + \xi)(t + T_{el}) + \varphi_n - \varphi_{n-1} + \xi T_{el}]$.

Its dependence on the value of the random frequency deviation ξ , which at $\xi T_{el} > 0.5$ leads to error appearance due to the sign change in $U[\bullet]$, indicates the absence of invariance to random frequency variations of the receiving signal.

We can achieve invariance with respect to random frequency shifts of the input signal during autocorrelation reception by complicating the signal processing algorithm, for example, taking into account the reciprocal correlation of three consecutive pulses.

The matched filter, discussed in detail in Chap. 5 [73], is an example of a device that is invariant to the magnitude of time delays between echoes in multipath reception of signals in data transmission systems.

4.2.2 Adaptive Systems

One of the possibilities to increase the noise immunity of the communication system and to decrease the error probability is the application of the current information about the channel for variation of its parameters or the system structure. Adaptation represents the process of parameter variation and/or the structure of the radio communication system on the base of the current information about the interference situation in the channel for achievement of the required quality of message reception under conditions of *a priori* uncertainty of information about interferences and for varying operation conditions.

Adaptive systems have become widespread in the high-frequency range in the development of frequency-adaptive radio communication and broadcasting by dividing a set of operating frequencies into groups that are equidistant from neighboring frequency groups. Usually, two operating modes are used: frequency changing within the limits of the frequency group (a fight with the fast fading and narrowband interferences) and replacement of frequency groups themselves (a fight with wideband interferences and the slow fading). The operation algorithm of the simplex radio link of the adaptive system anticipates the inclusion in the link structure of the analyzer of interference situation (AIS). The received signal $S^*(t)$ may differ significantly from the transmitted $S(t)$, the indicators of which were influenced by constantly changing parameters of radio channel, the influence of external interference, nonlinearities of the transmitting and receiving paths, delays in the signal processing

and execution of control commands transmitting through the reverse channel, etc. The analyzer of the interference situation (AIS) evaluates the influence of acting factors taking into account their weight coefficients and generates a signal about the interference situation. On this base, the control commands are formed in the control device of the receiving system, and these commands are transmitted through the auxiliary command channel back to the transmitter.

The decision about system formation can be made both at transmitting and at receiving sides. The main algorithm of frequency-adaptive radio link functioning is reduced to a choice of operating frequency, which provides the largest operation time at the error probability less than admissible ($p_0 \leq p_{0adm}$).

The adaptive control may not cover the entire system, but be limited only to change in configuration of the RF front-end on the receiving (or transmitting) side. Such functional elements in modern radio receivers are *antennas with spatial signal processing*, the *interference compensators* in high-frequency sections, *advanced (majority) demodulators of combined signals*, digital signal demodulators with *robust processing algorithm*.

4.2.2.1 Channel Diversity

Fading is one of the very significant distortions introduced by the radio channel. At that, it can be manifested not only in the frequency domain (smooth and selective) but in the time domain (slow and fast) as well. Several approaches are known for decreasing the fading influence on the effectiveness of the radio system. At microwave digital communication systems use fighting fading techniques that are different from those previously used in VHF analog systems. The signal fading at the receiver input is caused by multipath propagation of the signal, leading to an increase in the probability of erroneous reception. In the UMTS system, the multipath in the radio channel is its useful property, which makes it possible to carry out separate processing of the signals of the received beams in the correlation receiver. This is possible if the difference in the appearance time of individual signal beams at input of the receiver is not less than T_{ch} duration, the elementary pulse of the spreading code, which ensures their orthogonality. This will allow, after elimination of time delays, to perform coherent (in phase) addition of the demodulated data.

Signal reception in different channels and their following summation according to some laws is called the *diversity reception*, which can be considered as the combination of methods with the simplest spatial signal processing and the following summation of non-correlated signals.

We can distinguish between explicit diversity when repeating (serial or parallel) signal transmission, for example, transmission of the same signal at the different frequencies. In this case, two received signals are detected separately, and then they are summed according to a certain law. Two received signals are detected separately and then their summation occurs according to some laws.

At *implicit diversity*, the signal is transmitted once only, but owing to properties of the transmission medium and special methods of received signal processing, the possibility to create several reception channels happens to be.

We know several types of the diversity reception applied in the radio communication systems:

- Spatial diversity
- Frequency diversity
- Time diversity
- Multipath diversity
- Polarization diversity

In mobile communication systems, we can use the diversity both at transmitting (transmitter diversity) and at receiving (receiver diversity) sides.

Let us consider the main methods of diversity reception. *Spatial diversity* (Fig. 4.7) is based on summation according to a certain law of signals received by at least two antennas spaced in the space.

The distance between the antennas should be large enough so that the LOS and its echo-copies arriving at the input of receiver are incoherent. This is achieved at a distance between the antennas of at least half the wavelength, which, for example, for the GSM900 system, is:

$$\lambda = \frac{c}{f} = \frac{3 \cdot 10^8 \text{ m/s}}{900 \cdot 10^6} = 0,33 \text{ m} = 33 \text{ cm},$$

where c is the light speed in the free space.

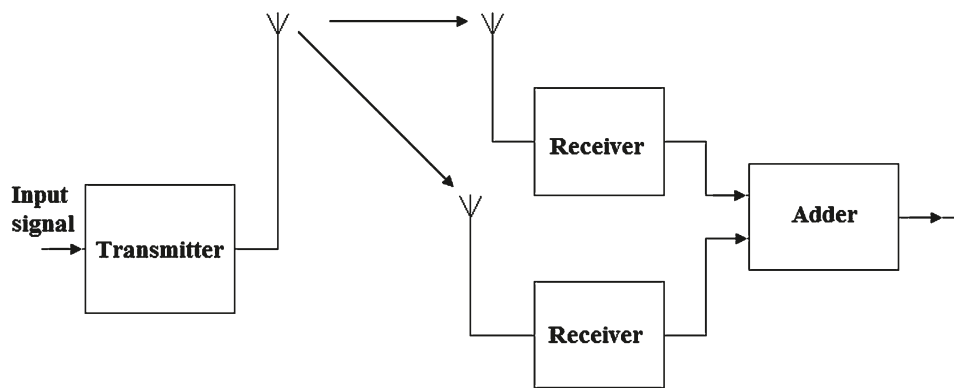


Fig. 4.7 An example of using the method of receiving signals to spatially separated antennas to combat multipath

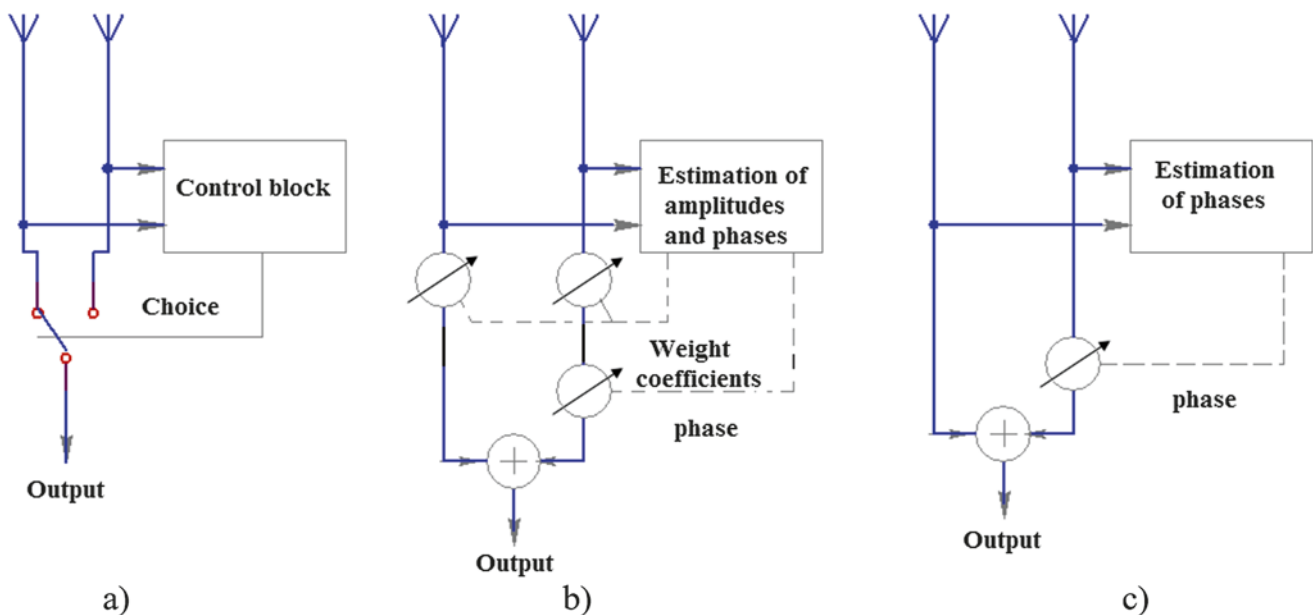


Fig. 4.8 Summation of signal vectors performed with the help of the control block

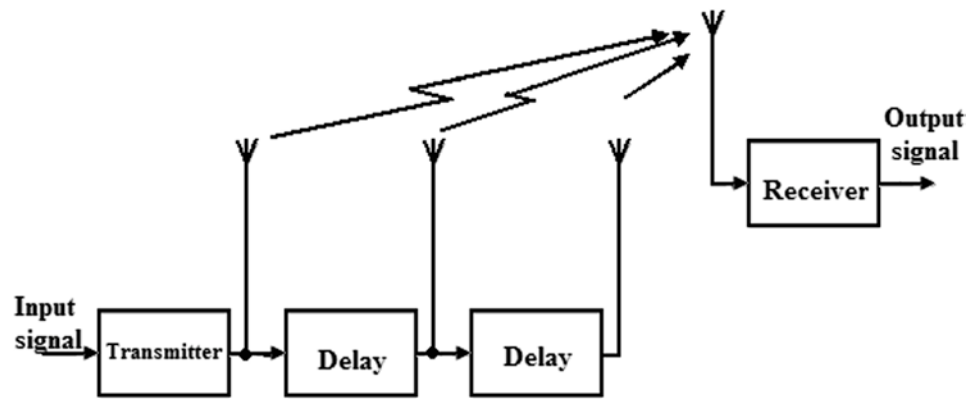
Summation of signal vectors performed with the help of the control block (Fig. 4.8a) uses the algorithm of *summation with the auto-choice* (selection combining). The control unit estimates the SNR in each antenna diversity branch and connects the antenna with the highest SNR to the receive input.

At optimal weighted summation (maximum ratio combining), signals of different branches are equalized in phase (Fig. 4.8b) and are amplified proportionally to their level. Thus, on the adder output, the signal acts with maximal SNR. The most complicated in this scheme is the estimation of signal amplitude at the output of each antenna. The simpler variant of this scheme is when signals are summed with the *equal weight coefficients* (Fig. 4.8c) after phase estimation of input signals (equal gain combining). The spatial diversity requiring the significant complication of the receiving part of the radio station is usually used in the base station.

At **frequency diversity**, the additional transmission channel is organized, which doubles the signal transmission on the main frequency and has the frequency shift sufficient for elimination of the channel correlation.

The **time diversity** is used in channels with relatively fast fading. In this case, the same signal is transmitted through the same single channel in the time intervals, when the signal can be considered as non-correlated. Summing them in the receiver by means of the appropriate approach, we can obtain the decrease of the error probability in the receiving signal. Such an approach for the fight with fading by means of the simple repetition of the transmitted signal is less effective compared with the error correction with the utilization of interleaving.

Fig. 4.9 The approach of diversity transmission



Polarization diversity in communication systems is a type of implicit diversity where antennas with orthogonal polarization are used at the base stations (Figs. 1.5 and 1.6 [73]). In this case, two orthogonal information signals appear at the input of the MS receiver. If they have uncorrelated statistics of signal attenuation, including in a multipath channel, this makes it possible to improve the quality of signal reception, even in the absence of polarized antennas on the MS. The spatial-polarization antenna diversity is the most often used method of signal separation. The simplest antenna system consists of two nonsymmetric vibrators (monopoles), which are oriented with regard to each other on 90° . Often, the vibrators are positioned at an angle $\pm 45^\circ$ relative to the vertical axis, providing identical conditions for signal propagation in each channel. These cross-polarized antennas provide at least 20 dB decoupling between the transmit and receive directions. At the same time, this provides the doubling of the link capacity (for conditions of direct visibility and identity of the receiving-transmitting antennas). However, with a decrease in the operating frequency (HW – LW), the effectiveness of these methods significantly decreases due to loss of directional properties of the antenna and the appearance of depolarization of radio waves by the Earth's magnetic field and water vapor in the atmosphere.

The Multipath Diversity. The described technologies are mainly concerned to the signal reception diversity. Application of the only frequency diversity requires the doubling of some units of the transmitter and, at the same time, requires the diversity of both the reception and the transmission. The diversity reception, in conformity with the cellular communication phone, is difficult for realization from engineering positions (Fig. 4.7). The approach of diversity transmission (Fig. 4.9) is used much more often.

The base station transmitter sends delayed copies of the same signal when using several antennas, which are spaced in the distance. This leads to the formation of several radio channels, which have the different directions of signal propagation from the base station to a receiver, i.e., several branches of diversity are created. At rather large distance between transmitting antennas, each branch is characterized by independent fading. There is a large probability that not all signals of independent channels are weakened by fading, and the receiver can perform the detection of the combined signal, which is formed from several shifted-in-time impacts and weakened by the inter-symbol interference. This will allow the growth of the reception quality.

Multipath diversity is used in wideband systems, if the receiver can distinguish components of the receiving signal, which arrive by different ways at its input.

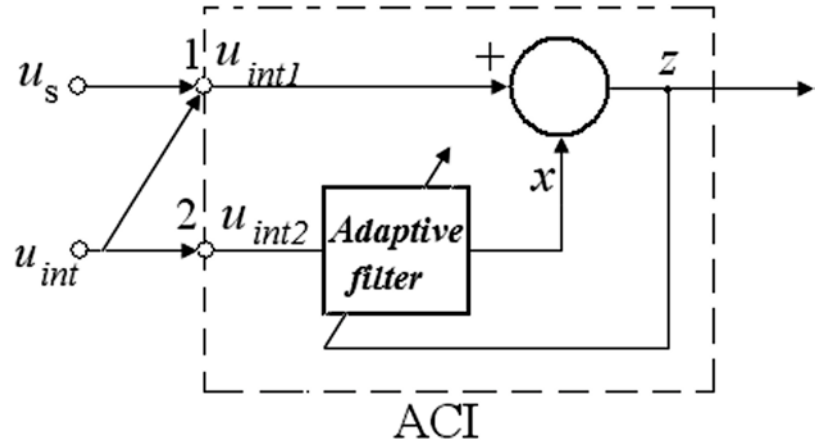
This is possible in the case when relative time delays introduced at signal passage by different ways do not exceed the chip duration of the spreading sequence and the spreading sequence is "white." Signals of different beams are extracted with the help of correlation processing with the spreading pulse sequence, the signal delay time from each station is eliminated, and then signals are multiplied by weight coefficients and are summed. Factually, such a receiver realizes the optimal weighting adding and is called the RAKE receiver, which is widely used in systems with code channel division.

4.2.2.2 Adaptive Units in Receivers

The RF front-end of the receiver performing various operations have been designed together with the digital part to solve the problem of increasing the interference immunity of data transmission systems by circuit and algorithmic methods.

The Adaptive Compensator The adaptive compensator (Fig. 4.10) is one of the units of such a type.

Fig. 4.10 The adaptive compensator



The useful signal u_s and the interference signal u_{int1} are applied to the main input (1) of the adaptive compensator of interferences (ACI), and the interference u_{int2} acts in the compensation input (2) of the device; x is the output signal of the adaptive filter; z is the ACI output signal. If to consider that u_s , u_{int1} , u_{int2} , and x are stationary random processes with the zero mean, and the signal and interferences are non-correlated, while the interferences themselves u_{int1} , u_{int2} are mutually correlated, then:

$$z = u_s + u_{int1} - E;$$

$$z^2 = u_s^2 + (u_{int1} - E)^2 + 2u_s(u_{int1} - E),$$

and the mean value is:

$$M(z^2) = M(u_s^2) + M[(u_{int1} - E)^2].$$

It follows from the last expression that achieving a minimum of average power $M(z^2)$ of the signal at the ACI output with a constant average value $M(u_s^2)$ of the signal at the input is possible only if the interference is compensated by the signal from the output of the adaptive filter x , which corresponds to the maximum excess of the h_s^2 signal.

There are different methods of compensating voltage formation based on the spatial, polarization, or frequency difference of signals and interferences.

Adaptive Antenna Systems The adaptive signal processing using the *spatial diversity* of the signal and the interference has got the widest distribution. This turns out to be very claimed in digital systems of the mobile communication, where besides the signal passed through the main channel, the large quantity of its copies dispersed in time is always present in the receiver input. From the point of view of correct reception, the adaptive weight processing of received signals gives the best result: when signals received in n parallel channels are added taking into account their noisy weights:

$$u_{s\Sigma}(t) = \sum_{i=1}^n q_i u_{si}(t),$$

which is widely used in phased antenna arrays.

In the usual cellular system, the communication between the base station and the mobile station, which is situated in the cell, is arranged with the application of the non-directed or sector antenna. Antennas of the base station emit and receive signals in the sector of 180° , 120° , or 60° . The subscriber devices usually use the isotropic antennas, which at some slight sensitivity decrease simplifies the channel organization. Even at operation with sector antennas, the greater part of the energy emitted by the base station is wasted in vain because the power intended to the specific mobile station is distributed over the whole sector. The formation of the narrow beam directed to the specific mobile station with the varying emission angle at its displacement is the optimal variant of the directional antenna pattern of the base station. Such the controlled pattern would permit the essential increase of the transmitter effectiveness (to increase the SNR in the receiver input of the mobile station or to narrow the radio covering zone of the base station). At the same time, the multipath propagation influence is restricted since in the antenna system input, there are no beams having the large phase shift with regard to the main beam.

Phased Antenna Arrays. The narrow beam of the base station antenna pattern can be realized with the application of the array of spaced antenna elements with the control of the voltage phase, which is applied to each of them from the digital signal processor (DSP) in the real time scale. Such a complex called the phased antenna array (PAA) allows the control of the aperture angle and the direction of the main antenna pattern lobe by means of varying the phase of the control voltage coming from the BB to each antenna element.

PAA application in the receiving part of the base station permits the formation of the spatial characteristics of the antenna array in such a manner to provide the maximal receiving power from the specific mobile station and to decrease interferences from other mobile stations. Signals from the separate receivers, after correction of the phase shifts by the weight coefficients, are added in phase, which increases the output amplitude by M times and the power by M^2 times compared to the output signal of the single-antenna receiver. Noises produced by receivers can be considered as mutually non-correlated, which creates the noise power in the adder output, which is more by M^2 times more than the power of single receiver. Thus, the SNR on the adder output of the receiver with PAA increases by M times compared with the single-antenna receiver.

In the simplest case, the weight coefficient values are calculated in advance at the stage of starting works and are stored in the receiver memory. On the base of these data, by selection of weight coefficients, the required antenna pattern is synthesized. In such a manner, the PAA with switched lobe can be implemented.

The antenna array has a possibility to track for mobile station displacements and provides the maximal values of SNR by directing of the main lobe of the pattern to specific mobile station. At the same time, the PAA control algorithm can assign zero weights to the received rays from the interferers. Such a PAA is called adaptive.

The most complicated in the implementation of such an algorithm of the PAA pattern control is the calculation of signal weight coefficients in the real time mode. In the simplest case for the receiving antennas, which are made from M isotropic antenna elements and linearly and uniformly distributed, their values are calculated as:

$$w_{k,n} = \frac{1}{M} \exp[-j2\pi f_0 \tau_k(\theta_k)], \quad (4.25)$$

where K is the main channel and θ_k is the signal phase shift with regard to the main channel.

In this case, mobile stations are considered as the point sources of signals located at a large distance from the base station. Weight coefficients are calculated only from the condition of phase shift compensation from outside channels.

When the phased antenna array operates for transmission, we must form the antenna pattern with the main lobe direction to the required mobile station. For this, in systems with the duplex channel division (TDD), the weight coefficients calculated for transmission in the uplink direction (at operation at reception) can be used on the base station. In the system with FDD, the main lobe of the antenna pattern is formed in this direction from which the uplink signal is received [109]. In this case, it is simply required to estimate the signal angles of arrival from the mobile stations. There are also other methods of signal arrival direction, for example, on the base of determination of coefficients of the correlation matrix of signals received from M antenna elements [3, 111].

Application of antennas having the narrow pattern («intellectual» antennas) happens to be rather effective in the case of communication organization between base stations.

Adaptive compensators based on the *difference of polarization* of received signals happen to be rather effective for mutual superposition of the interference and the signal in frequency and in time, and they are used on the main lobe of the antenna pattern.

The utilization of antennas made on the mutually orthogonal elements with the application of the following summation provides the interference suppression by 10–15 dB.

Adaptive Demodulators In digital systems of mobile communication, the *adaptive demodulators* are used, which are realized with the application of correlators or matched filters. At correlation signal processing, it is necessary to form on the received side the signal, which is coherent with the transmitter signal. On the receiver side, we only know the carrier (central) frequency of the signal. Its initial phase is provided by a certain error determined by the PLL system of the receiver's master generator. To increase noise immunity and combat fading, hopping of the carrier frequency of the transmitter in mobile communication systems is often used. To search for a carrier that varies according to certain law and capture in receivers, a frequency automatic frequency tuning (AFT) system is used. It provided a lower accuracy of the LO frequency tuning than the PLL, and therefore such demodulators are called quasi-coherent.

Application of *matched filtering* becomes more complicated due to a necessity of the filter design with AFC of the complex shape and, at the same time, having small losses.

With an increase of the operating frequency at the input on quasi-coherent demodulator of the phase-modulated signal, it is required to use high-speed AGC (HSAGC) systems when forming reference voltages. To a large extent, such a demodulator

4.2.2.4 Integral Reception

A very important variety of transmission systems that operate in conditions of a high level of interference with variable and hardly predictable characteristics are systems that monitor the state of the channel and, based on its analysis, select signal demodulation method. Adaptation to the communication channel is performed by choosing a modulation and coding scheme based on information about the success of the transmission of several previous packets. The link adaptation process is dynamic, and the signal and protocol parameters change, including through the use of channel state information from the transmitter. Thus, adaptive modulation systems increase the transmission rate and reduce the bit error rate. In such systems, the decision on the receiving signal (a symbol) is made by the sequential analysis of the sign group.

4.3 The MIMO Technology

There are various ways of increasing the noise immunity of communication channel, that is, reducing the number of incorrectly received information bits without reducing the data transmission rate. Most often, in mobile systems, the counter transmission of information to the BS about the excess in the AT of the admissible error probability is used, with a hybrid method of error correction by automatically generating requests for retransmission of HARQ data packets (Hybrid Automatic Repeat Request).

The method of forming a communication channel with several antennas in MS and BS “Multi Input Multi Output” (MIMO), providing an increase in the data transmission rate due to increased noise immunity, is based on the use of several technologies:

- «Intellectual» antennas allowing the formation of the narrow pattern excluding the multipath reception, which at the same time permits to increase the spectrum utilization effectiveness due to data transmission in parallel beams
- Application of the polarization signal diversity
- Utilization of spatial-temporal coding (STC)

The possibility of increasing the transmission data rate by simply increasing the bandwidth occupied by signal, for the known SNR, as follows from Shannon formula [16], in simplest single-beam SISO channel (One Input One Output) is determined:

$$R = B_{sp} \log_2 (1 + SNR), \quad (4.26)$$

where R is the peak rate of data transmission [bit/s] at the specified bandwidth and SNR for the single-beam channel.

The following growth of the data transmission rate is possible by means of the signal spectrum width increase, the SNR growth, or application of multi-position signals.

When the MIMO technology is used, the peak rate of data transmission in the communication links is defined as:

$$R_{MIMO} = M \cdot B_{sp} \log_2 (1 + SNR). \quad (4.27)$$

Here, the M parameter depends on MIMO configuration: $M = \min\{M_{tr}, M_{rec}\}$, where M_{tr} is the number of transmitting antennas and M_{rec} is the number of receiving antennas. So, if the communication system with the antenna system has $M_{tr} = 2$ and $M_{rec} = 3$, it is equivalent to the communication system with two spatial streams, i.e., $M = 2$.

For nonsymmetric antenna configurations with MIMO (for instance, 1×2 or 2×1), the data transmission rate is:

$$C_{MIMO_{nsym}} = B_{sp} \log_2 (1 + M \cdot (SNR)). \quad R_{MIMO_{nsym}} = B_{sp} \log_2 (1 + SNR). \quad (4.28)$$

It should be noted that in digital systems, when subscribers work in several standards, several antennas can be used to operate in separate frequency range of another standard (or frequency band) on one carrier in order to improve the reception quality, but this is not related to the MIMO technology. In the fourth generation standard, when a large number of subcarriers operate simultaneously in radio channel, they are all transmitted on a common (center) carrier frequency using a single antenna. Multistranded AT with the same purpose can use one or more antennas, for example, to implement separate channels in preselector, for receiving and processing a signal from the frequency range of a 3G, 4G, or WiFi system. In the RF front-end transmitter using MIMO technology, the total information stream with parallel structure at the output of the inverse

Fourier transform is divided into several orthogonal substreams fed to the inputs of separate antennas. Each of them in the space-time domain of the common radio channel generates a signal containing part of the transmitted information, orthogonal to everything created by other antennas. Group of such antennas, united by MIMO technology, generates a signal in the radio channel containing data from all sources at input of the transmission path. On the receiving side, the inverse procedure of forming from the substreams received from the outputs of the antennas, the total information stream of the parallel structure at input of the Fourier transformer is performed.

4.3.1 The MIMO Technology in Mobile Communication Systems

All modern wireless technologies of the third generation as well as the LTE technology, as well as IEEE 802.11 and IEEE 802.16 standards, support the multi-antenna systems (MIMO). To obtain the low cost of the mobile station and taking into consideration the huge market of wireless communication subscribers, the MIMO technology in LTE is maximally simplified [112].

The use of “smart” antennas to allow increase the network throughput due to the complication of their design, control system of directional pattern of the phased antenna array (PAA) and the formation of narrow radiation patterns. This allows to use SDMA Space Division Multiple Access (beamforming), providing data transmission to one or more subscribers, which is especially important at the edges of the cell, increasing the system capacity. It required the application of complex algorithms for control and processing of signal and service commands.

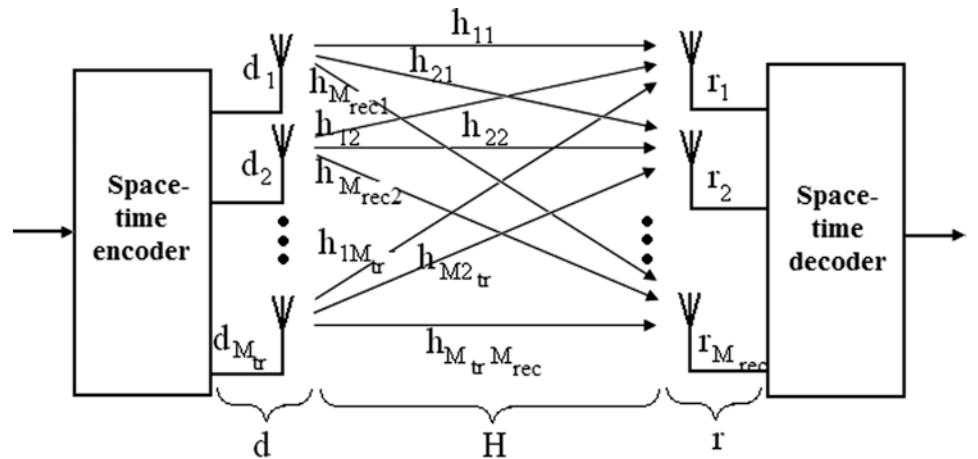
The multi-antenna structure (PAA) does not fight the multipath phenomenon, but thanks to use of a highly directional beam, it does not allow it to be formed (beamforming). Two spatial multiplexing options are used: Per-Antenna Rate Control (PARC) data transmission and Double Transmit Antenna Array (A-TxAA). In the first case, the data transmission rate in each spatial stream is estimated and the information is transmitted to the BS. After analyzing the results, the BS changes the data transfer rate and leaves only the beams that are effectively used. With D-TxAA, the transmit antennas are divided into subgroups, and in each subgroup, the transmitted data is duplicated with orthogonal weights. The scheme of physical channel formation at presence of the same number (M) of antennas in transmitting and receiving sides is shown in Fig. 4.12.

The MIMO technology together with STC provides separation and equalization of signals of different channels. The model of the communication link (propagation of the radio signal) is described by the $\mathbf{H}$ -matrix with dimensions $n \times n$, which elements are the *complex transfer coefficients (fadings) of channels*. The matrix elements of main diagonal h_{ii} are called direct coefficients, and the rest h_{ij} are called indirect coefficients. The unit matrix, of size $(n \times 1)$, describes the signal vector of the transmitting antennas, is denoting as $\mathbf{d}$, and at the inputs of the receiving antennas it is denoted as $\mathbf{r}$ is described by matrix of size $(n \times 1)$. Then, the equation connecting the receiving signals ($\mathbf{r}$) with parameters of transmitting signals ($\mathbf{d}$) and the channel properties ($\mathbf{H}$) under the influence of additive noises ($\mathbf{n}$) has a form:

$$\mathbf{r} = \mathbf{H}\mathbf{d} + \mathbf{n}, \quad (4.29)$$

where $\mathbf{n}$ is the vector on additive noises with dimension $n \times 1$.

Fig. 4.12 The scheme of physical channel formation at presence of the same number (M) of antennas in transmitting and receiving sides



Matrix $\mathbf{H}$ elements with dimensions $(N_{tr} \times N_{rec})$ should be known at the receiving side for realization of the signal processing algorithms. Their values, which are the complex transmission (attenuation) coefficients of the spatial channels, are determined by comparing the received training sequence with a previously known (reference) one on the receiving side. Measured transfer coefficients of spatial channels on the receiving side can be sent to the transmitting side for usage at coding, which increases the effectiveness of the MIMO technology.

Considered principles of construction of the digital radio communication system are close to the well-known approach of the diversity reception. Application together with the MIMO technology of the system for space-time coding (STC) allows the power increase in the receiver input. The space-time encoder, connected in the radio channel (Fig. 4.12), converts the sequential stream of information signals with the help of the demultiplexer on N_{tr} parallel sub-streams. Each sub-stream is coded by the space-time code and after modulation is emitted into a space by the appropriate antenna. All N_{tr} parallel sub-streams are emitted into a space synchronously (simultaneously) at the same frequency band. On the receiving side in each receiving antenna, an EMF is generated as a result of the vector addition of signals from all spatial channels. The level of each impact from each transmitter antenna on separate antenna input of the receiver is determined by the state of separate spatial channel. To find the values of attenuation (fading) in each spatial channel, each antenna of the transmitter emits an orthogonal training sequence processed by each receiving antenna. In other words, the signals described by vector matrices (4.29) act at the input of the space-time decoder, and the receiver solves the system of equations N_{rec} with N_{tr} unknowns, the coefficients in which are the values of the channel matrix. The channel matrix is defined with the learning sequence, and in the receiving side, it is considered as known.

In a specific case of MIMO technology application, when several antennas are used on receiving side and only one antenna at the receiving side, this corresponds to MISO mode – Multiple Input with Single Output. Both of these technologies assume the distribution of the transmitted data between the antennas, while simultaneously ensuring the orthogonality of the transmitted streams. This is done in transmitter after converting the serial baseband signal structure to parallel and modulating the subscribers (QPSK, 16QAM, 64QAM) with orthogonal sequences and performing precoding procedure (*Precoding*). Its essence lies in the use, lock coding (multiplication of radio signal with the matrix $\mathbf{W}$ of code word), that is, performing a “weighted” transformation of each symbol, for example, using the Alamouti code (*C.Alamouti*). For a 2x2 code matrix, it looks like this.

The sequence of complex signals (d_1, d_2) at demultiplexer output is divided into two pairs: even and odd. When in the transmitter, two antennas are used, then in the first time interval, the d_1 and d_2 are emitted, while in the second interval, $-d_1^*$ and d_2^* are described by equations:

$$\begin{aligned} d_1 &= a_1 + jb_1, \\ d_2 &= a_2 + jb_2, \\ -d_1^* &= -a_1 + jb_1, \\ d_2^* &= a_2 - jb_2, \end{aligned} \quad (4.30)$$

where a_i, b_i are the real and imaginary parts of the complex signals, relatively.

From (4.30), we see that:

- If in the first time interval, signals are in phase, then in the second interval, they are out of phase (and vice versa).
- If in the first time interval, signals are out of phase, then in the second interval, they are also out of phase.

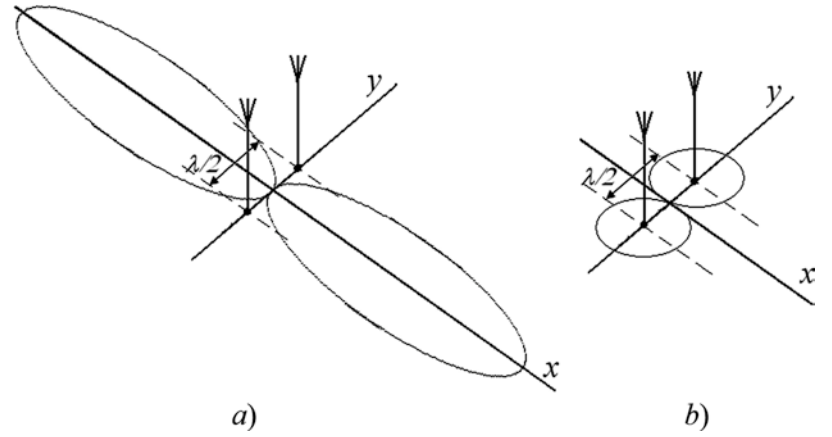
The emission direction of the double-antenna system depends on the phase relations in antennas (antenna elements) (Fig. 4.13).

At in-phase excitation of both transmitter antennas (Fig. 4.13a), the antenna pattern has clearly expressed directional properties, which improves the signal passage conditions under condition of direct visibility absence. Variations of phase relations in different time intervals permit the easier detection of variations arising in the radio channel, for instance, at variation of signal propagation conditions.

In the case of reception from the single antenna, the signal in the receiver input will be combined from signals emitted in the first and the second time intervals:

$$\begin{cases} r_1 = h_1 d_1 + h_2 d_2 + n_1, \\ r_2 = -h_1 d_1^* + h_2 d_2^* + n_2, \end{cases} \quad (4.31)$$

Fig. 4.13 The emission direction of the double-antenna system depends on the phase relations in antennas



where r_1, r_2 are signals in the receiving antenna input in the first and second time moments, relatively; h_1, h_2 are complex transfer coefficients of the channel; and n_1, n_2 are components of the additive noise in the channel in the appropriate time intervals.

On the base of received signals r_1, r_2 and known transfer coefficients of the communication channel h_1, h_2 , we obtain the recovered signals similar to transmitted ones:

$$\begin{cases} \hat{d}_1 = h_1^* r_1 + h_2^* r_2, \\ \hat{d}_2 = h_2^* r_1 - h_1^* r_2. \end{cases} \quad (4.32)$$

Substituting (4.31) into (4.32) and neglecting by the influences of additive noises in the channel, we have the equations defining the connection of the recovered signal (estimation) with the channel characteristics:

$$\begin{cases} \hat{d}_1 = (|h_1|^2 + |h_2|^2) d_1, \\ \hat{d}_2 = (|h_1|^2 + |h_2|^2) d_2. \end{cases} \quad (4.33)$$

The analysis of (4.33) shows that the signal level in the channel output is proportional to the square of modules of channel fadings, which at the same SNR in the receiver input allows the growth of data transmission rate in the channel.

Application of the large number of receiving-transmitting antennas using the space-time coding changes the order of transmitting signal formation. The signal power in the communication channel output, using the MIMO STC technology, equals to the sum of powers which arrive into the receiver from all transmitting antennas.

The Bell Laboratories Layered Space-Time (BLAST) algorithm and its modifications are widely used in transceivers using space-time coding MIMO technology. It is used when the number of transmitting antennas does not exceed the receiving ones. As a criterion of optimal reception, a criterion based on the achievement of the minimum mean square error MMSE is used.

The MIMO technology can provide two types of transmission: spatial multiplexing and diversity transmission.

In the mode of *spatial multiplexing*, each antenna channel transmits the independent information stream, and all channels are non-correlated. In the downlink, two types of transmission with spatial multiplexing are possible: diversity transmission for one subscriber unit MISO (Single User MIMO: 2x1, 4x1) and MU MIMO (Multi User MIMO: 2x2, 2x4) for a group of mobile station. In the first case, the base station forms several independent streams of the single mobile station with the number of antennas, which is not less than the number of streams from the base station. In the MU-MIMO mode, the information stream from the base station with the similar frequency-time parameters is transmitted to several mobile stations. At that, the procedure of the antenna pattern formation with the help of the adaptive antenna system is not used. The growth of the used antenna number does not increase the number of information streams since, in all channels, only two code words can be transmitted, i.e., operation with two mobile stations only.

The *transmission with diversity* assumes the transmission of the single data stream through several antenna channels, which lightens the fight with fading and increases the information transmission quality and (indirectly) increases the information transmission rate.

Subscriber devices having the single antenna also can use the MU-MIMO technology in transmission mode, since even at the utilization of the single frequency resource, the base station forms non-correlated information streams that provide the simultaneous communication with several mobile stations.

In the UTAN network, high-speed data transmission HSPA technology is widely used to organize a communication channel in conjunction with multi-antenna structures (MIMO technology). This suggests that several factors must be considered when choosing a MIMO signal generation algorithm, the combination of the number of transmit and receive antennas. The main of them are:

- A number of receiving and transmitting antennas
- Modulation types in each antenna and data transmission rates
- Methods of input data distribution between subgroups at demultiplexing

The variant of organization of the communication channel, which transmits the single stream by several wideband signals with different weight coefficients defining with the help of the feedback (the closed MIMO loop), is possible. In [113], we can be acquainted in detail with features of channel organization and algorithm operation principles.

4.3.2 Adaptive Antenna System and the MIMO Technology in the IEEE 802.16 Standard

The most common technology is MIMO based on space-time coding (STC), which is used as mandatory in local wireless information networks (WiFi), and is also used in IEEE802.16-2004 wireless networks broadband access (WiMAX) and global (UMTS) and regional (cdma2000) wireless networks of the third generation, as well as in networks of the fourth (LTE, LTE-Advanced) generation.

In networks of the wideband access, which have got the wide distribution, the methods are actively used, which increase their effectiveness and noise immunity. Application of adaptive antenna systems (AAS) and the MIMO technology is concerned to them.

The MIMO Technology for the IEEE 802.16 Standard. In the IEEE 802.16 standard, the AAS supporting means are embedded in base stations and mobile stations. Application of AAS essentially increases the potential capacity of the wideband access network of the IEEE 802.16 standard, since the operation is possible in the same channels (frequency and OFDM) in different sectors of the base station. Application of directional antennas allows the decrease of the output transmitter power, which at the same time reduces the inter-channel interference.

Application of multielement antenna systems or several antennas to construct the communication channel allows the utilization of methods of space-time coding (diversity), which improves the signal passage in channels with a memory. The aim of the STC application is to disperse the signal from the single source in space or in time, i.e., to change conditions of the signal passage. The probability of the correct signal reception after primary processing in the receiver essentially increases.

In contrast to systems of mobile communication, which use the mobile station of the limited physical sizes, usually two transmitting antennas are mounted for the single sector of the base station in the IEEE 802.16 standard and two receiving antennas in the mobile station. The configurations 2×2 , 4×4 , 4×2 , 8×2 , 8×4 , 8×8 are permitted to use (transmitting $\times$ receiving antennas). The number of antennas in the base station should not be less than two at application of the MIMO technologies, while on the mobile station there are one, two, or four transmitting antennas. Introduction of the MU-MIMO mode widens possibilities of the MIMO technology providing simultaneously and at the same frequencies the information transmission to different subscribers. At that, for two transmitting antennas on the base stations, up to two subscribers are supported, while at four, eight transmitting antennas, up to four subscribers can be supported.

We know several methods of the space-time diversity. In the IEEE 802.16 standard, the circuit is used offered by Alamouti [113], when the output symbol stream is divided into two streams formed in parallel according to some algorithm, for example, containing only even or odd symbols (Fig. 4.14).

The architecture of the base station and the subscriber device with regard to the radio channel realizes the structure of the MIMO type (several inputs and several outputs). A receiver and a transmitter have the conventional radio front-end structure of the digital radio communication station, which uses the orthogonal frequency multiplexing and demultiplexing.

On the base station, two antenna channels are created, which act in parallel and provide the synchronism of transmitted radio signals (the digital stream in each channel is formed from the common digital sequence).

Application of multiplexing with OFDMA together with STC requires the introduction of the set of pilot frequencies to form OFDM symbols taking into account the parity and the number of antenna channels. In WiMAX standard, in addition,

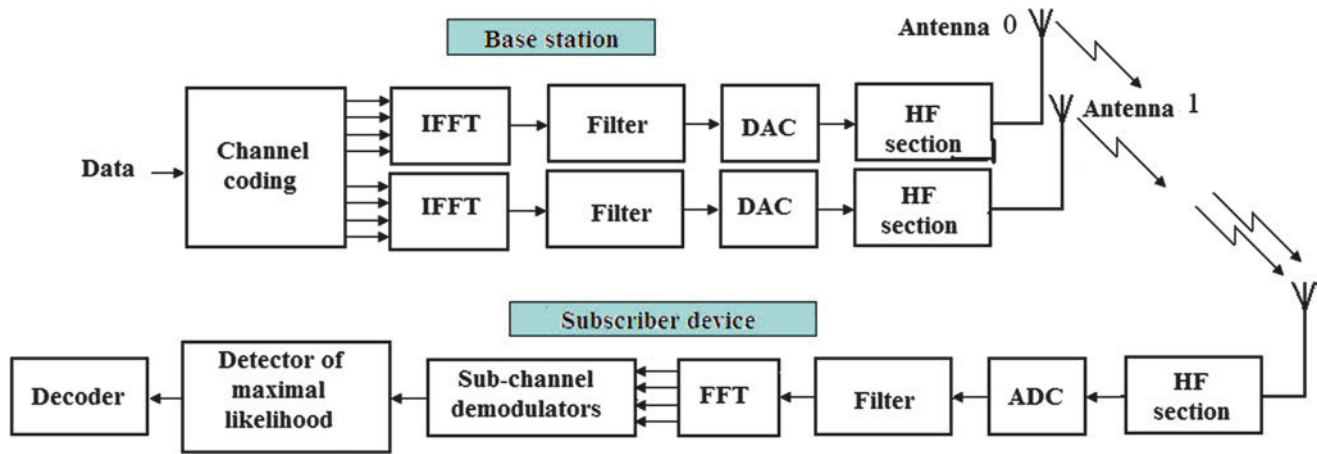


Fig. 4.14 The circuit for space-time diversity offered by Alamouti

with OFDM technology and space-time diversity (MIMO STC), frequency diversity implemented by frequency hopping diversity coding (FHDC). The formation of the OFDM symbol at its transmission by the antenna with the number 0 occurs at modulation by the signal vector S_0 of carriers in the sub-channel N and in the $N+1$ sub-channel by the signal vector S_1 . The antenna with the number 1 transmits the signal, in which the carrier frequencies of the sub-channel N are modulated by the vector $S \times I$ and the channel $N+1$ by the vector S_0 . Antenna number 1 transmits a signal in which the carrier of the subchannel N is modulated with the vector S^*_{-1} , conjugated with the vector S_1 , and the sub-channel $N+1$, with the vector S^*_0 . Recovery of the transmitted signal at the receiving side is the same as with STC technology, but due to the use of cross-modulation, instead of the transmission functions of spatial channels h_0 and h_1 (4.29), characteristics, obtained from the analysis of carriers of N and $N+1$ subchannels, are used. According to the received signals r_0 and r_1 in the sub-channels N and $N+1$, the transmitted signals S_0 and S_1 are recovered.

The OFDMA technology allows the STC/FHDC application not only in the downlink channel but in the uplink channel as well. The use of space-time coding is possible when placing not only two but also four antennas in the AT. In this case, in added two antennas ($0'$, $1'$), the applied signal is shifted in phase ($S'_0 = S_0 \cdot e^{i\phi_0}$).

Impetuous development of local computing networks and their active competition with systems of mobile communication led at present (recommendation of the IEEE 802.16e standard) to practically obligatory implementation of the MIMO technology in the mode of space-time coding and adaptive antenna systems.

The MIMO Technology for the IEEE 802.11 Standard. Features of construction of local computing networks led to the appearance of the macro diversity handover mode (MDHO), which provides the mobile station operation located in the environment of several base stations and transmitted information of the specific mobile station.

At application of two, three, and four antennas in the transmitter, the STC mode can perform the sub-channel distribution according to different algorithms: **PUSC** (partially used subcarrier), **FUSC** (fully used subcarrier), and **AMC** (adaptive MIMO switching). The utilization of the direct signaling mode at operation with adaptive antenna systems is not recommended to application because, in this case, the exact knowledge of the base station beam is required, in which the mobile station is situated and this is practically unrealizable for mobile subscribers.

To implement the MIMO technology (using Alamouti algorithm) or the adaptive pattern formation in the platform *4Motion*, it is assumed to use several antenna variants.

For the formation of independent channels in BS transmitters for the implementation of the MIMO technology, various options for antenna configurations are used: one-dimensional, two-dimensional, and three-dimensional. The most widespread are the designs of antennas with one-dimensional linear geometry [73] and a variable angle of inclination of the antenna panel in azimuth. They allow organizing a large number of spatial channels, the independence of which is ensured by placing monopoles on the antenna plane at different angles, forming channels with different polarizations. Such a polarization provides the channel diversity not less than 20 dB. The utilization of cross-polarized antennas is the effective variant of the antenna channel diversity, since at beam re-reflections, the polarization on the receiving side can be changed to the opposite compared to the transmitted polarization.

For adaptive formation of the pattern, the antenna arrays from fourth closely located elements with vertical polarization can be applied.

Fig. 4.15 The device realized on the chipset Intel RD2 (a) has six embedded antennas located inside the case and on the chipset from Beceem company has two embedded omnidirectional antennas (b)

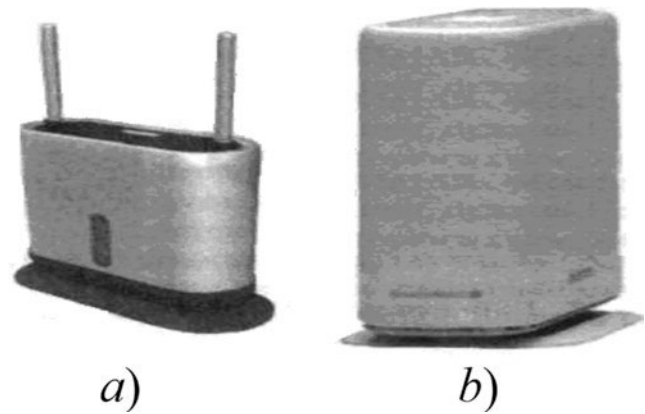


Fig. 4.16 The device BreezeMAX 4Motion PC Card



All mentioned antenna configurations can be used with larger or smaller effectiveness both for the formation of the controllable pattern and for the organization of two or four MIMO channels.

For connection of the terminal equipment with the *BreezeMAX 4Motion* platform, we can offer the different variants of the terminal telecommunication customer premises equipment (CPE), which is mounted on the client territory. These can be devices including external and internal module-installed outdoors (*BreezeMAX 4Motion PRO CPE*), or similar equipment installed separately indoors (*BreezeMAX 4Motion Si*), as well as PC Card modems and a USB port.

The subscriber device *BreezeMAX 4Motion PRO CPE* consists of the outside and inside modules. The outside module contains the plane integrated antenna and the control system for the antenna pattern. The outside module is connected to the inside module by an Ethernet cable, which transmits control signals to the external module, the monitoring signals for the state, and the power source voltage. The internal module is connected to the network of subscriber data transmission or to the personal computer through the standard IEEE 802.3 Ethernet interface, as well as can include two voice ports for transmission via VoIP (Voice over IP). The Wi-Fi module of IEEE 802.11b/g standards is anticipated for organization of the local access point.

The *BreezeMAX 4Motion Si* is a compact unit made in the form of a chipset and is installed by the customer indecently inside the premises and has all the capabilities of the *BreezMAX 4Motion PRO CPE* device. The device realized on the chipset Intel RD2 (Fig. 4.15a) has six embedded antennas located inside the case and on the chipset from Beceem company has two embedded omnidirectional antennas (Fig. 4.15b).

This device is directly connected to the personal computer (PC) and is activated through the SIM card or with the help of special application. Several variants of the device *BreezeMAX 4Motion Si* are produced, which are intended on operation in the frequency ranges of 2.3, 2.5, and 3.5 GHz.

The device *BreezeMAX 4Motion PC Card* is the network adapter permitting to PC connection to the network of the mobile WiMAX (Fig. 4.16).

Fig. 4.17 The device US210

It is designed to operate in the 2.3 frequency bands and 2.5 and 3.5 GHz with 10 MHz channel bandwidth and provides downlink speed of up to 20 Mbit/s and uplink speed of up to 7 Mbit/s. The bandwidth of the RF front end can be changed by choosing it to be equal to 5, 7, 8.75, and 10 MHz, varying its configuration. Two sliding antennas allowing the optimization of the signal reception by the position adjustment are used, and this provides the stable operation in the network at a distance to the base station of up to 5 km.

Development of mobile networks of the wideband access led to the appearance of WiMAX USB adapters for PC. The device US210 (Fig. 4.17) corresponds to the IEEE 802.16e standard and supports the mobile wireless connection at a speed of up to 130 km/h. The device has a USB port, with the help of which it connects to the AT and is user-configurable providing a peak speed in downlink up to 33 Mbit/s, while in the uplink up to 7 Mbit/s. This adapter is intended for operation in ranges 2.3 GHz and 2.5 and 3.5 GHz providing the output power 23 dBm, and it has the gain of 2 dB. Having the one transmitting antenna and two receiving antennas, the adapter US210 supports the MIMO technology.

Mobile WiMAX has been promoted since 2008 in Moscow and St. Petersburg by Scarlet Company (YOTA trademark). Formable networks should provide coverage of these cities and adjacent main highways in the range of 2.5–2.7 GHz. The equipment used provides access to Internet at a speed of up to 10 Mbit/s for each user without interrupting the communication session during intranet roaming when the AT moves in space up to 120 km/h.

4.3.3 *The MIMO Technology in the IEEE 802.11 Standard*

The IEEE 802.11 standard, which includes a large group of standards (IEEE 802.11a–IEEE 802.11z), was created as a wireless Ethernet. Let us limit ourselves to considering the standard's indicators at the physical level in the reference model of an open network. The main task that was solved when building the IEEE 801.11 and HIPERLAN standard is its compatibility with the connected wired networks Internet, PSDN, iSDN, etc. and transparency for user applications. The diversity of standards is explained by the great need to replace wired E2E connections and intensive work in the development of new technologies and circuit solutions. The standard IEEE 802.11b sets the parameters for the operation of equipment in the range 2.4000–2.4835 GHz. Network users with random access ("ad hoc") are classified as nomadic, that is, they can move in the zone of mutual radio accessibility of terminals, but at a speed not exceeding 1.4 m/s. The basic data transfer rate in physical channel is 1 Mbit/s. To combat multipath and increase the noise immunity of signal, DSSS technology is used, when user data is presented as a sequence of chips with a speed of 11 Mbit/s. The 11-element Barker code is used as an expanding sequence. The resulting sequence modulates the carrier according to the DQPSK or ODQPSK scheme, forming a radio

signal spectrum with a width of approximately 22 MHz. To accommodate partially overlapping spectra in the frequency band of radio channel, their central frequencies will be spaced by 30 MHz.

The IEEE 802.11b standard provides for the transmission of a signal with a speed of 11 Mchip/s, and 5.5 Mchip/s. To do this, instead of the expanding Barker sequence in data field, the Complementary Code Keying (CCK) proposed by M.J.E. Golay is used. A sequence a of length k is called complementary to another sequence b of the same length, having autocorrelation functions:

$$k-1 \quad k-1$$

$$A(j) = \sum_{j=1}^{k-1} a_j \cdot a_{j+1} \quad \text{and} \quad B(i) = \sum_{j=1}^{k-1} b_j \cdot b_{j+1}.$$

$$j=1 \quad j=1$$

where k is possible shift.

Sequences are complementary if: $A(0) + B(0) = 2k$ and $A(i) + B(i) = 0$ for any w not equal to zero. When transmitting data in a multipath channel, this property, when using the CCK, means that the main beam signal acting on the receiver input and its echo copies will be mutually orthogonal (independent) if they are delayed in time more than T_{ch} (elementary symbol). When using complementary encoding of CCK, 8 bits ($d_0, d_1, \dots, d_7$) are allocated from the information stream. They form complex symbol $C(c_0, c_1, \dots, c_7)$ consisting of 8 complex chips that modulate the carrier with frequency of 11 MHz. Each chip c_i of symbolic sequence C , using the Euler transform, can be represented as: $c = e^{j\psi_i} = \cos\psi_i + j\sin\psi_i$, where ψ_i is the phase of i -th chip. The chips are complex and, in relation to QPSK quadrature modulation, correspond to in-phase (I) and quadrature (Q) channels.

The IEEE 802.11n standard is designed for 5 GHz frequency range using OFDM technology (Multi-station Access Method). In the USA, unlicensed parts of range from 5.15 to 5.35 GHz and from 5.725 to 5.825 GHz was allocated for this, only 300 MHz (instead of 80 MHz in the IEEE 802.11b standard). The new standard used MIMO technology with multiple antennas for the first time.

A dedicated frequency band of 5.15–5.35 GHz allows you to organize 8 non-overlapping channels of standard width of 22 MHz. When using OFDM technology in each such band, the inverse Fourier ($N=2=64$) is performed at 64 points, forming 64 orthogonal subcarriers. The main part of them (subcarriers 52) is used for information exchange (48 for data transmission and 4 frequencies – pilot). The remaining 8 subcarriers are included in the protective gap between the channels and do not carry useful information.

The frequency diversity of subscribers for the selected number of Fourier transform points is: $\Delta f = 20 \cdot 10^6 / 64 = 312.5$ kHz. Each of these subcarriers can be represented by $f_k(t) = a_k \sin[2\pi(f_0 + k\Delta f)t + \varphi_k]$, where $k = -26 \dots 26$. Subcarriers are modulated, depending on type of channel being formed, according to the law: BPSK, QPSK, 16QAM, or 64QAM.

Elementary pulses (chips) at the output of each modulator can be represented as:

$$C = a \sin x + jb \cos x = A e^{jx}, \quad \text{and the total signal: } S(t) = \sum_{k=-26}^{26} c_k e^{j2\pi(f_0 + k\Delta f)t}$$

where c_k is the complex amplitude of the modulating subcarrier chip. When using the quadrature modulating in the RF front end, c_k components correspond to the in-phase (I) and quadrature (Q) channels. The subcarrier with frequency f_0 is not used, its amplitude is always zero. Technology OFDM, using orthogonal subcarriers modulated with sequences orthogonal in time and frequency [73], is a very effective means of dealing with multipath. This is typical for WLN that allow a station to communicate with other stations or servers, terminals, printers, inside and outside buildings, resulting in signal fading caused by multipath. The combined use of OFDM and MIMO technology is formulated in the standard IEEE 802.11n.

The simplified diagram explaining the principle of space-time coding (STC) is shown in Fig. 4.18.

The diagram reflects the algorithm of space-time coding (MIMO with STC) at the utilization of two antennas. The main task of MIMO with STC technology is division and equalization of signals of different channels. In this case, the communication channel model represents a 2×2 matrix, the elements of which are complex transfer coefficients (fadings) of channels. Matrix element knowledge is necessary for signal processing on the receiving side. Complex transfer coefficients of the matrix are defined with the help of the learning sequence, which is known in advance on the receiving side. If fadings measured on the receiving side are transmitted to the transmitter and are taken into account in the coding process, this additionally increases the effectiveness of the MIMO with STC technology.

Because the emission direction of the double-antenna system depends on signal phase relations in antennas (between its elements), the direction of the resulting emission will change in time. This essentially improves the signal passage, especially under conditions of the direct visibility absence.

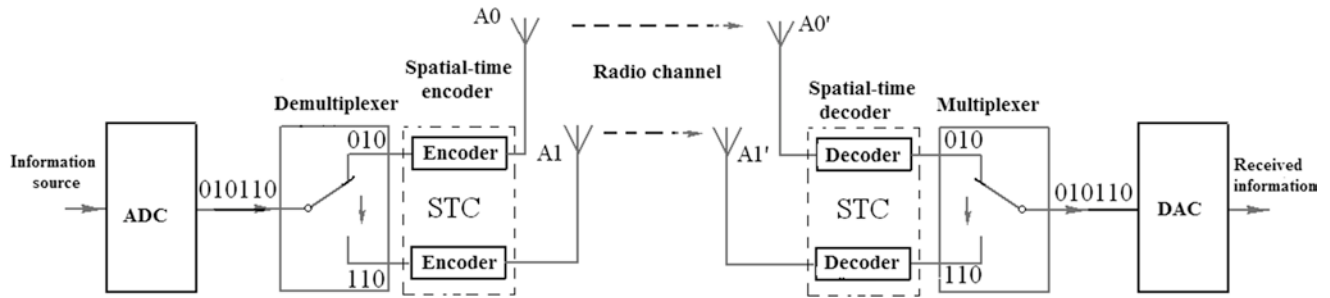


Fig. 4.18 The simplified diagram explaining the principle of space-time coding (STC)

For the simplest case (Fig. 4.18), the sequential digital stream from the ADC output is divided by demultiplexer into two parallel streams and passes to the STC encoder. To increase the speed of information stream, an encoder with a rate of $R=5/6$ is used, and if the base data is more than 300 Mbit/s, even two convolution codes. The coded stream is divided into several spatially separated streams, the number of which (N_{ss}) cannot be less than the number of transmitting antennas (N_{tx}).

When using a larger number of antennas, the encoded stream is divided into several spatially separated streams, the number of which (N_{ss}) cannot be less than the number of transmitting N antennas. Two complex conjugate streams ($x = a + jb$ and $x^* = a - jb$) acting at the input of the spatial streams shaper are distributed at its output to an even number of antennas with the number of bits determining the number of complex symbols $S(ss)$ in each stream. After the alignment of the number of bits on each subcarrier sequence of each stream, it is arrived at the interleaver and then at the quadrature modulator. After performing the inverse Fourier transform for each time sequence and equalizing the time shifts between the modulated subcarriers of the same stream, a protective between OFDM symbols is introduced. Then the generated stream is passed through the Nyquist filter and transmitted to the input of the frequency converter “up,” transferring spectrum of separate spatial stream to the general operating range. Then, decoding is performed in the receiver and the task of unified digital stream formation is solved.

Application in the IEEE 802.11n standard of several antennas in the receiver and the transmitter in base configuration anticipates the support of two antenna channels by the equipment of access points and the single channel for mobile consumer stations. Totally, the subscriber radio station and the mobile station can have up to four antenna channels for reception-transmission.

The structure “Multi Input Multi Output-Orthogonal Frequency Division Multiplex” uses several antennas for transmission and reception of signals along with simultaneous utilization of a large number of subcarriers in the radio channel. The structure of such a type allows signal processing for the wide class of multi-station access systems, which operate under conditions of the direct visibility absence (NLOS).

The MIMO-OFDM configuration enables the utilization of its advantages under the absence of direct visibility by means of application of the equipment for the multipath reception on the base stations. The “Multi Input Multi Output-Orthogonal Frequency Division Multiplex”(MIMO-OFDM) uses several antennas for transmission and reception of signals and the large number of subcarrier at the radio channel. This process forming spatial sequences using a space-time encoder (STBC) provides an increasing transmission rate by shortening the control sequences and reducing the length of the protective intervals. In the system, MIMO-OFDM uses several antennas to transmit segments of information flow in the direction of another station. When the station switches from signal reception mode to transmission mode, it uses Carrier Sense Multiple Access Technology with Collision Avoidance. It allows the station to listen to the radio broadcast before transmitting information and when an IFS inter-frame interval is detected, in which the channel remains free, and transmits the packet through a random delay interval. This allows you to avoid collisions when the device accesses the network. In addition, since the data transmission occurs at the same time and in the same frequency band, and spatial signal division is used, such a structure increases the spectrum utilization effectiveness. This technology has got the development in the VOFDM (vector OFDM) method.

The IEEE 802.11 (very high throughput) standard, on the base of which the ultrahigh-frequency local wireless networks will be realized in the range below 6 GHz, must provide the information transmission rate up to 500 Mbit/s. This task is solved by means of integration of four channels with a width of 20 MHz, utilization of the OFDMA multiple access with orthogonal frequency division, and application of eight antennas with diversity on the reception and transmission sides.

Checking Questions

1. What are principal differences in planning of the service zone in systems of fixed and mobile communication (and radio access)?
2. Which types of interferences and distortions are typical for digital systems?
3. Designate the main numerical characteristic for estimation of spurious impacts in digital systems.
4. What is characterized by the parameter “quality of service” QoS?
5. What is the numerical estimation of the “level of service”?
6. Enumerate the measures which decrease the probability of blocking.
7. What is the numerical estimation in the co-channel interferences?
8. How can we decrease the level of co-channel interferences?
9. What is the reason of inter-channel interferences? How their reduction can be provided?
10. What are the reasons of nonlinear distortion appearance and methods of their reduction?
11. Enumerate the reasons of fading appearance at propagation of narrowband and ultra-wideband signals in the free space.
12. Which distortions are arisen at radio signal transmission in the channel with the restricted bandwidth?
13. What do we mean by the term “invariance” with respect to communication systems?
14. Enumerate the methods to increase the channel noise immunity.
15. Give examples of adaptive methods’ utilization in the structure of digital radio communication and radio access systems.
16. How does the use of multi-antenna systems affect the signal transmission rate in the radio channel? Explain by the example of using MIMO technology in standard IEEE 802.15n.
17. Calculate the signal power at the input of the subscriber receiver of the LTE system in W for the power matching mode with the antenna, if the power at the antenna output is minus 102.2 dBm.

The power expressed in dBm indicates the normalization of the absolute quantity with regard to 1 mW:

$$_{dBm} = 10 \lg \frac{P_A, W}{1 mW} = 10 \lg \frac{P_A, W}{10^{-3}, W};$$

$$\frac{_{dBm}}{10} = \lg \frac{P_A}{10^{-3}}.$$

Fulfilling potentiation of this expression, we obtain:

$$10^{0,1 P_{A, dBm}} \cdot 10^{-3} = _{W}$$

$$_{W} = 10^{0,1 \cdot _{dBm} - 3} = 10^{-0,1 \cdot 102,2 - 3} = 63,1 \cdot 10^{-15} W.$$

Appendix: Interference Influence on the Main Receiver Indices

Application 4.1 Calculate the mean carrier power for the ideal QPSK system with the coherent demodulation and the linear amplification under conditions of Rayleigh fadings and a noise.

Analysis conditions:

- Error probability $P_e = 10^{-3}$
- Transmission rate $f_b = 48$ kbit/s
- Receiver noise factor $NF = 7$ dB

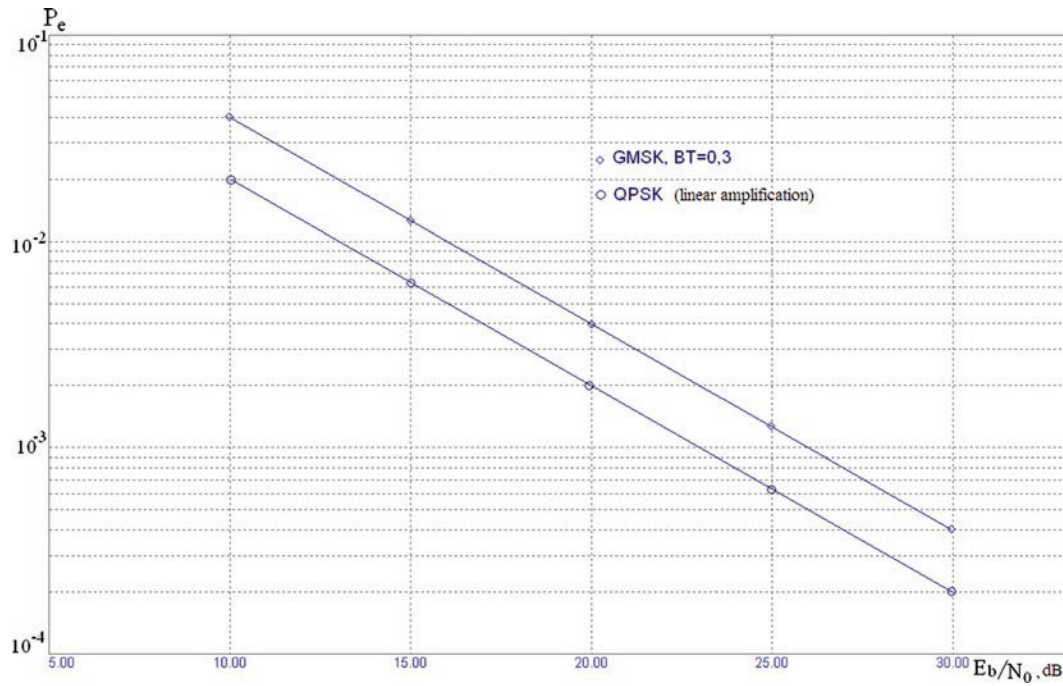


Fig. 4.19 Error probability graphs versus SNR for various types of manipulation

Only one transmitter signal acts in the receiver; the slow fading with the Rayleigh distribution acts in the channel. We can neglect the Doppler frequency shift and interferences from external sources; therefore, thermal noises (4.4) only act in the receiver input:

$$NF_T = kTB_n NF, [\text{dB}].$$

For the room temperature $t = 17^\circ\text{C}$, which is equivalent to $T = 290^\circ\text{K}$ and $k = 1.38 \cdot 10^{-23}$ J/Coul, the spectral density of thermal noises, which are created by the tuned antenna receiver input is $kT = -174$ dBm/Hz.

For the given error probability $P_e = 10^{-3}$, we obtain from the plot in Fig. 4.19: $E_b/N_0 = 23$ dB (the ratio of the energy per 1 bit to the power spectral density of thermal noises) from the ratio

$$\frac{E_b}{N_0} = (\text{SNR}) \frac{B_n}{f_b}.$$

Taking into consideration that in the ideal QPSK system, the noise band $B_n = f_b/2$, we calculate the ratio of the carrier power to the thermal noise power in the receiver input:

$$\frac{E_b}{N_0} = (\text{SNR}) = \frac{24 \cdot 10^3}{48 \cdot 10^3},$$

$$\frac{E_b}{N_0}, \text{dB} = (\text{SNR}), \text{dB} - 3, \text{dB}, \text{ then}$$

$$(\text{SNR}) = \frac{E_b}{N_0} + 3 = 23 + 3 = 26 \text{ dB}.$$

When there are no outside electromagnetic spurious impacts in the receiver input, the noise power in the frequency band of $B_n = 24$ kHz in the receiver input is:

$$NF_T = kTB_n NF = -174, \frac{dBm}{Hz} + 10 \log(24 \cdot 10^3) + 7, dB = 123.3 dBm.$$

To satisfy the condition $SNR = 26$ dB, the carrier power in the receiver input is:

$$C = NF_{T,dBm} + SNR_{,dB} = -123.2_{,dBm} + SNR_{,dB} = -97.2 dBm.$$

The obtained value satisfies the requirement to the sensitivity of the subscriber mobile receiver of the IS-95 standard.

Application 4.2 Calculate the signal power in the input of the mobile station of the UMTS standard taking into account the inter-channel interference influence.

The examples given in the appendix can be divided into two types: when the initial information is from the standards of specific system, according to which the cascade structure is calculated, or when, based on the known characteristics of the device, the possibility of achieving the requirements of the standard is evaluated. The real power of signals, which act in the subscriber equipment input, is defined by many conditions: radio route characteristics, the network loading, the data transmission rate and subscriber displacements, and the input SNR. The speed of information transmission in the radio channel of a separate cell largely depends not only on its characteristics but also on reception conditions in neighboring cells, as well as the electromagnetic compatibility of UTRAN and GERAN radio access networks and a number of other factors that determine the bandwidth of the network as a whole. Due to the complexity of the mutual influence account of the mentioned reasons on the reception quality, we shall be limited by the impact of the following interferences in the consumer equipment input:

- Arising due to the noncoherent signal reception from the proper base station and caused by the multipath propagation
- Created by base stations of the adjacent cells

BS and AT transceivers of modern mobile communication systems, as follows from the open systems interconnection (OSI) of interaction at the physical level, at addition to the traffic channel, support the constant exchange of service information (synchronization channel, pilot-channel, broadcast channel, etc.) over dedicated or shared channels. At the formation of the group signal, they are multiplied by their channel-formation codes and are exposed by scrambling. Then, each of them is multiplied by the weight coefficient providing the maximum power of the transmitted control signals and pilot signals. Their resulting power should not be more than 15–20% of the maximal transmitter power. The smallest coefficients, which are inversely proportional to the speed of information transmission, are assigned to data transmitted through the traffic channel and control channels.

In the receiver input, the total interference power [114] should not exceed the ratio of the energy per 1 bit to the power spectral density of thermal noises E_b/N_0 :

$$\frac{E_b}{N_0} \leq \frac{SP \cdot P_j}{P_n + P_{int.own} + P_{int.adj}}, \quad (4.34)$$

where P_j is the signal power of the j th subscriber in the receiver input; $P_{int.own}$ is the interference in the receiver input created by the own base station due to the noncoherent reception; $P_{int.adj.}$ are interferences created by base stations of adjacent cells; and P_n is the noise power in the receiver input.

Analysis conditions:

The multipath property of radio wave propagation and also nonideal channel filtering, which are formed by the base station, leads to the appearance of the noncoherent noise in the input of consumer equipment. We can consider that the level of such interferences is 0.1–0.125 of the power of the useful receiving signal.

The power of interferences produced by signals from adjacent base stations should not be more than 0.5 P_s of the useful signal power.

The ratio of the energy per 1 bit to the power spectral density of thermal noises in the receiver input should be $E_b/N_0 = 7.4$ dB [114].

The receiver noise factor is $NF = 5$ dB.

The radio signal transmission rate is $R_{bit} = 12.2$ kbit/s.

The rate of the radio signal transfer is $R_{chip} = 3.84$ Mchip/s.

The spread factor (SF) is $SF = R_{chip}/R_{bit} = 315$.

The reliable reception of the synchronization signal is provided under the condition $P_s = 1.5 P_n$, where P_n is the total noise power in the receiver input of the mobile station:

$$P_n = kT_0 B_n NF.$$

The power spectral density of thermal noises, which is created by the matched antenna for the room temperature $T = 290$ K°, is:

$$kT_0 = 1.38 \cdot 10^{-23} \cdot 290 = 4 \cdot 10^{-21} [W] \Rightarrow -174 \frac{dBm}{Hz}.$$

The noise power in the consumer receiver input is:

$$P = kT_0 + 10 \lg B_n + NF = -174_{dBm/Hz} + 65.84_{dB} + 5_{dB} = -103.2 dBm.$$

The interference power of adjacent channels is:

$$P_{int.own} + P_{int.adj.} = 0.125 P_s + P_s = 0.125 P_n \cdot 1.5 + 1.5 \cdot P_n.$$

The inequality (4.34) takes a form:

$$\frac{E_b}{N_0} = \frac{SF \cdot P_j}{P_n + 0.125 \cdot 1.5 \cdot P_n + 1.5 P_n} = \frac{SF \cdot P_j}{2.7 P_n}. \quad (4.35)$$

Then P_j is the signal power of j -th subscriber in the receiver input:

$$P_j = -10 \lg SF + 10 \lg \left(\frac{E_b}{N_0} \right) + 10 \lg (2.7 P_n) = -10 \lg 315 + 5 - 10 \lg 2.7 = -118.8 dBm,$$

which satisfies to the standard requirements.

Application 4.3 Calculation of the power at the input of the receiver of a mobile station of the IS-95 standard.

Calculate the average power of the carrier (C) for the ideal QPSK system with coherent demodulation and the linear amplification under conditions of the Rayleigh fading and a noise.

Condition of an analysis:

- The error probability $P_{er} = 10^{-3}$
- The transmission rate $f_b = 48$ kbit/s
- The noise factor of the receiver $NF = 7$ dB

Only one transmitter acts in the receiver input; the slow fading with Rayleigh distribution acts in the channel. We can neglect the Doppler frequency shift. We believe that in the UHF and microwave ranges, we can neglect interference from external sources due to large losses of the signal in free space in these frequency ranges and the absence of the effect of bending the Earth's surface, as well as the difficulty of predicting them, so we will take into account only thermal noise at the receiver input.

For room temperature $t = 17$ °C, which is equivalent to $T = 290$ K and $k = 1.38 \cdot 10^{-23}$ J/Coul, the spectral density of thermal noises produced by the tuned antenna is $kT = -174$ dBm/Hz.

For specified error probability $P_{er} = 10^{-3}$, we obtain $E_b/N_0 = 23$ dB (the ratio of the energy per 1 bit to the power spectral density of thermal noise from the plot in Fig. 4.19). Using the expression

$$\frac{E_b}{N_0} = \frac{C}{N} \frac{B_n}{f_b}.$$

Taking into account that in the ideal QPSK system, the noise band $B_n = f_b/2$, we calculate the ratio of the carrier power to the thermal noise power in the receiver input:

$$\frac{E_b}{N_0} = \frac{C}{N} \frac{24 \cdot 10^3}{48 \cdot 10^3} = \frac{C}{N} \frac{1}{2},$$

$$\frac{E_b}{N_0}, dB = \frac{C}{N}, dB - 3, dB, \text{ then}$$

$$\frac{1}{N} = \frac{E_b}{N_0} + 3 = 23 + 3 = 26 \text{ dB}.$$

When there are no external spurious electromagnetic influences in the receiver input, the thermal noise power, which is produced by an antenna in the band $B_n = 24 \text{ kHz}$ in the receiver input, is:

$$N_n = k'' B_n NF = -174, dBm / Hz + 10 \log(24 \cdot 10^3), dB + 7, dB = 123,2 \text{ dBm}.$$

To fulfill the condition $P_s/P_n = C/N = 26 \text{ dB}$, the carrier power in the receiver input is:

$$! = N_{T, dBm} + C / N_{dB} = -123,2_{dBm} + 26_{dB} = -97,2 \text{ dBm}.$$

The obtained value satisfies the requirement to the sensitivity of the mobile receiver of the IS-95 standard systems.

Application 4.4 Calculation of the power at the input of the receiver of a UE of the standard UMTS with account of inter-channel interference influence.

Conditions of an analysis:

The ratio of an energy per 1 bit to the power spectral density of thermal noises in the receiver input is $E_b/N_0 = 7.4 \text{ dB}$.

The receiver noise factor is $NF = 5 \text{ dB}$.

The speech transmission rate is $R_{bit} = 12.2 \text{ kbit/s}$.

The radio signal transmission rate is $R_{chip} = 3.84 \text{ Mchip/s}$.

The spread factor (SF) is $SF = R_{chip}/R_{bit} = 315$.

Despite the fact that a system with CDMA technology at input of the UE receiver, in addition to the useful signal, there are many signals intended for the subscribers of cell, we will limit ourselves to the action at the input of only the useful signal. The real signal power acting in the UE input is determined by a series of conditions: characteristics of the radio routes, the network load, the data transmission rate and movement of a subscriber, and the SNR in the input.

The bandwidth in each cell is limited not only by the width of the allocated radio frequency band, it is also necessary to take into account the influence of reception conditions in other cells, electromagnetic compatibility of radio access networks UTRAN and GERAN, and other factors.

Due to the complexity of the mutual influence account of the mentioned reasons upon the reception quality, we are limited by the interference impact in the UE input:

- Arising due to the noncoherent reception of signals by the own base station caused by multipath properties at propagation
- Produced by base stations of the adjacent cells

Signals emitted by the base station, besides traffic channels, include also the control channels (the pilot channel, P-CCPCH, etc.). In the process of forming a group signal, it is subjected to channel coding and scrambling, which increase its noise immunity. Then, each of them is multiplied by the weight coefficient, which provides the maximal power to control signals and pilot signals. Their resulting power must be not more than 15–20% of the maximal transmitter power. The minimal coefficients, which are inversely proportional to the information transmission rate, are conferred to the traffic signals and control signals.

The multipath propagation of radio waves as well as the nonideal filtering of channels, which are formed by the base station, leads to the appearance of the noncoherent noise in the UE input. We assume that the level of such interferences is 0.1–0.125 of the power of the useful receiving signal.

The interference power, which is produced by signals of adjacent base stations, should not be more than $0.5 P_s$ of the useful signal power.

The total power of interference acting at the receiver input must not exceed (E_b/N_0) the ratio energy per 1 bit to the spectral power density of thermal noise determined by the standards:

$$\frac{E_b}{N_0} \leq \frac{SF \cdot P_j}{P_n + P_{inter, own} + P_{inter, ad}}, \quad (4.36)$$

where P_j is the signal power of the j th subscriber in the receiver input; $P_{inter, own}$ are interferences in the receiver input, which are created by the own base station owing to the noncoherent reception; $P_{inter, adj}$ are interferences created by base stations of the adjacent cells; and P_n is the noise power in the receiver input.

The reliable reception of the synchronization signal is provided under the condition $P_s = 1.5 P_n$, where P_n is the total noise power in the receiver input of the mobile station:

$$P_n = kT_0 B_n NF.$$

The power spectral density of thermal noises, which are produced by the matched antenna for the room temperature $T = 290 K$, is:

$$kT_0 = 1.38 \cdot 10^{-23} \cdot 290 = 4 \cdot 10^{-21} \text{ W} \Rightarrow -174 \text{ dBm} / \text{Hz}.$$

The thermal noise power created by an antenna in the receiver input is:

$$P_n = kT_0 + 10 \lg B_n + NF = -174_{\text{dBm/Hz}} + 65.84_{\text{dB}} + 5_{\text{dB}} = -103.2 \text{ dBm}.$$

The interference power of adjacent channels is:

$$P_{inter, own} + P_{inter, ad} = P_s + 0.125 P_s = 1.5 P_n + 0.125 P_n \cdot 1.5.$$

Replacing the inequality (4.36) with the equation (a sticker condition), we obtain (4.37):

$$\frac{E_b}{N_0} = \frac{SF \cdot P_j}{P_n + 0.125 \cdot 1.5 P_n + 1.5 P_n} = \frac{SF \cdot P_j}{2.7 P_n}. \quad (4.37)$$

The inequality (4.37) resolved with respect to P_j permits to calculate the power of the j th subscriber in the receiver input:

$$P_j = -10 \lg SF + 10 \lg \left(\frac{E_b}{N_0} \right) + 10 \lg (2.7 P_n) = -10 \lg 315 + 5 - 10 \lg 2.7 = -118.8 \text{ dBm},$$

which satisfied to standard requirements UMTS [114].

Chapter 5

Technical Indicators of Devices for Signal Reception and Processing

This chapter is devoted to the main technical indices of modern radio receivers: their definition, methods of calculation and optimization. In Sect. 5.1, we discuss the main types of interferences important for communication systems. Then we introduced the most significant parameters of receivers: sensitivity (real, limiting, threshold), distinguishability factor, noise factor, selectivity (spatial, time, amplitude, phase, frequency), and RF front-end factor. Then we describe the problem of blocking, the transfer characteristic, the amplitude characteristic, the dynamic range, nonlinear distortions, and probability of error appearance. Section 5.2 studies the problems of internal noise sources, methods of noise property description, and models of noise sources. Special attention is attracted to noises in the receiver input and in bipolar and field-effect transistors (with their noise equivalent circuits). A large part of Section 5.2 is devoted to the problems of non-linear harmonic distortions in the preselector and mixer of the RF front-end receiver. After main definition, we discuss the intermodulation distortions of different orders, the intercept points (on input and output), the blocking phenomenon, the cross modulation, and methods for dynamic range broadening. At the end of this chapter, we describe the Appendix: the calculation of admissible (caused by antenna) and real noise factors.

5.1 Main Technical Indices

In addition to the useful signal at the input of the receiver, there are various types of external influences, as well as interference that are created inside the receiver and cause problems when receiving with the required accuracy.

Depending on the origin, these impacts differing from useful signals can be divided into:

- Interferences from the spurious radio stations, which can be considered as narrowband, create in its output signals of adjacent (in frequency) channels besides useful signals. In receivers of the superheterodyne type, they produce also interferences in the image channel, in the channel of direct passage, which are caused by insufficient selectivity of the radio front end.
- Atmospheric interference mainly arises from atmospheric electricity and, having a wide spectrum, has a significant impact on the parameters of useful signal (Table 1.1) in low (30...300 kHz), medium (300...3000kHz), and lower part high-frequency (3...30 MHz) of the range.
- Industrial interferences created by devices producing sparkles in the operating cycle (wideband spectrum).
- Space interferences, which sources are the Sun and several regions of the Galaxy, produce interferences of the noisy character for communication, radio navigation, and radar systems.
- Atmospheric noises, which have the fluctuation character and cause signal scattering or absorption on the water and oxygen molecules; noises of the thermal Earth's emission.
- Antenna noises caused by the chaotic (thermal) motion of charge carriers in the conductor, which has a resistance (losses).
- Self-noises of amplifying elements.

The main technical characteristics of the receivers are well known from the literature, as well as the ways to achieve them. However, the introduction of receivers into the structure of digital microwave communication systems, the use of digital

modulation and demodulation and new methods of multiple access have led to the expansion of the list of their technical parameters.

Let us consider the main characteristics, sources, and models of noise sources and their qualitative estimation.

One of the most important electrical characteristics of the radio receiver is the **sensitivity**, which is the receiver's capability to receive the weak signals. Quantitatively, the sensitivity is estimated by the minimal EMF in the antenna E_{Amin} , by the electric field strength near the receiving antenna E , or by the nominal power P_{A0} of the normally modulated signal, at which the signal in the receiver output is reproduced with the required quality. The required quality is understood as either obtaining a given signal level at output of the receiver at a certain signal-to-noise ratio (SNR), or providing one the probabilistic criteria for signal reception (e.g., BER).

Obviously, the sensitivity depends on the signal amplification in the receiver and increases when the signal level at the receiver output increases with increasing gain, which is true at low interference levels at the receiver input and a relatively high input signal level (with a low gain of the receiving path). However, with the receiver gain growth, the noise level in its output increases, which is caused by the transfer of active components of the radio front end into the nonlinear mode, which is accompanied by the appearance of additional frequency components. In this case, even with the sinusoidal nature of the input action, its harmonics, multiplying in the AE with the interference components, which is caused by the nonlinearity of its amplitude characteristics, will form in the RF front-end of a random process (noise), which reduces the signal-to-noise ratio at its output. Thus, the sensitivity at a specific reception quality *is restricted by the gain of the radio front end*.

In real conditions, the SNR at the output of the RF front-end, which determines the receiver sensitivity, can be very different. Simply increasing the RF front-end gain to increase sensitivity will not only amplify the signal but also increase interference, and the minimum signal level that can be received will be limited SNR at the detector input. So in analog systems the SNR is about 18 dB; in digital systems, thanks to efficient coding with error correction and multiple access, it is significantly reduced. For GSM standard, which has signal base $B = 1$, the SNR at output of RF front-end is 9 dB, and for the UMTS system using spectrum expansion technology, the SNR value is 5 dB. In this case, the signal power at the receiver input will be 20 dB less than noise power (at spreading factor $SF = 25$ dB). Since interferences can be external and internal, then even at the absence of external interferences, the amplified internal interferences will always be in the output of the radio front end, and they will mainly have the fluctuation character. The level of internal interferences defines the *limit of sensitivity* of the receiver.

In the UHF range, the main type of interferences is the receiver internal noise; however, this does not permit to provide any desired sensitivity, since the antenna is not only the signal source but also the noise source $E_{n,A}$. The equivalent circuit of the tuned antenna should contain the generator of noisy EMF $E_{n,A}$ together with the generator of signal EMF E_A .

Thus, when designing the receiver, we must use the concept of real sensitivity, by which we mean the ability of the receiver to receive weak signals for a given SNR at the output of the RF front-end. For digital radio communication systems, the concept of the receiver *threshold sensitivity* is used, which is the minimal received power necessary for the achievement of the acceptable level of characteristics, such as the maximal value of the error probability per bit: **BER** (bit error rate).

In this frequency range, the sensitivity is qualitatively estimated by the nominal power P_{A0} of the normally modulated signal, at which the output signal is reproduced with the required quality. By nominal (maximum) power, we mean the power that is dissipated at the input of the receiver in the matching mode between the signal source (generator) R_{gen} , which is an antenna with R_A , and the input resistance of a two-port radio receiver R_{in} ($R_A = R_{gen} = R_{in}$, Fig. 5.1).

The antenna radiation resistance R_{rad} usually essentially exceeds the loss resistance r_{loss} , i.e., we may consider $R_A \approx R_{rad}$, which usually is called the *internal resistance* of the receiving antenna. The gain of the radio front end in power $K_p = P_{s,out}/P_{s,in}$,

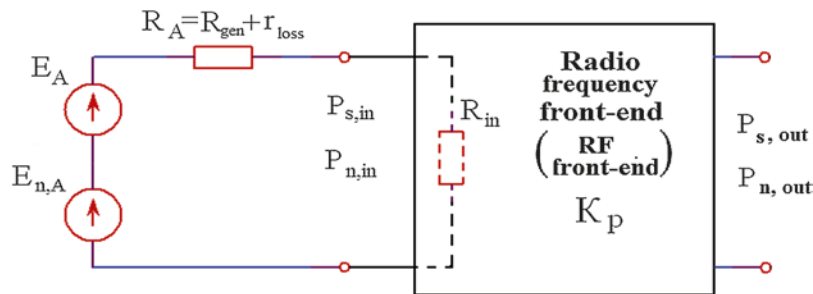


Fig. 5.1 The equivalent circuit of the input circuit

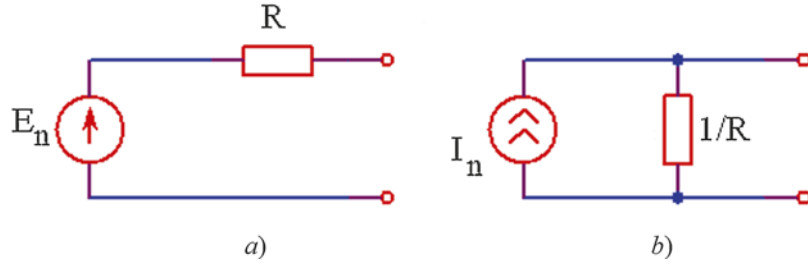


Fig. 5.2 Equivalent in series (a) and parallel (b) circuit of noise source

in . By radio frequency front-end (RF front-end), we mean the part of the radio receiver from its input (antenna connection point) to the input of the detector (demodulator) or DSP, which converts the radio signal into analog or digital signal of the modulation frequency. This may be the electrical signal passing after amplification onto the electric-acoustic converter, which reproduces the received audio signal or the signal in the form of message (video, text, etc.).

The signal exceeding over the noise (SNR) on the output of the radio front end can be estimated by the **distinguishability coefficient** $D = P_{s,out}/P_{n,out}$, which minimal value is defined by the requirement on the signal reproduction quality, the modulation type, and the character of the received message. Then, the signal power in the input of the radio front end in the matched mode is:

$$P_A = P_{s,out} / \kappa_p = P_{n,out} D / K_p. \quad (5.1)$$

The minimal value of distinguishability D_0 , which is limited by internal noises of the radio front end, defines the real receiver sensitivity:

$$P_{A0} = P_{s,in} = P_{s,out} / \kappa_p = P_{n,out} D_0 / K_p. \quad (5.2)$$

Internal noises of the first stages of the receiver, which are fluctuation noises [35], are mainly produced by all passive components having losses and active devices (transistors, electronic lamps). The thermal motion of the free carriers of charges creates the fluctuation currents and hence the appearance of the noisy voltage on the terminals of resistors. These currents and voltages are called thermal noise, which has homogeneous spectrum (“white noise”) up to frequencies $10 \div 10$ Hz and zero mean value. This allows the presentation of any component, which has losses, in the form of the generator of the noise EMF E_n or the generator of the noise current I_n , which expressed the generating properties of the passive component with losses. A real resistor that generates thermal noise is represented by EMF or current source that has an “internal” resistance R (Fig. 5.2a) that does not create noise (an ideal, not noise) and equal in magnitude to real resistor, or “ideal” conductance $G = 1/R$ (Fig. 5.2b)

Energy spectra of noisy EMF (or current) are defined by the Nyquist formulas, and its dispersions (the value of the rms noisy EMF) $\bar{E}_n^2 = 4kTRB_n$ or current ($\bar{I}_n^2 = 4kTB_n/R$), where $k = 1.38 \cdot 10^{-23}$ J/K·Hz is the Boltzmann constant; T is the absolute temperature of the circuit in Kelvin degrees; and B_n is the noise (integral or energy) band in Hz, within which limits the noise are calculated or measured.

Besides real or virtual resistors (for instance, the equivalent output resistance of the previous stage, which is the resistance of the signal source for the following stage), transistors and semiconductor diodes are also the sources of fluctuation noises. In bipolar transistors, it is the terminal noise of bulky resistances of base, emitter and collector regions; fluctuations of a number of charge carriers passing through E-B and C-B p-n junctions (the shot noise); fluctuations of collector and base currents caused by the process of generation and recombination of free charge carriers of opposite sign in the base structure (flicker noise). In the field-effect transistors (FETs), the main contribution in the level of self-noises is made by thermal noises of the body resistances of the source and drain, thermal noises of the channel, which is the resistor controlled by the gate voltage, and the shot noise of the gate leakage current.

Noises created by an antenna in the receiver input are noises of the outer space, noises of the Earth’s surface and its atmosphere, as well as thermal noises of the antenna losses r_{loss} . External sources, which include outer space, the Earth’s surface and its atmosphere, as well as thermal noise caused by antenna losses, create noise at receiver input. In the moderate frequency range, industrial interference of different nature and spectral characteristics contributes significantly to the overall noise level.

For the convenience of calculating the RF front-end of the receiver, the antenna noise generated by all the listed sources is included in R_A of equivalent antenna circuit (Fig. 5.1), heated to the temperature T_A , which is called the *effective noise temperature of the antenna* T_N . A large number of noise influences form $E_{n,A}$ (5.3), representing R_A as non-noise resistance (Fig. 5.1). To describe the antenna as source of thermal noise, it is necessary to increase the antenna noise temperature (T_A). Neglecting the thermal noise caused by the power loss in antenna (r_{loss}), the noise temperature in the antenna will be: $T_A = T_{cos} + T_{atm} + T_{Eart}$ where T_{cos} , T_{atm} , and T_{Eart} are values of the noise temperature caused by the penetration of cosmic noise, atmospheric noise, and noise into the antenna caused by the thermal radiation of the earth. These components are complexly depend on antenna direction pattern, nature of the spectral noise sources in receiving area, operating frequency range, etc.

For quantitative estimation of the noise properties of the radio receiver, we can use the concepts of the *noise factor* N and the *noise temperature* T_n , defined for the radio front end, i.e., without the influence of spectrum distortions (noises), which are produced by the first nonlinear device – the detector.

The value of $E_{n,A}$ is defined by the Nyquist formula:

$$E_{n,A} = \sqrt{4kT_A R_A B_n}, \quad (5.3)$$

where B_n is the noise band, which is approximately equal to the pass-band of the radio front end, and $k = 1.38 \cdot 10^{-23}$ [J/K·Hz] is the Boltzmann constant. For a given R_A and B_n , the antenna noise is determined only by the value $T_A = t_A T_0$, which is the antenna noise temperature, where $t_A = T_A / T_0$ is the relative antenna noise temperature, $T_0 = 293$ K.

The antenna noise power entering in the matched input of the receiver ($R_A = R_{in}$) is defined by the value of $E_{n,A}$ (5.3) and is:

$$P_{n,A} = E_{n,A}^2 / 2R_A = kT_A B_n = E_{n,A}^2 / 4R_{in} = P_{n,in,nom}. \quad (5.4)$$

If the condition of matching will be disturbed ($R_A \neq R_{in}$), the lesser input power will act in the receiver input:

$$P_{n,in} = b k T_A B_n, \quad (5.5)$$

where $b = P_{n,in} / P_{n,in,nom}$ is the *mismatch coefficient* depending on the ratio of resistances R_A and R_{in} : $b = (R_{gen} + R_{in})^2 / (4R_{gen}R_{in})$.

Noises produced in the receiver are reduced to the receiver input for comparison with the antenna noises, which is connected to the same point of the radio front end. Noise properties of the receiver are often estimated by the noise factor for some standard conditions: we consider that the receiver input is connected to the equivalent signal generator ($P_{s,in}$) matched in power, which internal resistance (R_A) has the room temperature. The noise power ($P_{n,in}$) generated at the receiver input impedance (R_{in}) by the thermal noise of the antenna, the rms EMF of which is calculated in (5.3) for the matching mode with antenna resistance $R_A = R_{in}$, is determined by ratio (5.4).

Equation (5.5) does not take into consideration the contribution in the power level of the noise in the output of the radio front end caused by nonlinear distortions produced in limiters, a detector, and other nonlinear stages. The frequency converter, which is the voltage multiplier of signal frequency and the LO frequency, is considered as the linear device producing self-noises only for the input impact with the signal frequency.

The noise factor (N) of the two-port shows how much the noise power in the output of the real two-port exceeds the power in the ideal (non-noisy) two-port for the matched mode with the signal source:

$$N = P_{n,out} / P_{n,out,id} = 1 + P_{n,own} / P_{n,out,id} = 1 + P_{n,own} / k T K_p B_n \quad (5.6)$$

where $P_{n,out,id} = k T K_p B_n$ is the noise power in the output of the ideal two-port with the transfer function K_p and $P_{n,own}$ is the power of the own noise produced inside of the two-port itself. As we see from (5.5), the noise power in the output of the real two-port exceeds the noise power in its input, i.e., the noise factor for the real two-port is always more than 1.

If to use Equation (5.1) and take into account that and the absence of a matching mode ($b \neq 1$, (5.5)), then for the room temperature ($T = T_0 = 290$ K), the value of the rms noise EMF generated in the antenna circuit is:

$$E_A = \sqrt{4k \cdot N \cdot B_n \cdot D} \quad (5.7)$$

and the value of the signal power is relatively:

$$P_A = k \cdot N \cdot T_0 \cdot B_n \cdot D \quad (5.8)$$

For the estimation of noisy properties of low-noise two-ports, we can often use the equivalent *noise temperature*, which shows how much we must increase the absolute temperature of the resistance (R) of the noise EMF source (Fig. 5.2a) compared to the normal (T_0), which is connected to the input of the non-noisy two-port in order to provide that the noise power at its output would be equal to the power of the real (noisy) two-port:

$$T_n = (N - 1)T_0 \quad (5.9)$$

The noise temperature of the ideal (non-noisy) two-port is equal to zero.

The *noise property of the radio front end* including (depending on the structure of the receiver) the different number of stages, is defined by the following relation (for instance, for the structural receiver circuit of the superheterodyne type with the single frequency conversion, see Fig. 3.7):

$$N_{RFE} = N_{IC} + \frac{N_{RFA}}{K_{pIC,nom}} + \frac{N_{FC} - 1}{K_{pIC,nom} K_{pRFA,nom}} + \frac{N_{IFA1} - 1}{K_{pIC,nom} K_{pRFA,nom} K_{pFC,nom}} + \dots \quad (5.10)$$

where N_{RFE} is the real noise factor of the receiver radio front end reduced to its input; N_{IC} , N_{RFA} , N_{FC} , and N_{IFA1} are noise factors of the input circuit (IC), RFA, the frequency converter (FC), and the first stage in the intermediate frequency amplifier (IFA1); and $K_{pIC,nom}$, $K_{pRFA,nom}$, $K_{pFC,nom}$, and $K_{pIFA1,nom}$ are nominal values of gains in power of the input circuit, RFA, the frequency converter, and the first stage of IFA. As follows from (5.10), the noise figure of the RF front-end, recalculated to input of the receiver, is mainly determined by its first stages. The noise figure N contribution of each stage connected behind the IC is reduced, due to its recalculation to the input, by dividing its N by the power transmission coefficient of the previous stages. As a result, the influence of the terms in (5.10) of the noise components of the amplifying stages of the IF path on the overall noise coefficient of the RF front-end can be neglected due to their smallness. Relation (5.10) is valid for a superheterodyne type receiver (Fig. 3.7) and shows how the influence of the strongest noise source in the RF front-end – the mixer – can be reduced. For this, between its input and the IC output, it is necessary to include an RF amplifier with a low noise figure (LNA), which has a sufficiently high gain K . Formula (5.10) is valid for any other types of structural schemes of the RF front-end (Figs. 3.10 and 3.19). To apply it for a specific structure, it is necessary to sequentially enter into an expression identical to (5.10), the indices (N and K_p) of real cascades of the radio path, starting with IC.

The expression for the noise factor (5.10), which shows the relative contribution in power of each of radio front-end stages in the total noise level in its output, is true for the case when the energy spectral density of the noise source can be considered as constant in the operation frequency range. At that, the signal source must provide the linear operation mode of the active element in the whole range of variations of its amplitude. In this connection, the transistor noise factor, which is given in the reference books, can be true only for the specific frequency region and for the measurement mode. For transistor operating conditions other than those specified in the manual, it is required to measure the noise figure or carry out simulations, for example, using equivalent transistor circuits.

For mobile communication systems, instead of the introduced definition of the noise figure (5.6), we can often apply another definition, depending only on the change in the SNR_{in} at input and output SNR_{out} of a two port. When the tuned antenna is used in the terminal operation, the *noise factor* is defined by the noise power (or the average spectral density of the noisy EMF) of the radio front end, reduced to the input of the receiver (SNR_{in}) to the (SNR_{out}) on its output:

$$N = \frac{(SNR)_{in}}{(SNR)_{out}} = \frac{(P_{s,in} / P_{n,in})}{(P_{s,out} / P_{n,out})} \quad (5.11)$$

where $P_{s,in}$ is the signal power in the input and $P_{n,in}$ is the average noise power produced by the resistance of the signal source R_A .

In (5.11), the noise generated by antenna, which is an external source, with respect to input of receiver $P_{n,in}$. The noise of the RF front-end, with the power gain $K_{p,rfe}$, and having its own noise, the power $P_{n,rfe}$, of which, recalculated to the input of receiver, will change the form, expressions (5.11), which will become:

$$N = \frac{P_{s,in} / P_{n,in}}{K_p P_{s,in} / (P_{n,rfs} + P_{n,in})} = 1 + \frac{P_{n,rfe}}{P_{n,in}}. \quad (5.12)$$

As it follows from (5.12), the equation for the signal power in the receiver input is:

$$P_{s,in} = P_{n,in} \cdot N \cdot (SNR)_{out} \quad (5.13)$$

The noise source having the uniform power density S_n in the reception band B_n will produce in the receiver input of the power:

$$P_{n,in} = S_n B_n / R_A, \quad (5.14)$$

at spectral density $S_n = 4kTR_A$, where k is the Boltzmann constant, T is the absolute temperature, and R_A is the antenna resistance (the signal source).

Then (5.13) takes a form:

$$P_{s,in} = P_{n,in} \cdot N \cdot (SNR)_{out} B_n / R_A \quad (5.15)$$

Having calculated the common logarithm from the obtained expression, we present the input signal power in decibels:

$$P_{s,in,dB} = (S_n / R_A)_{dB} + NF_{dB} + (SNR)_{dB} + 10 \lg(B_n), \quad (5.16)$$

where NF [dB] = $10 \lg(N)$. For the known value of the receiver input resistance R_{in} , the noise spectral density produced by an antenna is:

$$S_{n,in} = \left(\frac{R_{in}}{R_A + R_{in}} \right)^2 S_n. \quad (5.17)$$

For the *matched mode* $R_{in} = R_A$ and under the condition that the *noise impact* in the receiver input is defined by the *thermal noises of the antenna*, which has the resistance R_A , and taking (5.14) into account, we obtain:

$$S_{n,in} / R_A = \frac{1}{4} 4kTR_A \frac{1}{R_A} = kT. \quad (5.18)$$

For the room temperature ($T = T_0 = 290 \text{ K}^\circ$)

$$S_{n,in} / R_A = 10 \lg \left(\frac{kT_0}{1 \text{ mW}} \right) = -174 \text{ dBm} / \text{Hz}. \quad (5.19)$$

The sensitivity must be $P_A = -91 \text{ dBm}$ $\left(P_{s,in} = P_A = 10^{\frac{P_A[\text{dBm}]}{10} - 3} = 0.79 \text{ nW} \right)$.

The selectivity is the receiver's ability to distinguish the useful signal from spurious signals using the distinctive signs, which are peculiar to the interference (noise) and the signal. These include, for example, the direction of arrival of the signal and the direction of arrival of the parasitic effect, which allows them to be separated using directional antennas (*spatial selection*). Another distinguishing sign of the useful signal and the spurious signal is the time of its effect in the receiver input. The specific case of the *time selection* is the temporal duplexing, which is applied in mobile communication systems (TDD), when the time intervals of reception and transmission are spread in time: the signal reception occurs at the switched-off transmitter and vice versa. Differences between wanted signal and spurious signals, based on their differences in amplitudes, frequencies, and phases, underlie the application of *amplitude, phase, and frequency selection*.

In analog systems of radio communication and radio broadcasting, the *frequency selection* is widely used based on the utilization of selective active and passive circuits and filters to extract the useful signal from the range of operating frequencies and on the suppression of spurious frequencies. These methods are necessary for the construction of radio front ends of digital mobile communication systems. At that, we distinguish the single-signal selectivity and the real selectivity.

The *single-signal selectivity* is defined by the AFC of filters in the receiver front end *without taking into account the non-linear phenomena* caused by the effect of active components, diodes, varicaps, etc. at the impact of the single harmonic signal (useful or spurious) in the receiver input. The single-signal selectivity is qualitatively estimated by the ratio of the

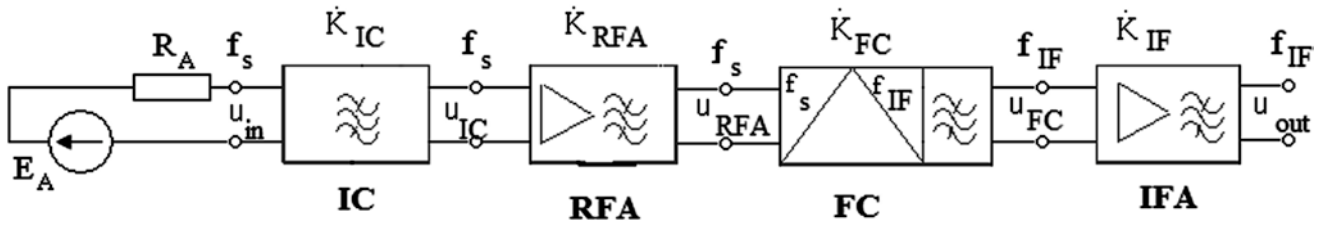


Fig. 5.3 The structural diagram of the superheterodyne radio front end

transfer function of the stage or the whole front end at action of the testing (useful) signal in the input to the transfer function on the interference frequency, which provides in the output the same voltage (without adjustment of radio front end):

$$Se_{int} = \frac{K_0}{K(f_{int})}, \quad (5.20)$$

where Se_{int} is the selectivity of the interference frequency, $K(f_{int})$ is the transfer function on the interference frequency, and K_0 is the transfer function on the central (resonance) frequency.

The amplitude-frequency characteristic of the radio front end of the receiver of superheterodyne type (Fig. 3.7) can be defined according to its equivalent circuit (Figure 5.3).

The gain of the radio front-end (RF front-end) on the arbitrary frequency is defined as:

$$\underline{K}_{rfe}(f) = \dot{U}_{out} / \dot{E}_A = \underline{K}_{IC} \underline{K}_{RFA} \underline{K}_{FC} \underline{K}_{IFA}, \quad (5.21)$$

where $\underline{K}_{Bx\&} = \dot{U}_{Bx\&} / \dot{E}_A$, $\underline{K}_{IC} = \dot{U}_{IC} / \dot{E}_A$, $\underline{K}_{RFA} = \dot{U}_{RFA} / \dot{U}_{IC}$, $\underline{K}_{FC} = \dot{U}_{FC} / \dot{U}_{RFA}$, $\underline{K}_{IFA} = \dot{U}_{out} / \dot{U}_{FC}$ are transfer (amplification) coefficients of separate stages: the input circuit together with an antenna, the RF amplifier, the frequency converter (usually together with the pass-band filter), and the IF amplifier.

For the central frequency of the main channel f_0 (for radio broadcasting, this frequency of the receiving signal is $f_s = f_{s0}$) and the average frequency of the converted signal $f_{conv} = f_{conv0} = f_{IF}$ at single conversion of the input signal (f_{s0}), the radio front-end gain takes a form:

$$K_{rfe0} = K_{0IC} K_{0RFA} K_{0FC} K_{0IFA}. \quad (5.22)$$

Gains of the stage or the RF front-end defined on the central frequency, are also a (relative) dimensionless quantity K_0 . When writing K_0 on a logarithmic scale (in decibels), they are expressed as: $K_{dB} = 20 \lg K$.

The normalized gain used to estimate selectivity and frequency distortion is ratio of the modulus of transfer function at an arbitrary (f) and the central frequency (f_0 are f_{IF}):

$$\gamma = \frac{\underline{K}_{rfe}(f)}{\underline{K}_{rfe0}} = \frac{\underline{K}_{IC}(f) \underline{K}_{RFA}(f) \underline{K}_{FC}(f) \underline{K}_{IFA}(f)}{\underline{K}_{0IC} \underline{K}_{0RFA} \underline{K}_{0FC} \underline{K}_{0IFA}} = \gamma_{IC} \gamma_{RFA} \gamma_{FC} \gamma_{IFA}. \quad (5.23)$$

To eliminate the dependence of the normalized gain versus the absolute value of the frequency f , we can use the dependence of γ from the offset (Δf) with regard to the central frequency: $\Delta f = f - f_0$, which allows the estimation of AFC of each block independently on the central frequency value.

Equation (5.23) defines the gain of the radio front end normalized to the maximal gain of each stage and the central frequency of the receiving signal. Gains of preselector stages have a somewhat larger bandwidth compared with FC and IFA stages (Fig. 5.4).

Since the resulting normalized gain γ_{rfe} is a product of stage gains, it is closed to AFC of the IF section in the shape.

By definition (5.20), the single-signal selectivity is the quantity inversed to the normalized gain (5.23); therefore, the selectivity on the interference frequency S_{int} is defined as:

$$Se_{int} = \frac{0. rfe}{rfe(f_{int})} = \frac{0. IC K_{0RFA} K_{0FC} K_{0IFA}}{IC(f_{int}) RFA(f_{int}) FC(f_{int}) IFA(f_{int})} = Se_{IC} Se_{RFA} Se_{FC} Se_{IFA}. \quad (5.24)$$

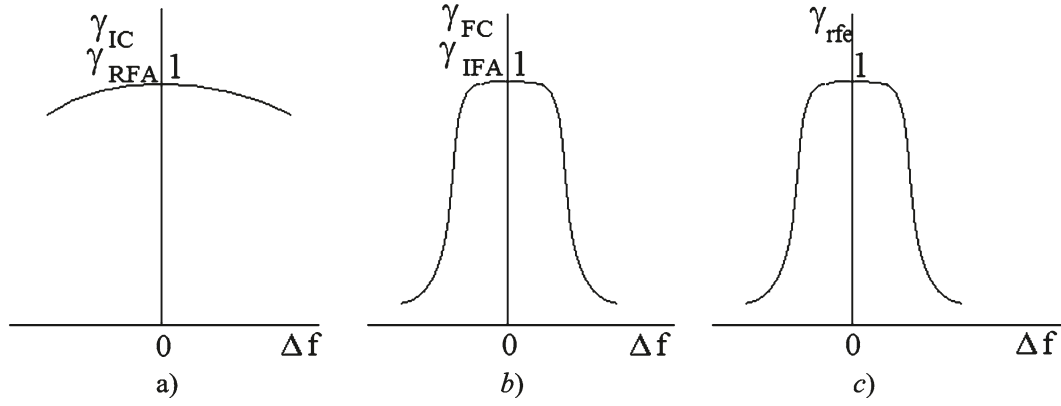


Fig. 5.4 AFCs of the radio front-end stages

As we see from (5.24), the resulting selectivity is defined by the product of selectivities provided by the separate stages. It is clear from a comparison of the radio front-end AFC (5.23) with Eq. (5.24) that selectivity defines the suppression degree of interferences, which has the frequency offset of Δf from the central frequency, and this offset is larger for the more abrupt AFC of the front end.

The single-signal selectivity characterizes the effect to the reception accuracy of the narrowband interferences in the *adjacent channel*, which are peculiar to *all types of receivers* and are produced by the external (adjacent) radio stations with the frequency offset $\Delta f_{adj. ch}$. For analog radio broadcasting AM receivers, $\Delta f_{adj. ch} = \pm 9$ kHz, and for receivers of FM signals, $\Delta f_{adj. ch} = \pm 150$ kHz. For receivers of mobile communication systems, the position of the adjacent channel is defined by standards for the appropriate equipment (Table 1.2) or is chosen as equal to the frequency bandwidth allocated to the radio channel. In superheterodyne receivers, it is technically easier to ensure the maximum slope of frequency response of band-pass filter outside the transparency band when choosing a low intermediate frequency allowed by standards. When using double (Fig. 3.14) or triple conversion of the frequency of the received signal, the most effective suppression of the adjacent channel is achieved at the lowest intermediate frequency. Suppression of adjacent channel in digital signal receivers with conversion to a low intermediate frequency (Fig. 3.24) is performed in two stages: complex filter extracts two orthogonal signals containing a small group of channels, including the useful channel; after their amplification, digitization, and conversion in mixers, LPF at their outputs, the quadrature components of signal with the modulation frequency are isolated and adjacent channel is suppressed. In the direct conversion receivers (Fig. 3.15), the input circuit (IC) and RFA cannot provide the adjacent channel suppression (Fig. 5.4a) up to a value ensuring the required probability of the correct reception for a given SNR. In this case, the main filtering is provided by the connection of the analog discrete filters in the detector output or in the informational (digital) section with the help of digital filters realized in the integral technology.

The *inclusion of the mixer* in the RF front-end of the superheterodyne receiver forms a huge number of spectral components in its output current, which are the result of multiplication of any influences at its input with the voltage of first and other harmonics of the local oscillator frequency. Among the input influences, interference (*collateral channels*) with a (quasi-harmonic) narrow-band spectrum, which form spectrum components at an intermediate frequency in the mixer current, is especially dangerous. The most interfering effect on the reception of useful signal is *mirror channel* and the *direct passage channel*.

The *image (symmetric) channel* is away by $\Delta f_{im. ch} = 2f_{IF}$ from the frequency of the received signal and can be higher than the frequency of the receiving signal f_s , if we use the upper tuning (Fig. 3.3) of the local oscillator ($f_{Lo} > f_s$), or lower than f_s , if we use the lower tuning ($f_{Lo} < f_s$).

The *direct passage channel* (as the image channel) is caused by addition of the mixer in the front end, which the transfer function on the intermediate frequency f_{IF} is, at least, two times more than the transfer function in the conversion mode. This leads to beat appearance between the converted useful signal and the signal in the direct passage channel and its harmonics.

The Squareness Coefficient. Suppression effectiveness of the collateral channels of reception and conversion is characterized by the slope of AFC decline (in the front end), which is estimated by the squareness coefficient (Fig. 5.5).

Fig. 5.5 To definition of the squareness coefficient

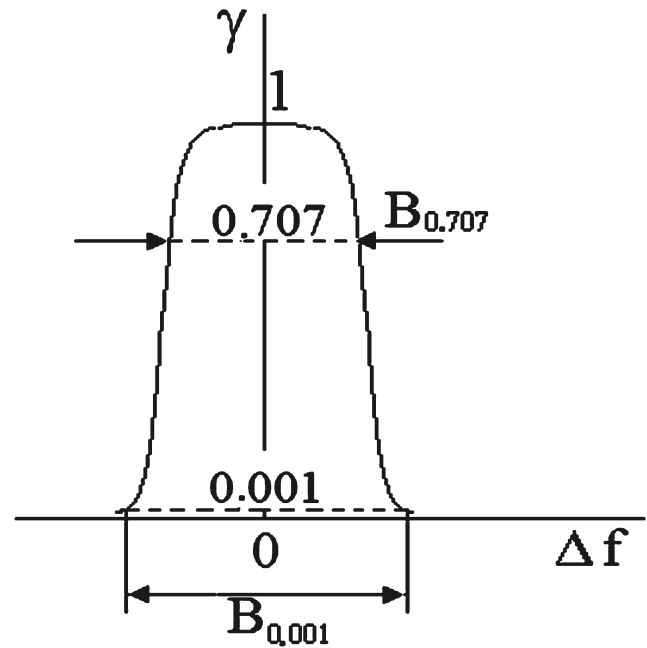
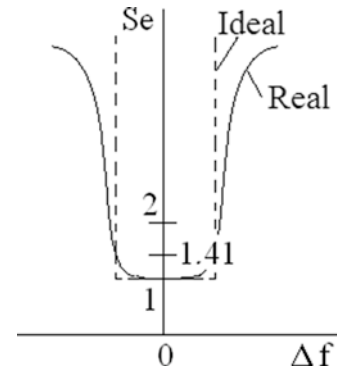


Fig. 5.6 The single-signal selectivity versus the frequency offset



The squareness coefficient k_{sq} equals to the ratio of the pass-band of the specific stage or the radio front end usually defined on the -3 dB level ($\gamma = 0.707$) of the normalized value (Figure 5.5) of AFC ($B_{0.707}$) to the pass-band ($B_{0.001}$) on the level of -60 dB ($\gamma = 0.001$):

$$k_{sq} = B_{0.001} / B_{0.707}.$$

As it follows from the comparison of (5.22) and (5.23), $Se = 1/\gamma$, and therefore, the closer k_{sq} to 1, the higher AFC coincidence of the specific stage to the square shape and the larger (dotted curve in Fig. 5.6) single-signal selectivity.

Blocking is the receiver's capability to receive the modulated useful signal in the presence of the spurious signal on frequencies differing from the collateral reception channels or the adjacent channels with errors, which do not exceed specified limits.

If the useful signal level is 3 dB higher than the sensitivity level and interference level, which are defined, for example, by the specification in [5, 6], the error probability RBER (residual bit error ratio) in the TCH/FS channels in static propagation conditions should not exceed 2.439%.

The **transfer characteristic** of radio receiver is the time function of the signal voltage envelope at its output when exposed to a radio pulse with an amplitude EA (amplitude-shift keyed signal envelope amplitude) at its input. The difference of signal envelope from the square shape indicates the appearance of *frequency* distortions. The appearance of the pulse *edge time* (rise or drop) reflects the suppression of the high-frequency components caused by the limited bandwidth of the RF front-end. Appearance of the *peak drop* of the rectangular pulse of the output signal is caused by the level decrease of low-frequency

components of the converted signal spectrum in the region of low frequencies. The time delay occurrence of the output shows the appearance of phase shifts between spectral components.

The amplitude characteristic is the function of the voltage amplitude (or power) on the receiver output versus the amplitude of harmonic EMF (or the power level) in an antenna.

The output power depending on the receiver type (e.g., for carrying radio broadcasting receivers) is defined as the *nominal*, which is the maximal power when the nonlinear distortions (for the 100% modulation) do not exceed the given norm; the *normal* corresponding to 10% from nominal measured at 30% modulation depth; and the *standard*, which is the output power of 50 mW (the specific values of indices are defined by standards on the appropriate equipment).

Operating frequency band is the frequency interval, within the limits of which, the receiver at retuning saves its main characteristics. This range is defined by the covering coefficient:

$$k_{cov} = f_{max} / f_{min}, \quad (5.25)$$

where f_{max} and f_{min} are the maximal and minimal frequencies of the receiver operation range. In professional receivers, $k_{cov} > 100$. Besides the wideband receivers, the receivers with fixed tuning are applied. In most cases, receivers of mobile communication systems are concerned to the narrowband ones ($k_{cov} \approx 1$), which is caused by the high operating frequency range.

The accuracy of **frequency tuning** is one of the parameters which defines a possibility to provide prolonged signal reception without degradation of the quality of reproducing messages and without any manual adjustments.

The frequency accuracy is defined, for instance, for receiving signal with the frequency f_s and the frequency of receiver tuning f_0 and is characterized by the relative detuning:

$$(f_0 - f_s) / f_s = \pm \Delta f / f_s. \quad (5.26)$$

Instability of receiver tuning is defined by the impact of a series of destabilizing factors (variations of temperature, atmospheric pressure, moisture, etc.). The main reason of frequency instability is the variation of local oscillator frequencies causing the variation of intermediate frequencies and hence frequency distortions causing nonlinear distortions due to the amplitude-phase conversion.

In radio broadcasting low-quality receivers (the third group of complexity), the main contribution in the resulting instability is introduced by the instability of local oscillator frequency, which is $\delta f_{LO} = \Delta f_{LO} / f_{LO} = 10^{-2} \dots 10^{-3}$, while the relative frequency instability in the receiving signal is $\delta f_s = 10^{-6} \dots 10^{-7}$. The high stability of the frequency of the reference oscillators and the accuracy of tuning to the operating frequency increase the reliability of the received message and efficiency of the channel in conditions of its high load. In modern professional receivers of the HF and VHF bands and MS digital cellular systems, the relative frequency instability of reference oscillators stabilized by quartz is $10 \dots 10^{-10}$. BS receivers using a thermostatically controlled reference oscillator and frequency synthesizers, the relative frequency instability of voltage-controlled oscillators is reduced by an order of magnitude. In navigation satellite systems, instability of the on-board synchronizing device decreases to $10 \dots 10^{-10}$.

The Dynamic Range. The *dynamic range* is the integral characteristic of the radio receiver, which connects the sensitivity, selectivity, and nonlinear distortions of the radio front end. Dynamic range on the main channel is $D = 20 \lg (E_{A, adm} / E_{A0})$, where $E_{A, adm}$ is the level of the signal generated by the antenna at input of the receiver, limited by the admitted nonlinear distortion at the RF front-end, and E_{A0} is the minimum level of the input signal created by its own noise. This indicator characterizes the limits of the input signal level change at which the permissible loss of information contained in useful signal is ensured. The minimal level E_{A0} is defined by the level of self-noises. The dynamic range of modern receivers can achieve 100–120 dB, and the value of the admitted distortion level of the input signal is estimated by the amplitude characteristic $U_{out} = f(E_A)$. Nonlinear distortions can be arisen not only due to the high level of the useful signal but owing to the impact of the strong out-of-band interference.

Nonlinear distortions on the *main signal* are defined by the harmonic coefficient of the modulating signal:

$$k_{har} = \sqrt{U_{out2}^2 + U_{out3}^2 + U_{out4}^2 + \dots} / U_{out1}, \quad (5.27)$$

where U_{out1-4} are effective voltage values of appropriate harmonics of the modulation frequency F_M . Harmonics of the modulating frequency appear in stages with the large level of the input signal, which leads to its limitation (the operation point transfer to the nonlinear part of the amplitude characteristic).

In digital systems of the mobile communication, the digital stream from the demodulator output is fed to the input of the amplifier with the logarithmic amplitude characteristic, which “equalizes” the amplitudes of the output signal. This allows obtaining the informational stream with the practically constant amplitude, which passes into the DAC input and forms the analog signal.

The probability of error appearance BER (the coefficient of error bits) is the main criterion of performance of digital communication systems:

$$\text{Bit Error Rate} = \frac{\text{number of received error bits}}{\text{total number of transmitted bits}} \quad (5.28)$$

When transmitting data over channel with additive Gaussian noise, the error rate depends both on properties of the channel and on SNR ratio, as well as on the selected processing technology of the received signal. The model of the additive white Gaussian noise (AWGN) assumes the thermal noise action in the receiver input with the constant *power spectral density* N_0 . For the *energy suited to one bit* of information in the receiver input E_b and for the power spectral density for AWGN N_0 , the required ratio is:

$$\frac{E_b}{N_0} = \frac{T_b P_{in}}{N_0} = \frac{1}{C} \frac{P_{in}}{N_0}. \quad (5.29)$$

The formula (5.29) allows the estimation of the power density level P_{in} in the receiver input at the noise power density N_0 [16], which provides the required ratio $(E_b/N_0)_{req}$ at the information stream rate c [bit/s]:

$$\frac{P_{in}}{N_0} = C \left(\frac{E_b}{N_0} \right)_{req} \quad (5.30)$$

Using (5.30), we can define the maximal possible rate of the informational stream for known other parameters: E_b, N_0, P_{in} .

5.2 Internal Noise Sources and Their Models

When designing radio receivers for a given SNR at its input, it is necessary to take into account not only the noise properties of transistors as a main source of noise, depending on the DC operating mode of the transistor, but also the bandwidth of operating frequencies. The transition to work in the microwave range increases the signal loss in the radio channel and, accordingly, increases the requirements for the receiver sensitivity. In some cases, this can lead to a transition from the matching mode in terms of the input power of the receiver with the antenna to the implementation of the matching mode with the antenna at a minimum of noise.

5.2.1 Methods of the Noise Property Description of Sources

There are a series of methods for the analysis and description of noise properties of radio front ends.

At present, the parameter called the *noise factor NF* is often used for it. This indicator is determined by the power of thermal noise generated by the equivalent noise impedance of an antenna with a uniform spectral density in the noise bandwidth of the RF front-end. The cascades of the radio RF front-end can be described in a similar way, replacing the transistors with equivalent circuits and displaying the actually acting noise, with equivalent noise sources included in the place of their occurrence. Such a characteristic (together with the parameter “noise temperature”) is often used also for the description of noise properties of transistors in the region of their application.

Another method of the noise property description of the linear circuits is based on the representation of the noisy two-port by two *equivalent noise sources* connected usually in its input (Fig. 5.7).

A real transistor, which creates noises of a different nature at its terminals, due to the properties of the processes occurring inside its structure, is replaced by equivalent two-port circuits: noise sources (EMF source $u_n(t)$ and current source $i_n(t)$) and an ideal noiseless transistor. Noise sources $u_n(t)$ and $i_n(t)$ are shown by shading.

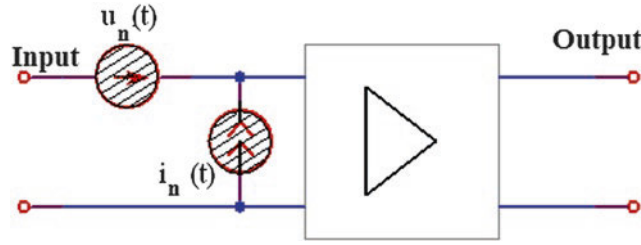


Fig. 5.7 Representation of the noisy two-port by two equivalent noise sources

Since noises represent the stochastic process, for their description [15], the average squares of noise voltages $|\bar{u}_n(f)|^2$ are used, which are called the *spectral density* of voltage noise $S_u(f)$ or the spectral density of the current noise $|\bar{i}_n(f)|^2 = S_i(f)$, having the dimensions $[V^2/Hz]$ and $[A^2/Hz]$, relatively. *Noise densities allow the definition of the spectral distribution of effective values of $u_{n,eff}$ and $i_{n,eff}$ describing the noise density as a function of the angular frequency ω with the definition domain $\omega \in (-\infty, +\infty)$ instead of the frequency f with the definition domain $f \in [0, \infty)$. Effective values can be found according to the noise density as:*

$$u_{n,eff} = \sqrt{\int_0^\infty |\bar{u}_n(f)|^2 df}, \quad (5.31)$$

$$i_{n,eff} = \sqrt{\int_0^\infty |\bar{i}_n(f)|^2 df}, \quad (5.32)$$

which are related to the spectral density of the appropriate noise sources as:

$$|\bar{u}_n(f)|^2 = \frac{d(u_{n,eff}^2)}{df}, \quad (5.33)$$

$$|\bar{i}_n(f)|^2 = \frac{d(i_{n,eff}^2)}{df}. \quad (5.34)$$

At transfer to the description of noise process properties through the angular frequency, we must take into consideration the connection between spectral densities, which can be expressed as:

$$|\bar{u}_n(f)|^2 = 4\pi |\bar{u}_n(j\omega)|^2. \quad (5.35)$$

In the general case, sources $u_n(t)$ and $i_n(t)$ are correlated between themselves; therefore, for complete description of the stage (radio front end), according to the method of noisy two-port, it is necessary to have a knowledge not only of spectral densities of mentioned sources $S_u(f)$ and $S_i(f)$ but also distributions characterizing their mutual influence on each frequency $S_{ui}(f)$ and $S_{iu}(f)$.

Such a description of properties of stages, units, and the radio front end as a whole allows obtaining their noise characteristics independently on the approach to include active elements and the method of the specific block construction. Nevertheless, the practical application of this method is possible only in the case of known distribution of spectral densities of sources $S_u(f)$ or $S_i(f)$, as well as their interconnections $S_{ui}(f)$ and $S_{iu}(f)$, which are rather difficult to determine. Besides, even at the absence of correlation between the noise sources inside the two-port, spectral densities and parameters defining their interconnection can have the frequency dependence. The description of the noise properties with the help of spectral densities $S_u(f)$ or $S_i(f)$ is complicated by the fact that they are functions of operation modes of transistors and the external temperature. Therefore, the described approach is used for the analysis of noise properties of complex structures: ICs, op-amps, etc.

Models of Noise Sources The method based on the utilization of physical equivalent circuits has got the widest application, especially for the analysis of noise properties in the wide frequency band.

Usually, all main internal physical noise sources are non-correlated, and the resulting noise power in the output $P_{n.out,\Sigma}$ is the sum of their powers:

$$P_{n.out,\Sigma} = P_{n.out,1} + P_{n.out,2} + P_{n.out,3} + \dots + P_{n.out,m} = \sum_{j=1}^m P_{n.out,j}, \quad (5.36)$$

and dispersions:

$$\sigma_{out,\Sigma}^2 = \sigma_{out,1}^2 + \sigma_{out,2}^2 + \sigma_{out,3}^2 + \dots + \sigma_{out,m}^2 = \sum_{j=1}^m \sigma_j^2, \quad (5.37)$$

where $P_{n.out,j}$ is the noise power in the output produced by the j th noise source, $\sigma_{out,j}$ is the effective value of the noise voltage in the output produced by the j th noise source, and m is the total number of noise sources affected on the resulting output power (dispersion).

The calculation of the noise characteristics of the RF front-end based on the use of physical equivalent circuits turns out to be very effective, since with a limited number of stages affecting its noise characteristics and neglecting the mutual correlation of the noise sources, we can also:

- To decrease the total number of noise sources estimating the degree of their effect on the resulting noise indices
- To be limited by the contribution estimation into the output noise process from only the input circuit and the low-noise input amplifier for the radio signal (or the frequency converter)
- Not to take into account of the frequency dependence of the noise spectral density $S_u(f)$ and $S_i(f)$ versus frequency considering that produced noise is white

Noise processes in the two-port are divided into **three types**:

- (1) Frequency-dependent noises with the current-independent spectral densities S_u and S_i
- (2) Frequency-independent noises with the current-dependent spectral densities $S_u(I)$ and $S_i(I)$
- (3) Frequency-dependent noises with the current-dependent spectral densities $S_u(f, I)$ and $S_i(f, I)$

Frequency-Independent Noises with Current-Independent Spectral Densities The *thermal noises* (Nyquist noises), which occur due to the chaotic motion of charge carriers in circuit components having a resistance and being at temperature differing from the absolute zero, can be attributed to noises of the **first type**. This motion of charge carriers produces the noisy EMF at the terminal of the one-port device, which leads to current appearance in the external circuit. Nyquist showed that the resistor R , which is in thermodynamic equilibrium with the environment at a temperature T , can be regarded as a source of fluctuation currents in external circuit. The reason for this is a chaotic change in the number of charge carriers at its terminals, which create a potential difference: the equivalent of an electric voltage (or current) generator. In this case, spectral densities of noise sources are frequency dependent in the whole radio range and are defined as:

$$S_u = 4kTR, V^2 / Hz \quad (5.38a)$$

$$S_i = 4kT / R, A^2 / Hz. \quad (5.38b)$$

In the general case of a two-terminal (one-port component) with a complex resistance $Z(f)$ (or conductance $Y(f)$), the spectral current density or EMF generated by such sources can be represented by relations (5.39) and are mutually recalculated depending on the used equivalent schemes like::

$$S_i(f) = S_u(f) / Z^2(f) \quad (5.39a)$$

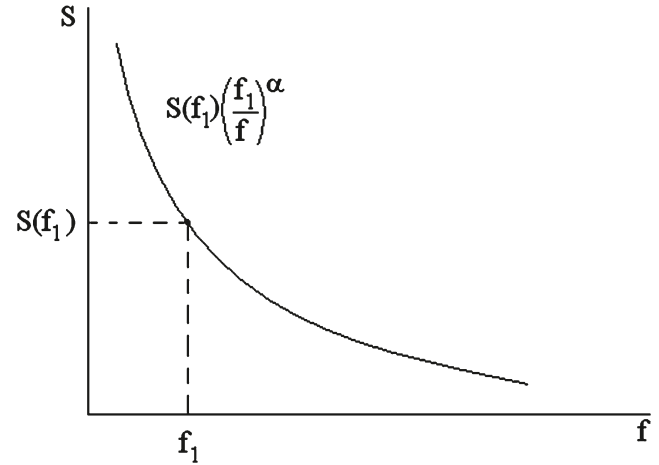
$$S_u(f) = S_i(f) / Y^2(f), \quad (5.39b)$$

where $Z(f) = 1 / Y(f)$ is the modulus of the complex resistance of the one-port network.

The frequency dependence of S_u and S_i in the description of these generators appears only in the infrared frequency range.

Frequency-Independent Noises with the Current-Dependent Spectral Densities The shot noises can be an example of the second type of noises. The *shot noise* arising at motion of charge carriers through p - n junctions is caused by carrier discreteness (electron-hole structure) and fluctuation of the number of carriers, which overcome the potential barrier in the p - n

Fig. 5.8 The flicker noise as a function of the frequency



junction during time unit. Application of the circuit modeling system MicroCAP to determine the noise properties of the radio front end allows taking into consideration the influence of active element modes in DC and variations of the external environment temperature.

The noise current arising at these processes has a uniform (frequency-independent) energy spectrum, which is proportional to the mean current value flowing through the p - n junction:

$$S_i = 2qI, \quad (5.40)$$

where I is the mean current value, which flows through the p - n junction, and $q = 1.6 \cdot 10^{-19}$ [Q] is the electron charge. It follows from this that the second type of noises has different values of spectral density of the noise current depending on the transistor mode in DC.

Frequency-Dependent Noises with Current-Dependent Spectral Densities The *low-frequency redundant* noise caused in the electronic tubes by the blinking (flicker) effect of the emitting electrode (in special literature – the flicker noise) is attributed to the **third** sort of the noise. In bipolar transistors, flicker noise is caused by the presence of traps, surface states, and other defects in the crystal lattice of the semiconductor structure, which leads to random generation and recombination of electric charge carriers. This noise plays a noticeable role in the low-frequency region, because it has the frequency dependence of the spectral density, which is proportional to $1/f^\alpha$ (Figure 5.8), where α is a parameter close to 1.

The graph of the spectral density of redundant noise obeys the law:

$$S(f) = S(f_1)\left(f_1 / f\right)^\alpha, \quad (5.41)$$

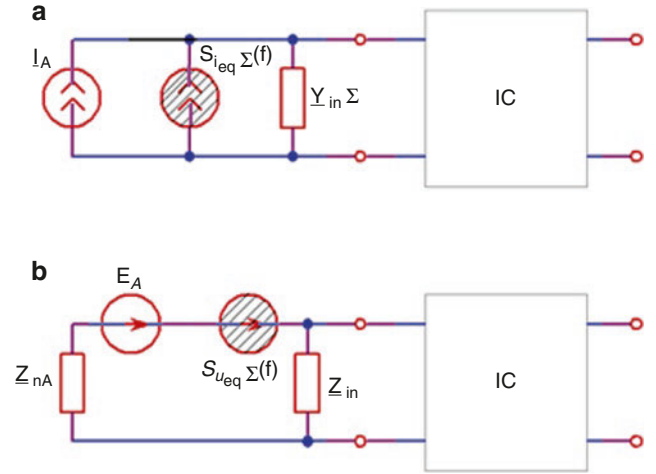
where $S(f_1)$ is the spectral density value of the redundant noise in the frequency f_1 ; the frequency f_1 is the frequency, relatively to which we perform the normalization of the spectral density $M(f)$:

$$M(f) = S(f) / S(f_1) = (f / f_1)^\alpha. \quad (5.42)$$

Usually, the normalization is performed at the fulfillment of equality of redundant noise spectral density and the spectral density of some other frequency-independent noise source, for instance, the shot source defined by the base current I_{b0} .

Changes during operation of the resistance of the pull-down resistors used to provide mode in transistor and semiconductor of integrated circuit structures can generate excess noise. Spectral densities of $S_u(f, I)$ and $S_i(f, I)$ fluctuations caused by these processes are proportional to the square of the potential difference, which is applied to the fluctuating resistor and has the frequency dependence defined by the $1/f$ law. We can indirectly estimate the presence and the effect of such type of noises on the base of the parameter (which is mentioned in the reference book) called the integral noise. These integral indicators include the effective value of the noise voltage σ_u or current σ_i , which is determined at a potential difference across the 1 V resistor in frequency band set by the upper and lower cutoff frequencies $\Delta f = f_{\max} - f_{\min}$.

Fig. 5.9 The input circuit together with the noise parameters in radio front-end blocks reduced to the receiver input and with noise parameters of an antenna



The additional reason of the redundant noises' origin is the insufficient filtering in the power supply circuits (Zener diodes) and in circuits of reference voltage formation (the local oscillator). The noise generated by these reasons has spectral densities obeying the $1/f^2$ law in the low frequency region.

In spite of the low-frequency character of the redundant noise spectrum (Fig. 5.8), it can introduce additional noises into the high-frequency signals, for instance, due to nonlinear properties of active elements in preselectors, which are caused by the small dynamic range of the applied transistor, or due to the modulation of the signal current by the fluctuating resistor in the input circuit. Such noises exert a significant effect on receiver sensitivity at application of the architecture of direct transformations.

For noises with a spectrum of the $1/f$ type, the effective value of the voltage σ_u or the current σ_i is defined by the ratio f_{max}/f_{min} independently on the pass-band Δf . For the spectrum of the noise described by (5.41) at $\alpha = 1$, the dispersion of the noise voltage is defined as:

$$\sigma_u^2 = \int_{f_{lo}}^{f_{up}} S_u(f_1) \frac{1}{f} df = S_u(f_1) f_1 \ln(f_{max} / f_{min}). \quad (5.43)$$

The use of effective values of voltages and currents (σ_u, σ_i) instead of spectral densities (S_u, S_i) for assessing the properties of noise is preferable, which allows the use of conventional units of measurement (volts and amperes) for assessing the intensity of noise sources. For example, as seen from (5.43), a effective voltage of the flicker-noise at the output of an ideal band-pass filter with a bandwidth equal to a decade ($f_{max} / f_{min} = 2$) remains constant.

5.2.2 Noises at the Receiver Input

The signal at the input of the receiver, created by the antenna by a multitude of quasi-harmonic influences with different amplitude, frequency, and phase, can be considered as random process acting at the input of the preselector (input circuit) and having a spectral density $S_{u,eq,\Sigma}$ or $S_{i,eq,\Sigma}$. Such a replacement of several sources of noisy impacts onto the single equivalent noise source in the receiver input is possible at known parameters of the antenna Z_{nA} . The input circuit together with the noise parameters in radio front-end blocks reduced to the receiver input and with noise parameters of an antenna can be represented as shown in Fig. 5.9.

In this figure, $Y_{in\Sigma} = Y_{nA} + Y_{in}$ is the resulting conductance defined by the conductance on the input circuit Y_{in} , $Y_{nA} = 1/Z_{nA}$ is the antenna conductance, and I_A is the current source produced by an antenna. The noise source acting in the radio front-end input with the spectral density $S_{eq,\Sigma}$ is equal to the sum of energy spectra of equivalent noise sources, which is caused by the superposition principle:

$$S_{eq,\Sigma}(f) = \sum_{j=1}^m S_{eq,j}(f), \quad (5.44)$$

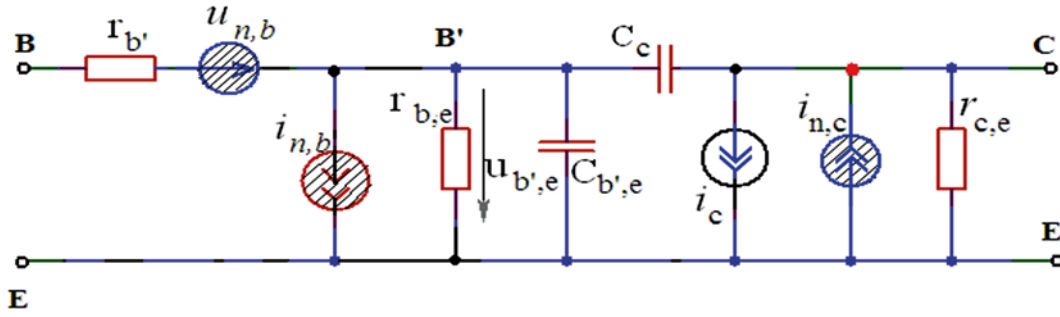


Fig. 5.10 The BT model in the mode of small signals with the sources of noise base voltage and noise currents of the base and the collector

where $S_{eq,j}$ is the noise spectral density from the j th equivalent source and m is the total number of noise sources causing the noticeable contribution in the total noise process.

The resulting conductance $Y_{in\Sigma} = g_{in\Sigma} + j\omega C_{in\Sigma}$ is defined by IC parameters reduced to the receiver input and depending on properties of input conductance of the first stage and matching circuits as well as the antenna distributed parameters. The reactive component in the input conductance is equivalent to a capacitance, which value is defined by not only the static capacitance between input terminals of the active element but the gain (the Miller effect [35]).

It is expedient to convert the internal noise sources into equivalent ones acting at the input of the RF front-end, through the gains of the blocks in terms of voltage, since when calculating the input stage, we usually indicate the sensitivity (EMF in antenna) at a given SNR at the detector input. The process of spectrum transformation of the j th source prototype $S_{u,j}(f)$ and $S_{i,j}(f)$ into the equivalent spectrum $S_{u,eq,j}(f)$ and $S_{i,eq,j}(f)$ in the case of resulting energy spectrum representation by the equivalent current source (Fig. 5.9a) gives:

$$S_{i,eq,j}(f) = S_{i,j}(f) \left[K_j(f) / K_\Sigma(f) \right]^2 \left[Y_{in\Sigma}(f) / Y_j(f) \right]^2$$

$$S_{u,eq,j}(f) = S_{u,j}(f) \left[K_j(f) / K_\Sigma(f) \right]^2 Y_{in\Sigma}^2(f),$$

where $K_j(f)$ is the modulus of the transfer function on voltage from terminals where the considered j -source is connected to the front-end input; $K_\Sigma(f)$ is the modulus of the transfer function on voltage of the whole front end; $Y_j(f)$ is the modulus of full circuit conductance shunting terminals, to which the considered j -source is connected; and $Y_{in\Sigma}(f)$ is the modulus of the full circuit conductance in the front-end input. Utilization of the current equivalent as the noise source simplifies the recalculation of internal noise current sources and appropriate conductances as parallel-connected components ($Z_{nA} = 1/Y_{nA}$).

Noises in Bipolar Transistors In the structure of semiconductors and in the p - n junction of the bipolar transistors (BT), noise voltages and currents appear. The thermal motion of charge carriers and fluctuations of the electric current at passage of discrete charge carriers through the junction are their sources.

Noises in BT in the radio range contain three components: thermal noises of the body resistance of the base, emitter, and collector regions; fluctuations of charge carriers passing through the emitter and collector p - n junctions (the shot noise); and fluctuations of collector and base currents caused by the random process of recombination of charge carriers (injected into the base from the emitter) with free charge carriers of the opposite signs. Figure 5.10 shows the BT model in the mode of small signals with the sources of noise base voltage ($u_{n,b}$) and noise currents of the base ($i_{n,b}$) and the collector ($i_{n,c}$).

Thermal noises are mainly defined by noises of the base region $|u_{n,b}(f)|^2 = 4kTr_{b'}$, and we can neglect other noises due to their relative smallness. The shot noise of the base current is defined by the relation:

$$|i_{n,b}(f)|^2 = 2qI_{b0} + \frac{k_{\beta} I_{b0}^{\gamma_{\beta}}}{f},$$

where I_{b0} is the base current in the operation point, k_{β} and γ_{β} are empirically obtained coefficients equal to 1–2 [7, 8], and $q = 1.602 \cdot 10^{-19}$ [K] is the electron charge.

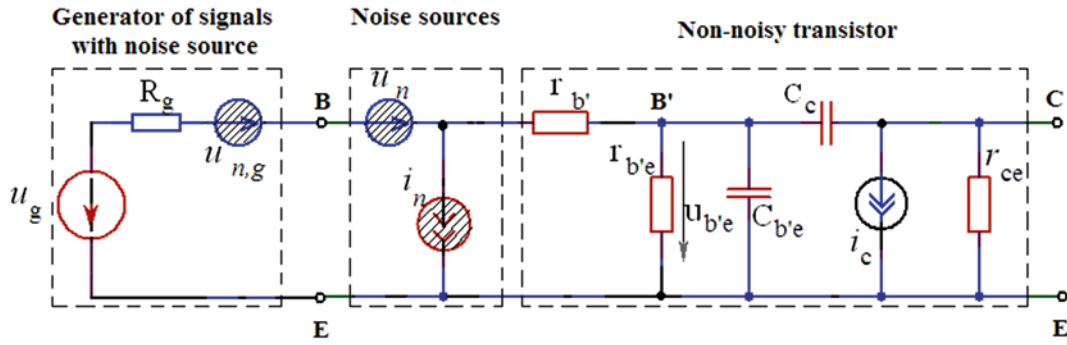


Fig. 5.11 The small-signal equivalent circuit of the bipolar transistor

The collector shot noise is described by the similar relation:

$$\overline{i_{n,c}^2(f)} = 2qI_{c0} + \frac{k_{fl} I_{c0}^{\gamma_{fl}}}{f}.$$

The flicker noise predominates in the shot noise on low frequencies, but on middle and high frequencies, the thermal noise components predominate. The frequency on which these components are equal is called the cutoff frequency on the current noise $f_{cut,fl}$:

$$f_{cut,fl} = \frac{k_{fl}}{2q}.$$

Its value depends on the operating point choice and increases with I_{c0} growth. For low-noise transistors, $f_{cut,fl}$ is 10 Hz–100 kHz at $\gamma_{fl} = 1, 2$.

Recalculating the noise sources in BT circuit (Fig. 5.10) to E and B terminals, which makes the transistor as noiseless, and thus simplifying the calculation, we represent the small-signal equivalent circuit of BT in the form shown in Fig. 5.11.

The input signal source u_g having the internal resistance R_g at the same time is the source of thermal noises produced by the nonideal generator. The transistor noises represented in the transistor input by the source of noise voltage u_n and the noise current i_n are equivalent sources (Fig. 5.10) of the noise voltage ($u_{n,b}$) and of noise currents ($i_{n,b}$, $i_{n,c}$) with extraction of the ideal non-noisy BT. Reducing all sources of the noise voltage and current to the single equivalent source of the noise voltage and assuming that it belongs to the signal generator, noise transistor properties can be described by the *spectral noise factor*, which is a ratio of the noise spectral density of the equivalent source $u_{n,eq}$ to the noise density of the signal generator $u_{n,g}$:

$$N(f) = \frac{|u_{n,eq}(f)|^2}{|u_{n,g}(f)|^2}.$$

Figure 5.12 shows the character of spectral noise factor dependence for some bipolar transistors [35] with the static gain of the base current in the common emitter (CE) circuit $\beta = h_{21e} = 100$, $r_b = 60 \Omega$, the generator resistance $R_g = 1 \text{ k}\Omega$, the cutoff frequency of the current noise $f_{cut,fl} = 100 \text{ Hz}$, the transistor cutoff frequency $f_T = 100 \text{ MHz}$, and the collector current in the operating point $I_{c0} = 1 \text{ mA}$.

The graph shows that flicker noise dominates in the frequency band below $f_{gr,flick}$ which has a function $\approx 1/f$ in the frequency band $f_{gr,flick} < f < f_T \beta^{-1/2}$, noise factor is defined by independent on the frequency of spectral noise densities and then increases sharply $\approx f^2$.

Since the generator has its own “noisy” internal resistance R_{gen} , then the noise properties of the stage significantly depend on the ratio of R_{gen} and r_b , which defines the base (collector) current. At *high requirements to the noise level*, we must use transistors with low volumetric base resistance and the high gain β in current in the small-signal mode. Therefore, with a low internal resistance of the signal source R_{gen} , it is required to choose a load of BT with a low volume resistance of the base (r_b) and a high current gain (β); and with a high internal resistance of the generator R_{gen} – with a high current gain (β), in the absence of such BT, use transistors with a low volume base resistance r_b .

With the growth of the operating frequency, the mutual correlation of the noise sources is amplified, which causes troubles in the choice of the optimal generator resistance R_g .

Fig. 5.12 The character of spectral noise factor dependence

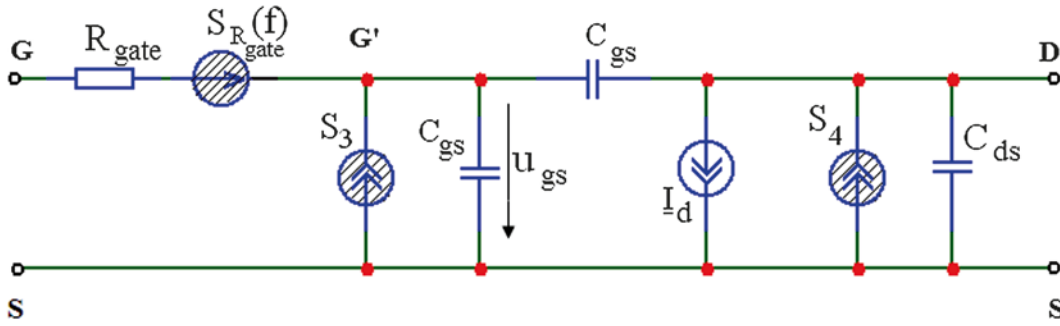
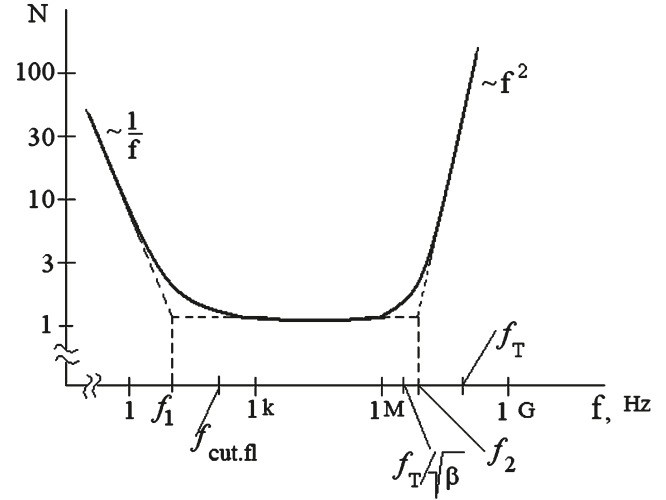


Fig. 5.13 The equivalent circuit of FET with noise sources

Noises in Field-Effect Transistors (FETs) Applying the similar approach as we used at consideration of BT noise properties to FETs, we are limited by consideration of the simplified description suitable for practical purposes.

For FET operating point I_{d0} , which corresponds to the A mode, the main noise sources in the saturation region are [35]:

- Thermal noises of the gate resistance R_{gate} :

$$\overline{u_n^2} = 4kTR_{gate},$$

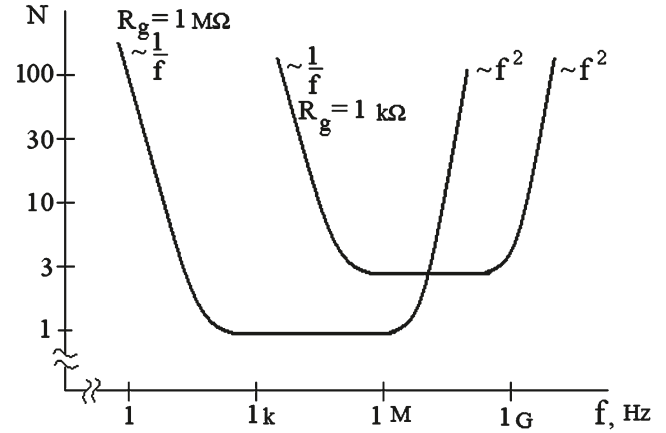
where k is the Boltzmann constant, T is the absolute temperature, and R_{gate} is the gate body resistance; in practice, when analyzing the dynamic properties of a MOSFET in the small signal mode (Fig 5.13), we can neglect the contribution of the thermal noise generated by the volumetric resistances of the source R_s , drain R_d , and substrate R_b . The volumetric resistance of the gate R_g must be taken into account, since it, together with the gate-source capacitance C_{gs} , forms the frequency properties of the MOSFET in the high-frequency region..

- Thermal and flicker noises of the channel are defined as:

$$\overline{i_n^2} = \frac{8}{3}kTS + \frac{k_{fl}I_{d0}^{Y_{fl}}}{f} = \frac{8}{3}kTS \left(1 + \frac{f_{cut.fl}}{f} \right).$$

The rms value of the noise current amplitude $\overline{i_n^2}$ is less than the noise produced by the resistor $R = 1/S$ at $\overline{i_n^2} = 4kTR = 4kT / S$, which is explained by the channel irregularity in the saturation region (S is the slope in the operating point). On low frequencies,

Fig. 5.14 The resulting noise factor of MOS FET



the component $1/f$ predominates in the spectral density of the drain current, while on middle and high frequencies, the thermal component predominates. This permits to determine empirically the values of k_{β} and Y_{β} and, according to them, to calculate the value of cutoff frequency of FET on the flicker noise $f_{cut,\beta}$. For MOS FET, $Y_{\beta} \approx 1-2$, and $k_{\beta} \sim I/L^2$, which allows (on the base of condition of equality of rms currents of thermal and flicker noises produced in the channel) the calculation of:

$$f_{cut,\beta} = \frac{3}{8} \frac{k_{\beta}}{kT} \sqrt{\frac{I_{d0}}{K}},$$

when $K = 2I_{d0} / U_{thr}^2 = S^2 / 2I_{d0}$ is the specific slope, I_{d0} is the DC component of the drain current in the operating point, and U_{thr} is the threshold voltage.

Since $k_{\beta} \sim I/L^2$, where L is the channel length, the ratio is correct for MOSFET, then flicker noise decreases faster than the channel length increases. where L is the channel length, the relation is true, then the flicker noise decreases faster than the channel length increase occurs. It follows from this that with an increase in the scattering power at the drain of the transistor, the lower boundary frequency $f_{cut,\beta}$ will decrease.

At comparison of MOS FET and the FET with the $p-n$ junction, the latter happens to be essentially better since it has significantly lesser spectral power of flicker noises at the same offsets with regard to the operating frequency.

- The induced gate noise is:

$$\overline{i_{n,gate}^2}(f) = \frac{4}{3} kTS \left(\frac{f}{f_T} \right)^2. \quad (5.45)$$

The thermal channel noises falling on the gate due to capacitance coupling between the gate and the channel are the reason of the induced noise current. This causes the cross-correlation of sources of the noise gate current $i_{n,gate}$ and the drain current $i_{n,d}$ with the spectral density S_3 and S_4 , relatively (Fig. 5.13).

The density of noise voltage produced by the gate resistance R_{gate} is inversely proportional to the transistor slope and, relatively, decreases with the drain current growth in the transistor operating point.

If FET is controlled from the signal source with the finite value of the internal resistance and the voltage density of thermal currents $\overline{u_{n,g}^2}(f) = 4kTR_g$, then its value R_g has the essential effect on the resulting noise factor of MOS FET (Fig. 5.14).

Figure 5.14 shows the dependencies of the noise coefficient of the MOSFET, which has technical indicators: the gate resistance $R_{gate} = 100 \Omega$ and the current limit frequency $f_T = 100 \text{ MHz}$, $f_{\beta} = 1 \text{ MHz}$ for an active mode with a steepness at operating point $S = 1 \text{ mA/V}$, the boundary frequency on the flicker-noise $f_{\beta} = 1 \text{ MHz}$ for two the resistances of the signal source $R_g = 1 \text{ k}\Omega$ and $R_g = 1 \text{ M}\Omega$ [35].

In plots, we can distinguish three regions:

- The region of approximate constancy of the noise-factor (in middle frequency region); its minimal value is $N \approx 1$ in condition, when the gate resistance $R_{gate} \ll 1/S$, and the signal source resistance exceeds the value inverse to the slope: $R_g \gg 1/S$.

- The region of noise factor reduction (the region of low frequency, where the flicker noise predominates) inversely proportional to frequency:

$$N \propto \frac{2}{3} \frac{1}{SR_g} \frac{f_{fl}}{f};$$

- The region of noise factor growth (proportional to a square of frequency):

$$N \approx \frac{2}{3} SR_g \left(\frac{f}{f_T} \right)^2.$$

FETs with the controllable p - n junction have the cutoff frequency of the flicker noise f_{fl} (and, relatively, the noise factor in the low-frequency region) by three to four orders lesser than for MOS FETs.

The analysis shows that at $\partial N / \partial S = 0$, the noise factor of MOS FET decreases with the growth of the slope in operating point at constant value of the signal source resistance and will be lesser for the larger drain current in the operating point. The optimal value of the signal source resistance ($R_g \gg R_{gate}$) in the frequency band $f_{fl} < f < f_s$ allows the achievement of the minimal noise factor:

$$N(f) \approx 1 + \frac{4}{3} \frac{f}{f_T}.$$

Achievement of wideband matching in the real devices is impossible due to the fact that the input resistance of the transistor becomes complex $Z_{gate}(f)$.

5.3 Nonlinear Properties of the RF Front-End

The basic technical indices discussed in Section 5.1 assess the conditions for receiving a useful signal occupying a certain frequency band, which is under the influence of various types of interference, among which the most dangerous is narrow-band interference at a known central frequency. The use of digital modulation and multi-station access methods, the development of new frequency ranges, and the structure of the data transmission system led to the need to introduce new methods for assessing the quality of reception and its optimization. In real conditions, the reception occurs on the background of single or several out-of-band interferences with the significant level. In these conditions, it is insufficient to have the above-mentioned technical indices of radio receivers, and we must use the integral characteristics, which connect its sensitivity and selectivity with nonlinear properties of the radio front end, and this integral characteristic is the *dynamic range*. This parameter defines such properties of the radio communication systems as the quality index, capability, EMS, etc.

In the simplest case, at the absence of out-of-band interferences, nonlinear distortions in the main channel are defined by the input signal level: the upper value by nonlinear effects caused by the volt-ampere characteristic nonlinearity of active components of RFA, FC, and IFA (e.g., U_{in1max} ; Fig. 5.15) and the lower value by the proper noises (U_{in1min}). The region between these levels defines the dynamic range of the receiver and corresponds to the linear part of the amplitude characteristic.

The deviation of the amplitude characteristic from the ideal one indicates the occurrence of nonlinear distortions in the amplifier. The effect of such deviations is stronger demonstrated in systems, which applies the amplitude modulation and noticeably weaker at application of angular modulation (FM, PM, FSK, PSK and their variations).

Simultaneous action of a useful signal at the receiver input together with intense in-and out-of-band concentrated interference, even when quasi-consistent filters are used in structure of the RF front-end, leads to a decrease in SNR at its output compared to its input value. The reason for this is the appearance in the RF front-end of combinational components created by harmonics of interference and interference together with a useful signal in the AE of a low-noise amplifier (RFA), frequency converter (Mix 1, Fig. 3.11), and other cascades. The result of such interaction estimated the change in the excess coefficient ζ showing how much the excess coefficient at the output of “ideal” RF front-end h is greater than the output of the “real” h one. The excess coefficient (1.1) is actually SNRs for the ideal and real structure of the RF front-end, and decrease of real SNR (growth ζ) indicates the existence of nonlinear distortions in the RF front-end and, accordingly, a decrease in the sensitivity of the receiver.

For AE protection against overloading in the radio front end, we can use signal filtering by linear circuits keeping their characteristics at different impact levels. At reception of signals with complex shape (with digital modulation), we can apply

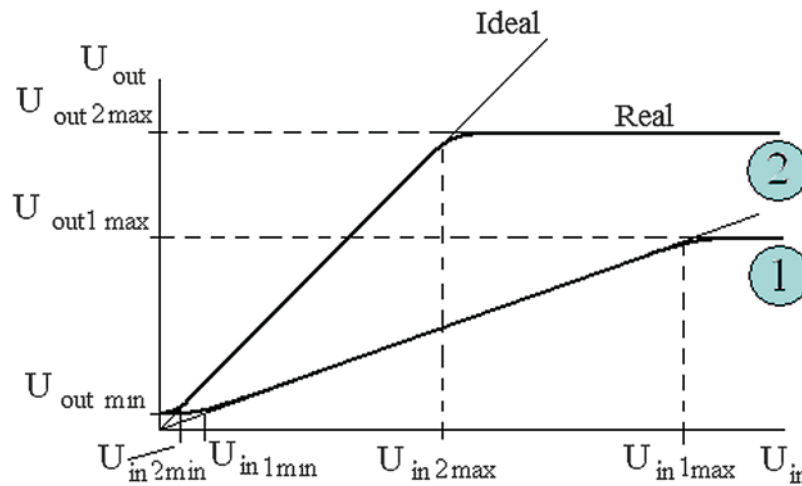


Fig. 5.15 Amplitude characteristic of the amplifier stage

quasi-matched filtering with the help of pass-band filters. From two types of filters applied in the radio front end (with maximally flat AFC and the Chebyshev filters), utilization of filters with maximally flat and closed to rectangular AFC with smooth AFC drop in out-of-band region, having the lesser number of links.

In the block diagram of receivers of modern SHF data transmission systems, the preselector structure usually contains a low-noise amplifier regardless of the construction method (direct conversion, superheterodyne type, with digital intermediate frequency). This ensures a reduction noise factor of the RF front-end (increases sensitivity of the receiver) and reduces the requirements to BPFs used for the fifth collateral channels in superheterodyne receivers. However, along with improving the technical parameters of receiver, the introduction of a low-noise amplifier designed to amplify a weak useful signal (at the sensitivity level) may be accompanied by the appearance of nonlinear distortions that occur when exposed, for example, to concentrated interference from neighboring base stations. MS preselector filters have a bandwidth allocated to the system in which base stations of other operators can operate. At the same time, their power is controlled by the ground radio access network of another operator and can significantly exceed the maximum allowable input level, which puts the AE into the restriction mode. Interaction of the interference (usually lumped) of the high level with operation signals, which is caused by AE nonlinearity, leads to the appearance in the radio front-end pass-band of the large number of conversion products (increasing the noise level) and the specific effects in the form of signal blocking, cross modulation, and intermodulation.

The transition of the active element of LNA to the limiting mode under the influence of strong interference and a weak signal leads to the appearance of a “mirror signal” (with respect to the frequency of strong interference), which has amplitude modulation of useful signal and inverse phase modulation of the useful signal. Such distortion is especially dangerous in communication system with noise-like signals, when areas with a significant excess of the spectral power density are formed in separate sections of uniform spectrum of the radio signal, violating the mutual orthogonality of the received sequences. The appearance of such a “signal” in the LNA at the input of an board repeater, for example, a satellite communication system with CDMA, can lead the end stage of the transmitter to the key mode and failure of the communication system.

The transition, under the influence of powerful interference, of AE of the RF front-end to the key mode leads to an increase in noise component of the output current. The reason for this is an increase in the average value of the constant current component and the appearance, in addition to the first, also of other harmonics, multiples of the interference frequency in the open state interval of active element. The noise factor growth of the stage is also defined by parameters of filtering circuits at its input.

Application of filtering circuits with electronic retuning in the preselector can also lead to reduction of the LNA dynamic range. Exposure to strong interference can cause the varicap, made in form of a reverse-connected diode, to go open. The appearance of a current through an open diode form in its spectrum a large number of components, multiples of the frequency of the interference and combination components. To reduce the risk of a varicap transition to the open state, it is necessary to increase its dynamic range, that is, a possible change in the control voltage on varicap without its transition to the open state. It should also be born in mind that the use of varicaps leads to a significantly smaller tuning range of the preselector counters over the frequency range $k_r = f_{\max}/f_{\min}$, where $f_{\max}$ and $f_{\min}$ are the upper and lower frequencies of the operating range

in comparison with mechanically tunable capacitors. In addition, the capacitance of the varicaps is nonlinearly dependent on the control voltage. To eliminate these problems, varicap matrices are used, consisting of 3-4 pairs of oppositely connected varicaps. This makes it possible to obtain their capacitance-voltage characteristic of the matrix is close to linear and to increase the coefficient of tuning the contours over the frequency range. These parameters depend on the constant charge of varicaps, which can be increased by the growth of the constant control voltage of varicaps and attraction by the larger number of varicaps in the single network. Another reason of dynamic range reduction of electronically retuning networks (amplifying stages) is noises and ripples in the varicap control circuit. Depletion of noises and ripples can be achieved by the growth of the loaded Q-factor of retuned circuits in the preselector.

Similar phenomena are also present in the intermediate frequency amplifiers of superheterodyne receivers, with the exception of issues related to varicap restructuring of the oscillating circuits. The use of a non-tunable preselector and filters in the intermediate frequency path eliminates the causes of nonlinear distortions in the FRF front-end created by varicaps but does not increase dynamic range of the active components used. The resulting nonlinear distortion in the preselector is often mitigated by lowering the LNA gain or using an AE with higher output power. In the intermediate frequency path of superheterodyne receivers, the problem of overloading the final stage is solved using an automatic gain control system, which limits the signal level at the final stage input.

5.3.1 Intermodulation Distortions

Small-signal AE models characterizing the linear mode are true for the larger part of the radio front-end units of receivers of different destinations and the frequency range. Most accurately they describe the properties of the AE in mode A, which is used in the RF front-end cascades of the receiver. At small input impact, parameters of element of physical equivalent circuits do not depend on these signal levels, and output signals are linear functions of the input signals. As we know, the preselector units should have small self-noises, and the input signal level (especially for mobile communication systems) can be close to the noise magnitude created in an antenna. However, at the restricted linear part of the AE amplitude characteristic, nonlinear distortions may occur caused by the wrong mode choice of active elements, which leads to AE transfer into the cutoff mode or in the saturation mode. This results in the appearance of the input signal harmonics and intermodulation distortions (IMD) in the AE current spectrum. In contrast to harmonic distortion arising from overloading AE of the amplifiers, frequency converters with an input signal, intermodulation distortion is the result of impact of narrow-band or in-band interference and their harmonics. Their combinational frequencies do not coincide with input impact harmonics, but results of their interconnection contain combinational components falling in the pass-band of the preselector or the intermediate frequency section.

The cause of IMD is the simultaneous effect of two interference (and their harmonics) at input of receiver with the carrier frequencies created by the BS of the operators, which form spectral components in the AE that coincide with the collateral receive channels. This is a serious problem for receivers AT of mobile communication systems in conditions of dense BS location, for example, near large shopping, sporting and entertainment centers. The effect of the adjacent transmitter impact can lead to the appearance of combinational components in the receiver preselector, which results in the suppression of the useful signal. The influence of lumped interference can be significantly weakened by the inclusion in the preselector of a notch filter tuned to the frequency of the parasitic source, as well as filters of input circuit with the lowest possible bandwidth and the use of amplifiers with a large dynamic range. The decrease of transmitter impact on the IMD magnitude in the receiver can be achieved using the frequency suppressor tuned to frequencies of interacting transmitters and/or their shielding, as well as correcting the frequency characteristics of preliminary amplification stages in transmitters.

For reception conditions, when the input power levels are close to values of limited sensitivity for specified SNR, which is more than 10 dB, intermodulation distortions take on special significance. Intermodulation distortions in the radio receiver (at output of the front-end) are the result of input action of two or more interferences, and the frequencies of which do not coincide with the frequency of main and collateral channels, but form in the output current of the AE as result of their multiplication, the combination components with the frequency: $f_{comb} = p_1 f_{n1} \pm \dots \pm p_m f_{nm}$, where $p_m = \pm 1, \pm 2, \dots$ is a coefficient that determines the order of multiplication of interference frequency f_{nm} is the frequency of the external interference of the 1, ..., m th sources.

Nonlinear properties of AE, which is the main amplifier component, can be estimated from the consideration of the transfer characteristic of the transistor. Let the impact in the transistor input be:

$$x = x_0 + U_1 \cos \omega_1 t + U_2 \cos \omega_2 t, \quad (5.46)$$

where U_1 and U_2 are the amplitudes of impact signals.

The transfer characteristic of AE $y = f(x)$ in the operating point x_0 can be approximated by the Taylor power series; then the output signal y will be:

$$y = f(x) = f(x_0) + \sum_{n=1}^{\infty} \frac{1}{n!} f^{(n)}(x_0) (x - x_0)^n, \quad (5.47)$$

where $f^{(n)}(x_0)$ is the derivative of the transfer function of the n -order in the operating point x_0 .

Substituting the input impact (5.45) into the expression (5.46) for the transfer function and being limited by the first three derivatives of expansion, the output signal can be presented as:

$$\begin{aligned} y = f(x_0) &+ \frac{1}{4} f''(U_1^2 + U_2^2) + \left[f'U_1 + \frac{1}{4} f''U_1 \left(\frac{1}{2} U_1^2 + U_2^2 \right) \right] \cos \omega_1 t + \\ &+ \left[f'U_2 + \frac{1}{4} f''U_2 \left(\frac{1}{2} U_2^2 + U_1^2 \right) \right] \cos \omega_2 t + \frac{1}{4} f''(U_1^2 \cos 2\omega_1 t + U_2^2 \cos 2\omega_2 t) + \\ &+ \frac{1}{24} f'''(U_1^3 \cos 3\omega_1 t + U_2^3 \cos 3\omega_2 t) + \frac{1}{2} f''[U_1 U_2 \cos(\omega_1 \pm \omega_2)t] + \\ &+ \frac{1}{8} f''[U_1^2 U_2 \cos(2\omega_1 \pm \omega_2)t + U_1 U_2^2 \cos(2\omega_2 \pm \omega_1)t] \dots \end{aligned} \quad (5.48)$$

From the above expression, we see that:

- (1) The variation of the operating point position is proportional to the second (in the general case, even) derivative of the transfer characteristic.
- (2) The amplifier linearity (fulfillment of condition $y = kx$) is saved in the case of the equality to zero of the third (in the general case, odd) derivate.
- (3) Even harmonics are caused by even derivatives, while the odd harmonics by odd derivatives.
- (4) Summary and differential components are caused by the even derivatives of the expansion.
- (5) Combinational (intermodulation) components of third and higher orders are caused by the odd derivatives (in the general case, at signal interaction in the load); components caused by even terms of expansion can be added in the resulting IMD value.

As we see from (5.48), the main reason of the nonlinear distortion rise is the existence of the third order nonlinearity (the linearity of the transfer characteristic is disturbed), which can also indirectly cause distortions of the second order. In the general case, all odd expansion components (5.47) are sources of intermodulation distortions, but the third derivative is the most important ($f'''(x)$) because it introduces the largest contribution in the resulting indices.

If in the amplifier input, the harmonic signal $x = U_1 \cos \omega_1 t$ acts, the output signal will be $y = f'U_1 + \frac{1}{8} f'''U_1^3$, and the gain is $f' + \frac{1}{8} f'''U_1^2$. The gain value normalized with regard to the small-signal f' value is $1 + \frac{1}{8} \frac{f'''U_1^2}{f'}$, which shows to its decrease proportional to amplitude square, because $f' > 0, f''' < 0$. Thus, with an increase in the input signal level, starting from some values of its amplitude, the linearity of the amplitude characteristic is violated due to the growth of the influence of the second term, which is proportional to the square amplitude. This leads to the power reduction of the main (first) harmonic and to the appearance of higher harmonics in the output spectrum.

Formation in the stage or the RF front-end output (as a whole) of products of mutual modulation is defined by the coefficient order (numbers of harmonics) of input impacts. Their spurious influence decreases with the growth of the product order number. This is explained by the fact that in devices with the smooth nonlinearity, the intensity of harmonic components decreases according to the $1/n^\alpha$ law, where n is the harmonic number and $\alpha > 2$ is a coefficient. Under the order of the mutual modulation product, we understand the total value of the coefficient $N = |p_1| + |p_2| + |p_3| + \dots$, which defines f_{comb} .

When evaluation intermodulation distortions in the RF front-end of the receiver, we usually restrict ourselves to the third order with biharmonic influence at intermodulation frequencies $u = U_{m1} \cos \omega_1 t + U_{m2} \cos \omega_2 t$, which is due to the highest distortion power of the second and third orders compared to higher-order IMD, that is, the closeness of the pass I-V characteristic of the AE to the cubic polynomial:

$$f_1 \pm f_2 \text{ are intermodulation distortions of second order.} \quad (5.49)$$

$$2f_1 \pm f_2 \text{ are intermodulation distortions of third order.} \quad (5.50)$$

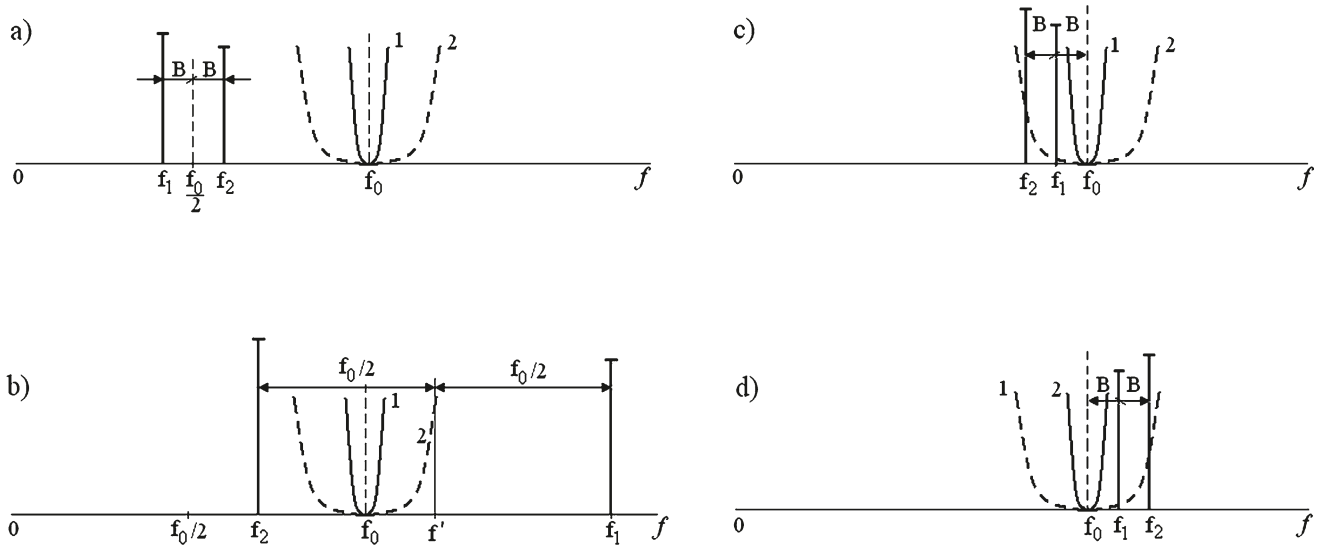


Fig. 5.16 Components of interferences forming intermodulation distortions in the superheterodyne receiver

The formed intermodulation oscillations are made in two types:

- (1) Coinciding with the main channel of amplification or radio reception at the *tuned preselector* or falling into the receiving frequency band of the wideband filter of the non-tuned preselector of the mobile terminal (the base station), that is typical for the mobile radio communication and radio access systems: $p_1 f_1 + p_2 f_2 + \dots = f_0 \pm B/2$, where f_0 is the middle value of the receiving frequency band and B is the pass-band of the intermediate frequency section.
- (2) Coinciding with the collateral receiving channels (on the intermediate frequency and on the frequency of the image channel for receivers of the superheterodyne types): $p_1 f_1 + p_2 f_2 + \dots = f_{IF} \pm B/2$, $p_1 f_1 + p_2 f_2 + \dots = f_{im, ch} \pm B/2$.

As it follows from (5.49) and (5.50), the interferences with the following frequencies have the largest spurious effect on the receiver indices:

$$f_1 \pm f_2 = f_0 \pm B/2, \quad (5.51)$$

$$f_1 \pm f_2 = f_{IF} \pm B/2, \quad (5.52)$$

$$2f_1 \pm f_2 = f_0 \pm B/2, \quad (5.53)$$

$$2f_1 \pm f_2 = f_{IF} \pm B/2. \quad (5.54)$$

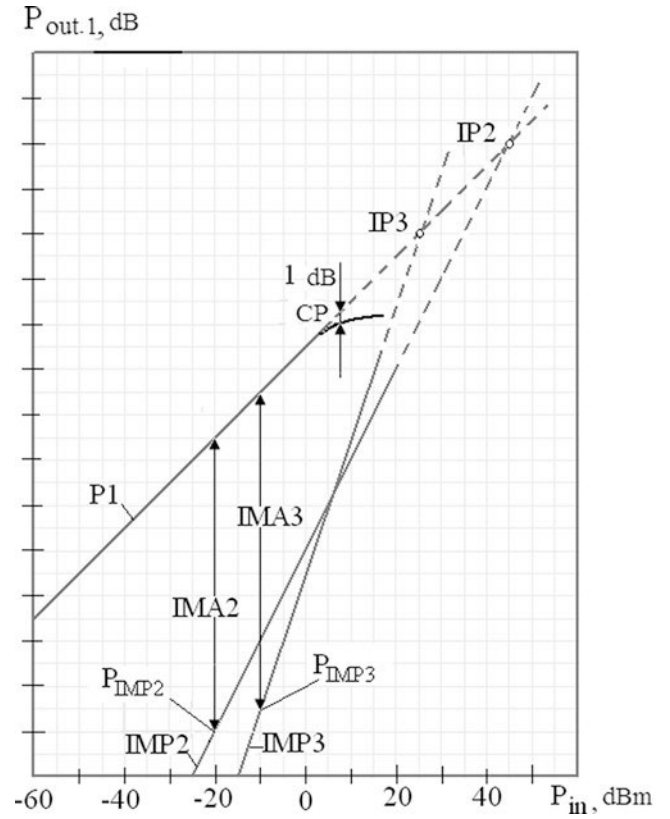
The values of input impact frequencies creating intermodulation distortions of the *second order* based on (5.49) will be $f_1 = (f_0/2) - B$ and $f_2 = (f_0/2) + B$ located symmetrically near the frequency $f_0/2$ (Fig. 5.16a). Its sum creates the interference on the frequency of the receiving signal: $f_1 + f_2 = [(f_0/2) - B] + [(f_0/2) + B] = f_0$. At impact of components with frequencies f_1 and f_2 , which are symmetrically located with regard to some frequency f' (Fig. 5.16b), the difference of these frequencies forms also the component with the frequency f_0 : $f_1 - f_2 = [(f_0/2) + f'] - [(f' - f_0/2)] = f_0$ in the nonlinear active element of the front-end amplifier.

Intermodulation components of the *third order* are formed as the conversion product of components with frequencies f_1 and f_2 separated between themselves and with regard to the useful signal frequency f_0 through the interval B (Fig. 5.16c): $f_1 = f_0 - B$, $f_2 = f_0 - 2B$. Then, from (5.50), it follows $2f_1 - f_2 = 2(f_0 - B) - (f_0 - 2B) = f_0$. At location of spurious impacts, according to Fig. 5.16d, their combination $2f_1 - f_2$ will create the component on the input signal frequency: $f_0 = 2(f_0 + B) - (f_0 + 2B)$.

In all presented figures, numbers 1 and 2 designate the selectivity plots of the wideband part of the radio front end (dotted curve, 2) and the narrowband (section of the intermediate frequency, solid line, 1).

As it follows from Fig. 5.16a and b, an influence of the out-of-band interference components on frequencies f_1 and f_2 is significantly weakened by the preselector, which decreases a danger of the intermodulation distortion appearance of the second order on the frequency of the receiving signal f_0 due to AE nonlinearity in the preselector and the mixer.

Fig. 5.17 Power characteristics for intermodulation distortions



If the third-order intermodulation distortion (Fig. 5.16c and Fig. 5.16d) created by interference at frequencies f_1 and f_2 are not suppressed in the preselector, then they pose greatest danger, forming at its output a component with the signal frequency. At large impact levels, active elements of LNA, RFA become nonlinear and the combination components on the frequency f_0 are formed in output currents.

For the estimation of components of (5.49) and (5.50) types, which arise in the AE output current, we can use the intermodulation coefficient. For estimation of components of (5.49) or (5.50) type, which arise in the AE output current, we can use the intermodulation coefficient. It defines in percent a ratio of oscillations amplitude, which are arisen on the fundamental frequency from the influence of interference at frequencies f_1, f_2 , for example (5.49), second order (U_{m11}), to the normalized signal amplitude (U_{mc}) in the absence of intermodulation for third order distortion (5.55)::

$$K_{11} = (U_{m11} / U_{mc}) 100\%, \quad (5.55)$$

and for the third order (5.49):

$$K_{21} = (U_{m21} / U_{mc}) 100\%. \quad (5.56)$$

In receiving devices with the structure of conversion to the zero frequency, intermodulation distortions of the second order are the most dangerous.

Their contribution to nonlinear distortions increases in the RF front-end of the receiver of microwave networks for various purposes. This is due to the cellular structure of the networks, when signals from the BS of other operators of their own or other mobile networks act at the input of the AT receiver. It is especially important for receivers using non-tunable preselector filters to ensure that each AT operates over the entire frequency range of the system. Taking into consideration that in this range, it is difficult to measure the voltage amplitude in the separate points, we can use the amplitude characteristics in power of separate stages (Figure 5.17).

The output power of the first harmonic ($P_{out,1}$) of the useful signal has the linear dependence $P_{out,1} = f(P_{in})$ to the point P1 of the 1 dB compression (CP, P1), where the difference between prolongation of the amplitude (ideal) characteristic and the real characteristic (solid line) achieves 1 dB.

The interferences' action in the stage input on frequencies f_1 and f_2 (5.49) and (5.50), which provide the total power equal to the input power of the useful signal (P_{in}), permits to obtain the amplitude characteristic for distortion of the second order (P_{IMP2}) or the third order (P_{IMP3}).

The output power of the fundamental harmonic (as well as distortion of the second and third order) increases the input power (P_{in}) of the useful signal, as well as interference at frequencies f_1 and f_2 , respectively (Fig. 5.17). Performing linear interpolation of the output power from the input power (P_{out} vs P_{in}) of the main signal (amplitude characteristic) as well as of the output powers occurring at the fundamental frequency and (P_{IMP2} , P_{IMP3}) caused by intermodulation distortion of the second order *IMP2* and third order (*IMP3*) are defined by the points the intersection of second order *IP2* and third order *IP3* when exposed to two-tone interference.

In the same way, the value the output power of the second-order intermodulation distortion is determined. The compression point (CP) (Fig. 5.17), when amplification decreases by 1 dB, is the absolute boundary of the practically linear dynamic characteristic. The value of **IP3** is some conditional value of the output power of intermodulation of the third order equal to the small-signal output power. Really, this value is never achieved because the stage (an amplifier, FC) before transfers in the saturation mode and admissible distortions will be restricted by the point of 1 dB compression (CP).

The estimation of intermodulation distortion effect of the second order obtains the special significance in connection with the development of wideband access systems with the tuned preselector and especially with active utilization of receivers realized in the direct conversion structure of the input signal and with conversion into zero intermediate frequency. In this case, frequencies $f_1 + f_2$, $2f_1$, $2f_2$, $f_1 - f_2$ are the sources of collateral reception channels, when frequencies f_1 and f_2 are located in the band of received frequencies. These collateral channels appear due to interference phenomenon (hashing) between the interference with the frequency f_2 and the signal on the frequency f_1 or superposition of two interferences with frequencies f_1 and f_2 . Application of wideband filters in the receiver preselector in mobile communication systems does not provide the required interference suppression on those frequencies, which leads (due to their hashing) to the appearance of components falling in the band of the receiver intermediate frequency.

Knowledge of these intercept points allows an indirect estimation of the power value of intermodulation distortions of the third order according to the known impact level equal in power:

$$P_{IMP3} = 3P1 - 2 \cdot IP3 \quad (5.57)$$

where $P1$ is the output level of the main signal and $IP3$ is the position of the intercept point for the intermodulation component of the third order.

The value of the output power of intermodulation distortions of the second order is defined by the relation:

$$P_{IMP2} = 2P1 - IP2. \quad (5.58)$$

The value of the protective gap *IMA2* (Inter Modulation Apart, Fig. 5.17), which separates *IMP2* (Inter Modulation Product 2) for the plot of intermodulation distortions of the second order from the plot $P1$ of the main signal (P_{in}) for some level of the impact P_{IMP2} (e.g., -20 dBm).

$$IMA2 = IP2 - P1. \quad (5.59)$$

Similarly, the *IMA3* value is calculated for the output power characteristic, which is defined by **IMP3** (Inter Modulations-Product 3) of the intermodulation component of the third order:

$$IMA3 = 2 \cdot (IP3 - P1). \quad (5.60)$$

Potentially, intermodulation distortions of the third (odd) order are the most dangerous when they fall in the band of the useful signal and therefore cannot be suppressed by the filtering approach.

The value of intercept point *IP2* for intermodulation distortions of the second order in the output (sometimes, the designation *IP2O* (*IP3O*) is used) can be recalculated to the input *IIP2* (Input Intercept Point 2-order):

$$IIP2_{dB} = IP2_{dB} - K_{p,dB}, \quad (5.61)$$

where K_p is the gain of the unit (radio front end) in power.

Similarly, the *IIP3* value for the third-order intermodulation distortion at the input of the stage is determined for the known value of the output power *IIP3* in the point of interception (Input Intercept Point 3-order):

$$IIP3_{dB} = IP3_{dB} - K_{p,dB}. \quad (5.62)$$

The greater the value of IIPN, the higher the receiver stability to intermodulate distortions of the Nth order.

The resulting magnitude of intermodulation distortions of the cascade connection of linear devices (for instance, receiver units of the front end) is defined as:

$$\frac{1}{IIP3_{RFE}} = \frac{1}{IIP3_1} + \frac{K_{p1}}{IIP3_2} + \frac{K_{p1} \cdot K_{p2}}{IIP3_3} + \dots + \frac{K_{p1} \cdot K_{p2} \cdot \dots \cdot K_{pn}}{IIP3n}, \quad (5.63)$$

where K_{pn} is the gain in power (in times) of the separate stage (unit) and $IIP3n$ is the value of the intercept point of intermodulation distortions of the third order of the n th stage of the radio front-end in the input (in times).

The connection between considered characteristics and the coefficient of nonlinear distortions can be obtained using the polynomial approximation of the transfer characteristic of the RF front-end (or any units).

For define a danger the appearance of intermodulation distortions of the second and third orders in individual stages and the entire RF front-end, the concept of *two-signal* and *three-signal* selectivity is used, which simultaneously is an indirect characteristic of their dynamic range. It is defined as the ratio of two equal interference levels with frequencies f_1 and f_2 (5.49) and (5.50), which form a component at output of the stage at the signal frequency f to the power of the useful signal. The power (or voltage) of the false signal corresponding to the admissible value of $K_{11,adm}$ or $K_{21,adm}$ (5.55), (5.56) should be lower than the power of the useful signal at a given frequency f_0 , usually by 20 dB.

5.3.2 Blocking

Blocking in the radio front end of the receiver is caused by the action in its input of the lumped interference, which does not coincide in frequency with the frequency of the main and collateral reception channels, but its increase leads to signal level and SNR decrease at its output.

If the transfer function (5.47) can be approximated by the Taylor series and the input impact is represented by a combination of the signal and the interference:

$$x = x_0 + U_s \cos \omega_c t + U_{int} \cos \omega_{int} t, \quad (5.64)$$

then, as follows from (5.47), the component with the signal frequency

$$y = \left[f' U_s + \frac{1}{4} f'' U_s \left(\frac{1}{2} U_s^2 + U_{int}^2 \right) \right] \cos \omega_s t \quad (5.65)$$

is defined by the first and third derivatives of the transfer characteristics. When exposed to a strong nearby interference ($U_{int} \gg U_s$), the steepness of the transfer function of the AE will be determined not by the useful signal, but by the level of interference. With an increase in the level of interfering influence, the slope of the transmission characteristic will be determined by influence of the second term in (5.65). When $f'''(x) < 0$, which is typical for the transfer function of real AE, the level of the useful signal at the output of the stage will decrease with an increase of the interference level (the effect of the useful signal can be neglected). Thus, the dynamic range at blocking (DB1) is determined by the compression point on the amplitude characteristic (CP; Fig. 5.17), given by the interference.

The blocking effect of the useful signal is most strongly felt from the influence of interference, which is in the preselector passband: $f_{adj,low} > f_{bl} > f_{adj,up}$, where $f_{adj,low}$ is the frequency value of the adjacent channel located below of the frequency of useful (received) signal and $f_{adj,up}$ is the frequency of the upper adjacent channel. Blocking is estimated by decrease in the power level of the useful signal at the output of the device (usually by 3 dB) when exposed to interference, for a given signal power at input, or example, of low noise amplifier. At that, the carrier frequency of the spurious impact f_{bl} differs from frequencies of collateral conversion channels and lays outside the filter pass-band (B) of the main selectivity. The frequency band in which the blocking effect manifests is called the blocking band. To estimate the blocking effect on receiver characteristics, we can use the blocking coefficient K_{bl} , the dynamic range on blocking DB1, and the coefficient of susceptibility on blocking $\Psi_{sus,bl}$.

The blocking coefficient K_{bl} is defined by parameters of the radio front end and is calculated as:

$$K_{bl} = (U_{s,o} - U_{s,int}) / U_{s,o}, \quad (5.66)$$

where $U_{s,o}$ is the voltage of the signal on the front-end output at absence of interference and $U_{s,int}$ is the signal on the output at interference impact in the input. The relative contribution of the noise factor of the separate stage N_i (IC, RFA, FC, IFA) into the resulting noise factor of the RF front-end N_{RFE} of the superheterodyne receiver (Fig. 3.11) and gains in power of each of them $K_{pi,nom}$ define the resulting blocking coefficient of the RF front-end $K_{bl,RFE}$:

$$K_{bl,RFfe} = 1 / \left(\beta_{n0} + \frac{\beta_{n1}}{K_{bl1}} + \frac{\beta_{n2}}{K_{bl1}K_{bl2}} + \frac{\beta_{n3}}{K_{bl1}K_{bl2}K_{bl3}} \right), \quad (5.67)$$

here

$$\begin{aligned} \beta_{n0} &= \frac{N_{IC}}{N_{RFfe}}; \\ \beta_{n1} &= \frac{(N_{RFA} - 1)}{N_{RFfe} K_{pIC}}; \\ \beta_{n2} &= \frac{(N_{FC} - 1)}{N_{RFfe} K_{pIC} K_{pRFA}}; \\ \beta_{n3} &= \frac{(N_{IFA} - 1)}{N_{RFfe} K_{pIC} K_{pRFA} K_{pFC}}; \end{aligned}$$

are parameters characterizing the contribution of each stage in the total noise factor (5.10). It follows from this that, for instance, at the same value of the β_i parameter produced in each stage, the stage RFA will be determining for the value of the blocking coefficient, since IC is the passive (linear) device. The formal reduction (5.67) of the contribution in $K_{bl,RFfe}$ of the following stages by increase of amplification K_{pRFA} practically is unacceptable due to reduction of the stage dynamic range (compare curves 1 and 2) at gain growth (Fig. 5.15).

Another reason, which limits the gain of radio front-end stages before the input of the last FC, is the pass-band width of each stage. If the signal amplitude in the output of the i th stage on the resonance frequency, at the presence of the harmonic interference in the input, is

$$U_{s,out,i} = K_{0i} U_{s,in,i} (1 + K_{bl,i}), \quad (5.68)$$

where K_{0i} is the stage gain on voltage and $K_{bl,i}$ is the stage blocking coefficient defined by the interference amplitude in its input, then the signal amplitude in the RF front-end output is:

$$U_{s,rfe} = U_{s,in} K_{oIC} \prod_{i=1}^n K_{0,i} (1 + K_{bl,i}). \quad (5.69)$$

The amplitude of the interference detuned with regard to the central frequency of the radio front end in the output of the i th stage is:

$$U_{int,out,i} = U_{int,in} \prod_{j=1}^i K_{0j} \gamma_j(\Delta f), \quad (5.70)$$

where $\gamma_j(\Delta f)$ is the value of normalized gain of the j th stage at the offset $\Delta f = |f_s - f_{int}|$, $K_{0j} \gamma_j(\Delta f) = K_{int,i}$ is the interference gain taking into account the filter selectivity.

Taking into consideration Eqs. (5.67) and (5.69), the blocking coefficient of the radio front is defined by the equation:

$$K_{bl,rfe} = \prod_{i=1}^n \left[1 + \varepsilon_{n,i} U_{int,in}^2 \prod_{j=1}^i K_{0j}^2 \gamma_j^2(\Delta f) \right] - 1. \quad (5.71)$$

At small AE nonlinearity, the blocking coefficient of the radio front end is:

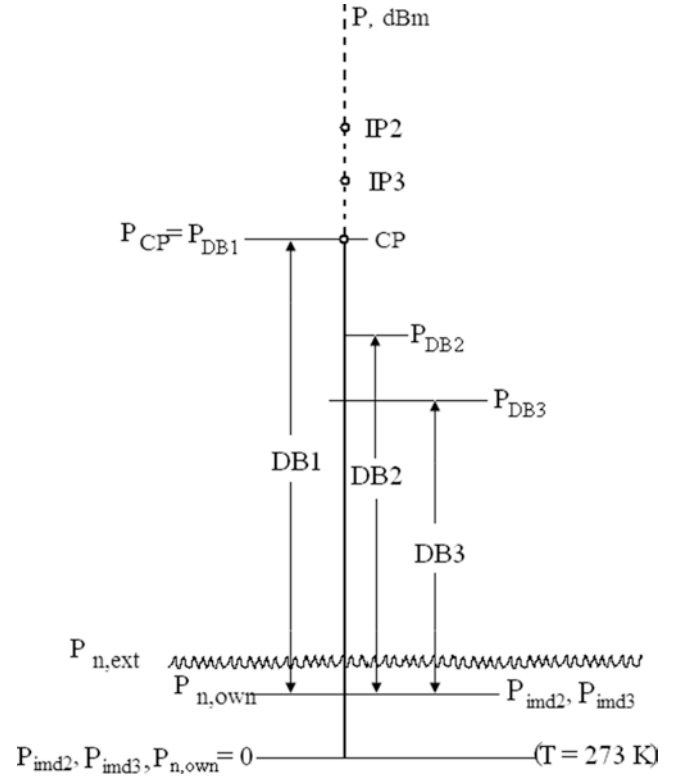
$$K_{bl,rfe} \approx \sum_{i=1}^n K_{bl,i} = U_{int,in}^2 K_{int,IC} (\varepsilon_{n1} + \varepsilon_{n2} K_{int,1}^2 + \varepsilon_{n3} K_{int,2}^2 K_{int,3}^2 + \dots), \quad (5.72)$$

where $\varepsilon_{n,i}$ takes into account AE parameters expressed through the slope values and its derivatives in the equivalent circuit in the Y -parameters system.

Evidently in the case of the poor frequency selectivity of stages, the highest voltage value (and interference) will be in the input of the last front-end stage and, relatively, will create the blocking effect.

At high selectivity of stages $K_n = K_{0n} \ll 1$, this effect will occur in the radio front end as a consequence of the transfer in the saturation mode of the first stage AE (large gain) and the blocking coefficient will be equal to the blocking coefficient of the first RFA stage. It follows from this that since in the real conditions it is impossible to provide such a suppression of out-of-band interference, which permits to neglect by it, then the resonant gain of RFA cannot be very large.

Fig. 5.18 To definition of the dynamic range on intermodulation



The dynamic range on blocking is defined as a ratio of the interference amplitude at specified offset with regard to average frequency of the main channel to the useful signal amplitude corresponding to the receiver sensitivity restricted by noises and is defined as:

$$DB1 \leq \frac{1}{\varepsilon_n K_{int}} \left(1 - \frac{\beta_{n1} K_{bl}}{1 - \beta_{n0} K_{bl}} \right) - \frac{1}{h_s^2}, \quad (5.73)$$

where K_{bl} is the admissible blocking coefficient of RFA at SNR in the receiver input $h_s^2 = U_s^2 / \sigma_n^2$; other designations correspond to (5.67).

As it follows from Fig. 5.18, in practice, we should achieve IP3 values as large as possible compared with the power level of the receiver's own noises $P_{n,own}$, which permits to increase such receiver indices as the dynamic range on blocking DB1 and the dynamic range on intermodulation DB3.

5.3.3 The Cross Modulation

The *cross modulation* occurs at the interaction of the signal with modulated interference, which carrier frequency does not coincide with the frequency of the main or collateral channels of reception and conversion. The reason for the occurrence of cross modulation is the nonlinearity of transmission characteristics of the active elements of radio frequency amplifiers and frequency converters, which cause voltage multiplications and the formation of new spectrum components in the output current. The interference causes the additional modulation of the modulated receiving signal leading to the appearance in its spectrum of components with the interference frequencies. The interference impact leads to the AE slope variation causing the variation of the modulus and the argument of the stage gain, which results in the appearance of spurious amplitude or angular modulation. The influencing interference impact is defined by its intensity compared to the useful signal power as well as by differential AE parameters and its operation modes on DC. To estimate the cross-modulation effect on the quality of the receiving signal, we can use the following indices: the coefficient of the cross modulation $K_{cr, mod}$, the dynamic range $D_{cr, mod}$, and the radio front-end susceptibility $\Psi_{cr, mod}$.

At cross modulation, the components with frequencies Ω_s , Ω_{int} , $\Omega_s \pm \Omega_{int}$, $\Omega_s \pm 2\Omega_{int}$, where Ω_s is the modulation frequency of the useful signal and Ω_{int} and m_{int} are the modulating frequency and the modulation depth of the interference carrying oscillation, are present in the output signal spectrum. If the interference modulation depth $m_{int} \leq 0.3$, then we can neglect the amplitude of the spectrum component with the frequency $2\Omega_{int}$ owing to its smallness compared to the component Ω_{int} .

The dynamic range $D_{cr.mod}$ on the cross modulation is defined by a ratio of the frequency selectivity on cross-modulation distortions at given interference offset with regard to the average frequency of the main channel to the receiver sensitivity. Such a characteristic is obtained by the double-frequency method.

Methods of the fight with the crosstalk are the same as applied for decrease of signal blocking.

5.3.4 Methods for Dynamic Range Widening

The range of possible relative changes in the power level of interference and signals at the receiver input can be from 60...80 dB under normal conditions and reach 140...160 dB when operating in an extreme electromagnetic environment. Without the application of additional approaches of dynamic range broadening, the dynamic range on blocking for the balanced diode circuits of the frequency converters is $D_{bl} \approx 125$ dB μV , and for transistors $D_{bl} \approx 82$ dB μV , which is insufficient for transceivers of the modern systems of the mobile communication, especially for 3G and 4G systems.

All methods used to expand the dynamic range of receivers can be divided into compensation and invariant. In the first case, it is supposed to use special compensating devices: for example, [73], in bridge circuits of frequency converters, parallel connection to diodes of resistors (linear elements) is used; the use of active elements in BT radio frequency amplifiers, which have a high dissipation power on the collector, which increases its dynamic range or, for example, to reduce the amplification factor of the cascade by changing the position of the operating point of the AE.

The dynamic range on intermodulation DB3 is the more important parameters, which defines the upper limit of the admissible level of intermodulation distortions of the third order. This boundary is defined under the condition that the double-tone (measuring) energy-symmetric signal acts in the input. The lower boundary for both dynamic ranges is the level of the own noise power level ($P_{n.own}$) in the receiver input (Figure 5.18).

The power level of own noises $P_{n.own}$ in the receiver is significantly higher than the thermal noise power, which has a value at temperature $T = 290$ K in the matched mode of -174 dBm. For superheterodyne receivers, the $P_{n.own}$ value is mainly determined by P_{IMD3} third-order intermodulation distortion power, and in direct conversion receivers, the P_{IMD2} second-order intermodulation distortion power. The value defining the upper boundary of intermodulation distortion power of the third order P_{DB3} , from which the increasing effect of proper intermodulation interferences from out-of-band signals begins, is calculated from the expression:

$$P_{DB3} = \frac{P_{n.own} + 2 \cdot IP3}{3}. \quad (5.74)$$

The value of the dynamic range by the intermodulation level DB3 is determined from the ratio:

$$DB3 = \frac{2(IP3 - P_{n.own})}{3}. \quad (5.75)$$

The limit value of the output power, which defines the upper boundary of the dynamic range on blocking, coincides the upper boundary in the compression point: $P_{CP} = P_{DB1}$.

The value of the dynamic range on blocking is defined as:

$$DB1 = P_{CP} - P_{n.own}. \quad (5.76)$$

To fulfill the condition $SNR > 10$ dB, it is necessary that on any frequency of the operating range, the input signal level would exceed the value of the average interference power, at least, by 10 dB.

As follows from Fig. 5.15, the intersection point $IP3$ is located on the continuation of straight lines $P1$ and P_{IMP3} and indicates the equality of the power at the output of the stage, created by the useful signal at frequency f_0 and power of the “false” signal at a frequency close to f as combination of frequencies: $2f_1 \pm f_2$ or $f_1 \pm 2f_2$, arising from the multiplication of two frequencies f_1 and f_2 . Its proximity to the point of single-dB compression CP indicates an increased risk of third-order intermodulation distortions $IMP3$ compared to $IMP2$. To reduce the level of intermodulation distortions $IMP3$, it is required that the $IP3$ point is located 10...20 dB above the single-dB compression point (CP). In the ranges of medium (MF) and high (HF) frequencies, reducing risk of intermodulation distortion is achieved by using a bandpass filter with a cutoff frequency ratio of no

more than 2:1 in preselector, which allows for stable reception with successful suppression of peak interference power, which is especially frequent arise in the evening.

When designing radio receivers of modern data transmission systems operating in a tense electromagnetic environment, one should not only take into account the technological features of implementation of the RF front-end but also choose its components and AE operation modes that provide a low level of intermodulation distortion. For the estimation of nonlinear effects, we can use the coefficient of mutual modulation K_{mut} , the dynamic range on mutual modulation D_{mut} , and the susceptibility Ψ_{mut} .

The mutual modulation coefficient K_{mut} is measured at the output of the RF front-end when a signal equal to sensitivity of the receiver is applied at its input and represents the ratio of the level of intermodulation interference to signal level. For the single RFA stage, K_{mut} is defined by the level of interacting interferences, by differential parameters of AE nonlinearity, and by parameters of the resonant load. Obviously, for the reduction of the mutual modulation coefficient, we must apply AEs with the high dynamic range, and increase of the frequency selectivity of the front-end stages with regard to interferences leads to the situation when the resulting value of $K_{mut, rfe}$ is close to the value of $K_{mut, RFA}$.

The dynamic range on intermodulation of second DB2 and third DB3 orders is defined by the ratio of the frequency selectivity on intermodulation at given interference offset with regard to the main channel (at given K_{mut}) to the receiver sensitivity. Factually, DB2 and DB3 are the measure of receiver linearity, which defines the interference level in its input at the required value of $K_{mut, rfe}$.

These indicators describe not only the parameters of individual blocks but also integral characteristics of the receiver, evaluating the device from the standpoint of different requirements: cost and weight and dimensions, dynamic range, intermodulation distortions, etc.

The relative wide dynamic ranges are achieved in the simple superheterodyne receivers (in III, IV groups of complexity), when it is possible to use the single frequency conversion. Each additional conversion leads to the decrease of KP, IP2, and IP3 (Figure 5.18) and, as a consequence, to narrowing of the dynamic range.

Necessity of multiple frequency conversion is caused by some reasons:

- High requirements on the interference suppression on the frequency of the image channel (more than 70 dB) and the channel of direct transfer
- When the optimal value of the intermediate frequency falls into the reception band (other operation ranges)
- When too high value of the intermediate frequency does not provide the required pass-band (on this f_{IF})

Mentioned circumstances restrict a number of conversions in the receiving section at two (at operation frequency below 1 GHz) to save the large dynamic range. At that, if the synchronous detector is used, which is also a converter, it is not taken into account.

The achievement of the wide dynamic range restricts the gain of the wideband receiver section (before the input of the last FC providing the necessary value of the single-signal selectivity). To achieve high sensitivity of the receiver, it is necessary not only to select the matching mode with the antenna to minimize the noise figure, but also to achieve the greatest dynamic range of the input part of receiver in a difficult electromagnetic environment, implementing amplifiers of preselector with a gain of about 3 dB more than the gain of the previous stage [117]. Taking into account the above, when designing the RF front-end, we must use a highly efficient frequency converter (balanced or bridge), made with diodes or MOS transistors to maintain a wide dynamic range. The main amplification of the radio signal is provided in the intermediate frequency path (or in the Base Band), depending on the selected structural scheme of the receiver. The main function of this amplifier is useful signal extraction ($f_{IF1} = f_s - f_{LO}$) as well as suppression of the image channel (on the second intermediate frequency) and the sum component in the FC output: $f_s + f_{LO}$. In the same manner, we can insert the filter of the main selection in the output of the last FC (*at first the preliminary amplifier and then the amplifier with the pass-band filter in the load*).

Existence of the spurious internal feedback in bipolar and field-effect transistors, which can be positive in some region of operating frequencies, exerts a serious influence on the AFC of the previous stage. This phenomenon is especially noticeable in the front-end stages with the resonance load leading to essential AFC irregularity and potentially to the self-excitation danger. To decrease the occurring distortions, it is necessary that the intermodulation distortion level of the third order (the value of IIP3 in the input of the following stage) would be, at least, 3 dB more than IP3O (in the output of the previous stage). With the same aim, we must provide the relation between values of the compression point of the following stage reduced to the input CP_i compared to the value of the output power in the compression point (usually on the level of 1 dB) of the previous stage CP_0 not less than 1 dB. The mentioned relations should be provided by selection of the power level value in the pass-band filter input on the main frequency of +10 dBm (≈ 10 mW) as the counting out point.

In superheterodyne receivers, the preselector and first frequency converter must be highly linear to ensure maximum dynamic range for decrease third order intermodulation distortions (DB3) and hazard blocking (DB1, Fig. 5.18). The gain of

cascades at the output of the last (or only) frequency converter, which forms signal with f_{IF} , are controlled by automatic gain control (AGC) system. The AGC system maintains a constant and optimal signal level at their inputs, which reduces the requirements on the dynamic range of intermediate frequency amplifiers.

Checking Questions and Numerical Examples

1. Give the definition of sensitivity of the radio receiving device. What is the difference between the concepts of “sensitivity” and “real sensitivity”?
2. Denote the sources of internal noises in stages of radio front end of the receiver.
3. Give the definition of the “noise factor” index.
4. Denote the sources of external interferences (distortions, noises) which affect on the receiver operation.
5. How can we calculate the noise factor of the receiver RF front-end (the Miler formula)?
6. Formulate the concept “single-signal selectivity.”
7. Why do we introduce the concepts of “double signal-” and “triple-signal selectivity”? Compare the noise properties of bipolar and field-effect transistors; make a conclusion about possibilities of its decrease.
8. Denote the reasons of distortion appearance in the radio front end of the receiver and methods of its decrease.
9. What is the difference between nonlinear distortions and intermodulation distortions?
10. Which properties are characterized by the “point of 1 dB compression”?
11. What is the reason of the phenomenon called “blocking”?

Denote the methods of intermodulation distortion reduction and dynamic blocking.

Appendix to Chapter 5

Application 5.1 Calculation of admissible and real noise factor.

There is the principal difference between receivers with the operating frequencies below 50 MHz and above 50 MHz.

As we see from [117] and from Fig. 5.19, in the first frequency region, the external and atmospheric interferences having the significant intensity act in the receiver input.

Therefore, in the frequency region below 50 MHz, receivers are under the impact of enough powerful input signals (Fig. 5.20), i.e., they must have the high noise immunity at the relatively low sensitivity.

In *professional receivers* of the range of moderate high frequencies, which have a higher sensitivity, SNR value can be chosen significantly lower than for broadcasting receiver and be::]

$$SNR_{in} = \frac{U_{s.in}}{U_{n.in}} = 3 \div 6 \text{ dB},$$

where $U_{s.in}$, $U_{n.in}$ are effective voltage values, relatively, of the signal and noise in the receiver input, owing to this, the sensitivity can be increased.

Evidently the spectral power of external noises (in relative units) acting in the input restricts the minimal power value of the receiving signal. The minimal value of the noise coefficient should correspond to conditions defined by the function “C,” and for systems of personal radio communication, by the function “G” (Fig. 5.19). For determination of the noise temperature in the operating frequency region, which is not presented in Fig. 5.19, we can apply the linear interpolation of the appropriate dependence.

With increasing frequency, the spectral density of external noise decreases (Fig. 5.19), which will reduce the level of signal arriving at the input of the receiver. However, to provide the required SNR at the input of the receiver, it is necessary to reduce its own noise figure (increase the receiver sensitivity), for example, using the spread spectrum technology of the received signal. For the case when the noise factor defined by external sources (NF_{ext}) is equal to the noise factor (real) recalculated to its input ($N_{real} = N_{RFfe}$), $NF_{ext} = NF_{RFfe}$, the noise level in the receiver output becomes higher by 3 dB than at the absence of external noises. In the high-frequency region (up to 30 MHz), it is unreasonable to fight for the reduction of the own noise of the radio path NF_{RFfe} in order to increase the sensitivity, since the noise generated by external sources and acting at the input of receiver (NF_{ext}) turns out to be significantly higher than the own noise at a given SNR. Reducing the level of

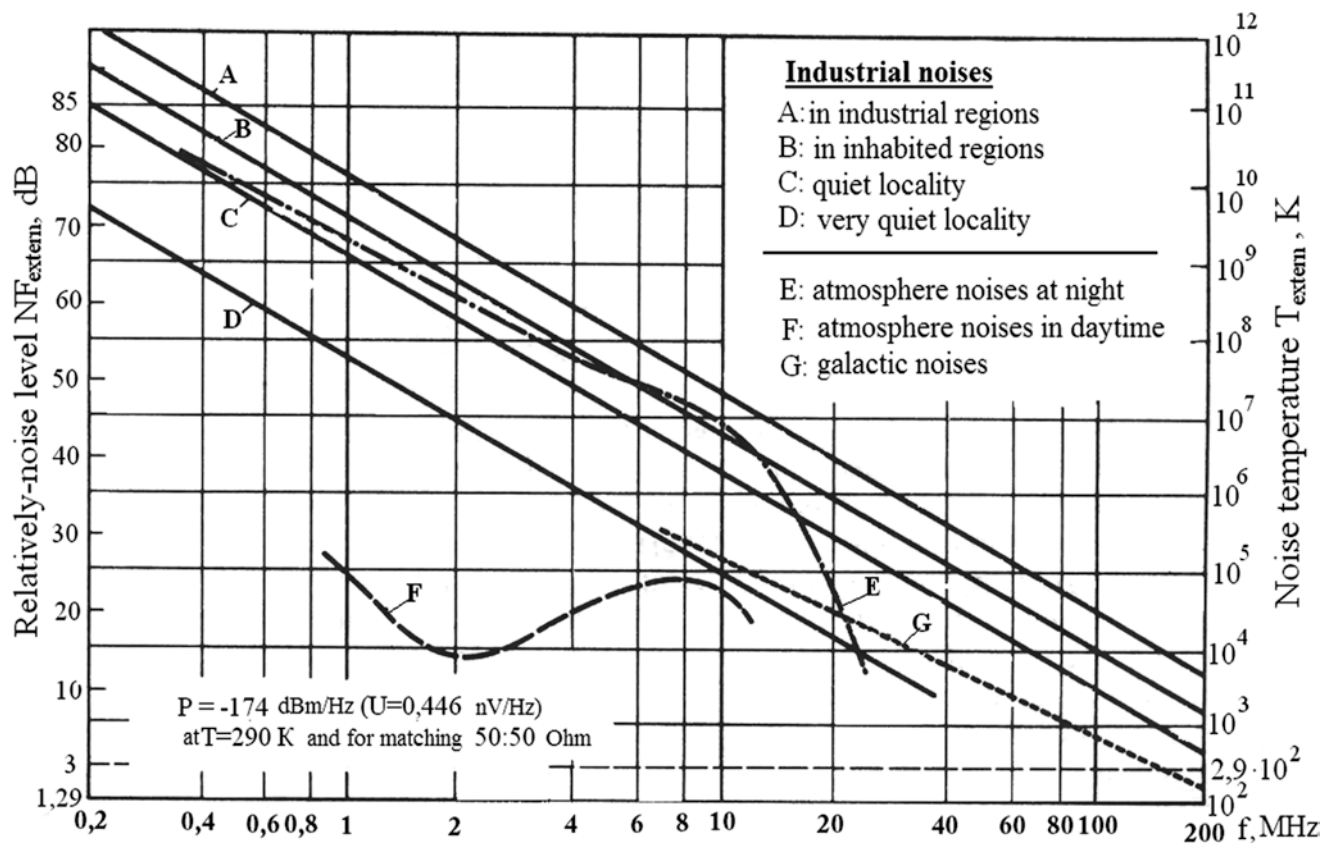


Fig. 5.19 The relative level of interferences at the antenna output from various sources

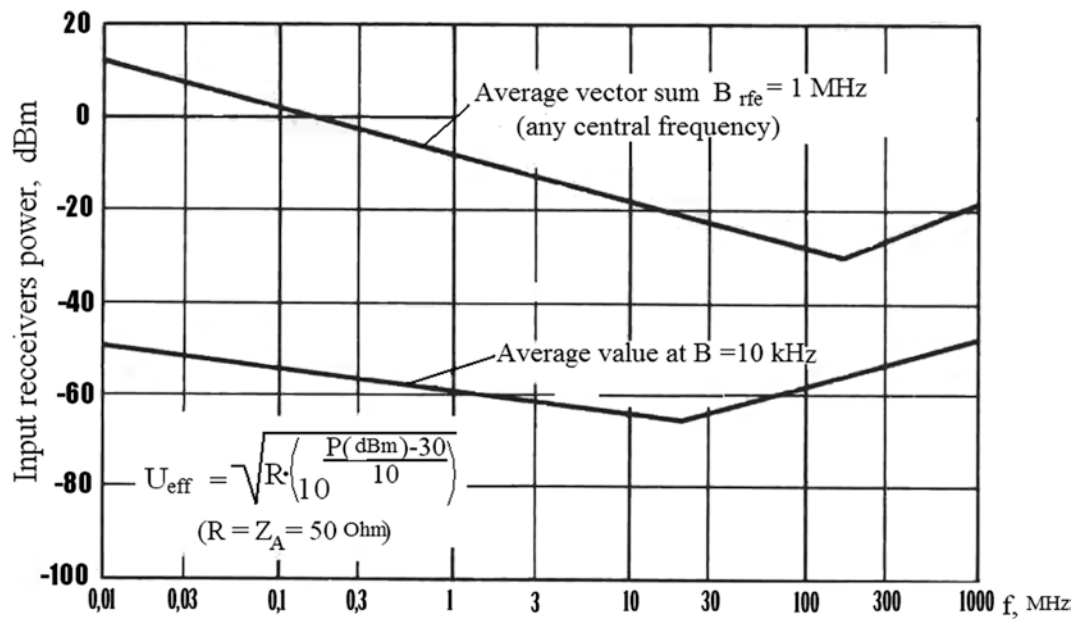


Fig. 5.20 Dependence of useful power at input receiver on the frequency

external noise (and reception of weak signals) can be achieved by using antennas with a narrow radiation pattern and using tunable filters in the preselector, narrowing the received frequency band. In the UHF and microwave region, we must take into account the contribution of Earth's thermal noises corresponding to temperature $T_n \approx 290 \text{ K}^\circ$ ($NF_{ext} \approx 3 \text{ dB}$; Fig. 5.19). For systems of personal communication applying the pencil-beam antennas, the noise factor NF_{RFfe} should correspond to the level of galactic noises (G; Fig. 5.19), and in practice, it is necessary to achieve the possibly lower values. On frequencies above 200 MHz, the noise temperature $T_n \approx 50 \text{ K}^\circ$ ($NF_{ext} \approx 0,7 \text{ dB}$), which allows increasing the receiver sensitivity.

For *radio broadcasting receivers* operating in the region of moderate frequencies, the main requirement is the high fidelity of reproduction of receiving signals. The fulfillment of this requirement can be achieved at the condition that the minimal signal exceeds the interference level acting in the input by 20–30 dB. Practical estimations of atmosphere and industrial interference level show that for receivers of such a type, the sensitivity should be within the limits: $E_{Amin} = 50 \div 200 \mu\text{V}$, where E_A is the signal EMF in the receiving antenna.

Achievement of the required single-signal sensitivity at a given SNR is defined, first of all, by the interference power acting in the receiver input and having the various nature and spectral properties as well as by the RF front-end noise power reduced to the input. At development of the receiver with the operation range below 50 MHz, its sensitivity is mainly defined by external interferences (Fig. 5.19), which decreases requirements to noise indices of the receiver itself. With the frequency growth, the level of industrial and natural interferences decreases, and implementation of receivers with increased sensitivity is mainly defined by noise properties of input stages of the radio front end.

When designing the RF front-end, the receiver noise figure (NF_{RFfe}), recalculated to its input (real noise figure), should be no more then the admissible noise figure generated by external sources, including the thermal noise of antenna NF_{adm} , calculated at its input, usually for conditions corresponding to a quiet area (C, Fig. 5.19). If the condition is met that the permissible noise figure is equal to the real $NF_{adm} = NF_{RFfe}$, the noise level at the output of the RF interface will be twice (3 dB) higher then in the absence of external noise. From this, the noise factor of the front end must be selected as: $N_{RFfe} = N_{adm} - 3$, (dB). If we take into account there are losses in the antenna feeder and thermal interference of the Earth is added to the signal received near the Earth's surface, then at $T = 290 \text{ K}$ the permissible value of the noise factor will be $NF_{adm} \approx 3 \text{ dB}$ to fulfillment the condition.

$$NF_{adm} \geq NF_{RFfe} \quad (\text{II.5.1})$$

it is enough to provide the value $N_{RFfe} \approx 3 \text{ dB}$.

The level of the own noise power (Fig. 5.18) coincides with the power level of intermodulation distortions arising in the front end for useful signal absence in its input due to the impact of out-of-side interferences. The level of the own noise power (Fig. 5.18) coincide with the power level of intermodulation distortions arising in the RF front-end for useful signal absence in its input due to the impact of out-of-side interferences and can be up to $\approx 3 \text{ dB}$. This leads to a decrease in the threshold sensitivity by 3 dB. This factor determines the requirements for the real noise figure of RF front-end NF_{RFfe} , which should be close to the value of admissible noise factor NF_{adm} . Achieving the adequacy of these values ($NF_{adm} \approx NF_{RFfe}$) is ensured by installing an attenuator directly after the antenna connector, which increases the output at point of single-dB compression P1 and the value of the intersection points IP2 and IP3, by reducing the power limit of intermodulation distortion. At the same time, it saves practically unvaried of the dynamic range width on blocking and on intermodulation distortions.

Modern radio communication systems form high requirements to $Se_{im,chk}$, which do not permit to realize them using simple preselector constructions. It is clear that it is technologically very difficult to apply tunable filters in the preselector, since even when the AT operates in the standard of a single system, for example, GSM, it is necessary to provide reception/transmission in the frequency band allocated to the system (about 25 MHz and more) when tuning to subscriber frequency with a relative error of $10 \dots 10$. In addition, the user for work in the VHF, UHF, and SHF band must be provided with the possibility of tuningless operation of the AT receiver, which is achieved by using quartz-stabilized reference oscillators. Tuning of the AT in frequency when operating in one or several systems is ensured by using frequency synthesizers that form a frequency grid. Clarification of AT tuning to the received frequency is achieved using voltage-controlled generators (VCOs). Such a two-stage system for achieving high accuracy (with a relative error of $10 \dots 10$) of AT tuning to the received signal is used in all data transmission systems. In systems of the wideband access, the common frequency band is presented to each user; therefore, at saving of high requirements to $Se_{im,chk}$, we can apply the pass-band filter usually realized on SAW systems and having the squareness coefficient close to 1 ($k_{sq,0,001} = 1.2, \dots, 2$). Increase of the intermediate frequency simplifies the preselector construction; however, without transfer to the lower intermediate frequency using the second frequency conversion, it is rather difficult to satisfy the requirements to $Se_{adj,chk}$ and required sensitivity.

Application 5.2 Calculate the frequency of the intermodulation channel of reception of the base station of GSM standard.

The operating frequency band of the base station of the GSM900 standard in the reception mode is 890–915 MHz (Table 1.2). It is assumed that two interfering influences from MS of other operators are simultaneously received at the input of the BS receiver. In this frequency band, we must receive the signal from any mobile station, which supports the active mode with the base station (Fig. 5.21).

The received MS frequency band in FDD mode is 935–960 MHz. In it, two subscribers are allocated a frequency band of 200 kHz for each connection. In this case, the frequency band of the working channel is provided randomly from the number of free channels.

The impact of a certain pair of frequencies emitted by mobile stations of other operators of the same or other systems can form frequency components on the AE of the preselector and frequency convertor the fall into the operating frequency band of the mobile station. The danger of such distortions, called intermodulation distortions, lies in their suddenness and values of affecting frequencies. The following intermodulation distortions of the third order

$$f_{\text{IMD3}} = 2f_1 \pm f_2, \quad (\text{App. 5.2})$$

$$f_{\text{IMD3}} = 2f_2 \pm f_1 \quad (\text{App. 5.3})$$

can be related to these distortions, where f_1 and f_2 are carrying frequencies in the frequency band of the mobile station transmission ($f_1 < f_2$). A choice of some values of f_1 and f_2 and their second harmonics from the band of transmitted frequencies can lead to the appearance of f_{IMD3} in the band of receiving frequencies of the base station. To reduce the danger of intermodulation distortions' appearance between allocated bands of reception/transmission, the band of the duplex diversity is allocated, which is 20 MHz for GSM900, in which the operating frequency bands of other systems may be located.

Intermodulation distortion of the third order can occur when conditions (App5.2) or (App. 5.3) are fulfilled. Evidently, the sum sign “+” cannot form components IMD3 in the reception band of the mobile station. The upper limit of the difference combinational component IMD3 from (App. 5.2) cannot exceed limit of the band of received frequencies of the MS, which makes it possible to find the value of the lower frequency $f_{2\text{IMD3}}$ at which this occurs:

$$2 \cdot f_1 - f_2 < 960 \text{ MHz; from which we obtain: } f_{2\text{IMD3}} > 820 \text{ MHz.}$$

From the condition that

$$2 \cdot f_1 - f_2 > 935 \text{ MHz, it follows: } f_{2\text{IMD3}} < 845 \text{ MHz, i.e., } 820 < f_{2\text{IMD3}} < 845 \text{ MHz (the region of combinational components). On the frequencies axis, we put the area corresponding to the band of the received of mobile station which is the source of IMD3 (Fig 5.22): (a) frequency band received by MS and formed combination components (from the condition App. 5.2), (b) frequency bands (from the condition App. 5.3)}$$

As follows from Fig. 5.22a, the region of components, which can form IMD3, is situated outside of the receiving frequency band of mobile station, therefore, it is not dangerous.

Using (App. 5.3), we calculate the frequency band in which there is a danger of IMD3 occurrence (Fig. 5.22b):

$$2 \cdot f_2 - f_1 > 935 \text{ MHz, from which } f_{1\text{IMD3}} < 985 \text{ MHz;}$$

$$2 \cdot f_2 - f_1 < 960 \text{ MHz, from which we obtain: } f_1 > 960 \text{ MHz. In this case, the double-side inequality defined the frequency band of combinational components: } 960 < f_1 < 985, \text{ MHz, also do not intercepts with the region of operating frequencies (935 ... 960 MHz) of mobile station receiver. The reason for this phenomenon, as shown above, is not the influence of high-power out-of-band transmitters, but the interaction of the spectral components of the MS of its own network at the input of the BS receiver, which is a feature of the mobile system. } \bullet f - f > 935 \text{ MHz, from which } f < 985 \text{ MHz; } 2 \bullet f - f < 960 \text{ MHz, from which we obtain: } f > 960 \text{ MHz. In this case, the double-side inequality defined the frequency band of combinational components: } 960 < f$$

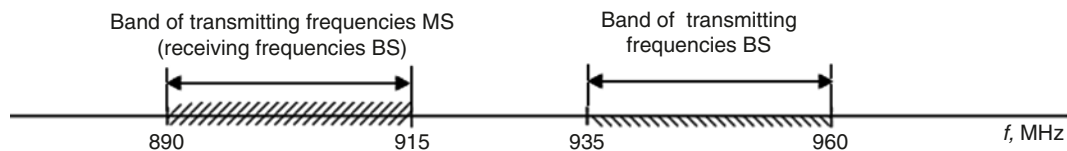


Fig. 5.21 Distribution of frequency bands between MS and BS of the GSM system in frequency duplexing mode

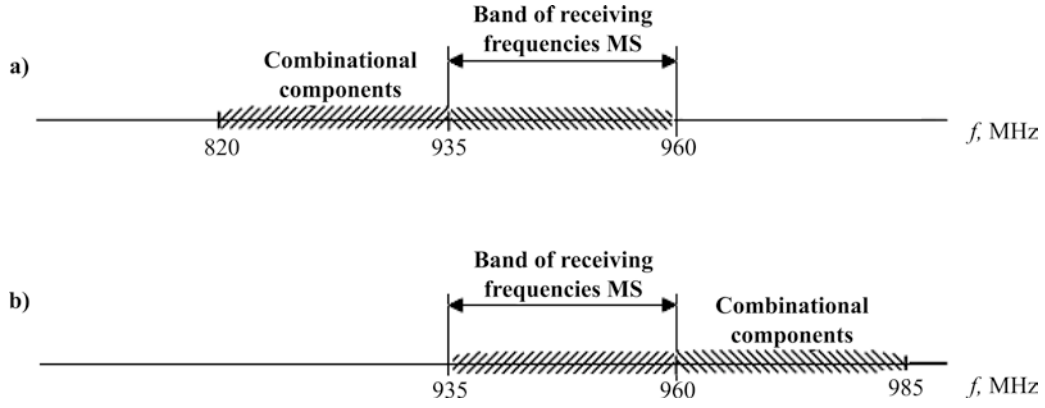


Fig. 5.22 (a) Frequency bands transmitted by MS (received by BS) and formed combination components (from the condition App. 5.2), (b) frequency bands transmitted by MS (received by BS) and formed combination components (from the condition App. 5.3)

Application 5.3 Calculate the value of dynamic range in third harmonic DB3.

Initial data:

- 1) The noise factor of the receiver: $NF = 5$ dB
 - 2) The noise pass-band of the radio front end: $B_n = 5$ MHz
 - 3) The intercept point for intermodulation for the third order on output: $IP3 = 50$ dBm
- Using (5.11) for the noise factor

$$N = \frac{P_{s,in} / P_{n,in}}{P_{s,out} / P_{n,out}},$$

and converting (5.14), we obtain the expression for the noise power produced by an antenna of the mobile station in the receiver input:

$$P_{n,in} = \frac{P_{s,in}}{N(P_{s,out} / P_{n,out})} = \frac{S_n \cdot N \cdot (P_{s,out} / P_{n,out}) \cdot B_n}{R_A (P_{s,out} / P_{n,out})} = \frac{S_n \cdot N \cdot B_n}{R_A},$$

where $S_n = 4kT_0R_{in}$ is the spectral density of the noise average power.

Assuming that an antenna is matched with the receiver input ($R_A = R_{in}$), the thermal noise power is determined as:

$$\frac{S_n}{R_A} = \frac{4kT_0R_{in}}{4R_A} = kT_0.$$

For the room temperature ($T = T_0 = 290$ K), we have:

$$\frac{S_n}{R_A} = 10 \lg \left(\frac{kT_0}{10^{-3}} \right) = -174 \text{ dBm / Hz}.$$

Then, the thermal noise power in the receiver input is:

$$P_{in} = kT_0 \cdot N \cdot B_n, [\text{W}],$$

or in the logarithmic scale:

$$P_{n,in} = kT_0 + NF + 10 \lg B_n = -174 + 5 + 10 \lg (5 \cdot 10^6) = -174 + 5 + 7 + 60 = -102 \text{ dBm}. \rightarrow (63.1 \cdot 10^{-15} [\text{W}]).$$

The equation for the noise factor through (5.11) can be written through the RF front-end gain in power and the average value of the own noise power:

$$N = \frac{P_{s,in} / P_{n,in}}{K_{p,op,p} P_{s,in} / K_{p,op,p} (P_{n,in} + P_{n,own})} = 1 + \frac{P_{n,own}}{P_{n,in}} \quad (\text{App. 5.4})$$

The value of the own noise power is calculated as:

$$P_{n,own} = P_{n,in}(N - 1).$$

For values $NF = 5$ dB ($N = 3.16$), the own noise power is:

$$P_{n,own} = 63.1 \cdot 10^{-15} (3.16 - 1) = 136.3 \cdot 10^{-15} \text{ [W]} \rightarrow 0.136 \text{ [pW]},$$

and the interception point coordinates are $IP3 = 50$ dBm $= 10^2$ [W].

The dynamic range on intermodulation distortions of the third order is calculated on the formula (5.75) correction completed:

$$DB3 = \frac{2(IP3 - P_{n,own})}{3} = \frac{2(10^2 - 136 \cdot 10^{-15})}{3} = 0.67 \cdot 10^2 = 67 \text{ [W]} \rightarrow 48.3 \text{ [dBm]}.$$

The results obtained for the selected conditions show that to achieve the target value $IIP3 = 50$ dBm, it is sufficient that the dynamic range of receiver is 48.7 dBm, which is quite simple to implement.

Chapter 6

The MicroCap12 System of Circuit Modeling

At the end of the first volume of our textbook, i.e., before the detailed investigation of the theory and structure of different stages of modern radio receivers, we briefly describe one of the popular modern software packages, which is extremely convenient for modeling of the most important elements of subsystems in receivers: the MicroCap12 system in this chapter. This software packet will be used in the second volume of our textbook. The main characteristics of this software for circuit modeling are discussed in Sect. 6.1. After a short description, we consider (Sect. 6.2) the structures of main menus and submenus of this packet. Section 6.3 is devoted to the mathematical models of components (passive, active). Special attention is attracted to modern models of bipolar transistors and field-effect transistors with controlling p - n junction and MOS transistors. In this model, noise processes are considered because noise properties are extremely important for all modes of modern radio receivers. Section 6.4 describes the menu for the most essential commands and modes: DC analysis, AC small-signal analysis, steady-state analysis, large-signal analysis, stability analysis, and transient investigations. AC analysis is executed in time and frequency domains.

6.1 Main Characteristics of the MC12 System of Circuit Modeling

The software packet MicroCap12 is the software system for circuit modeling of analog, digital, and combined (analog-digital) devices of the universal destination. A characteristic feature of this software package is the use of friendly graphical interface that allows users to access it for numerical analysis of device that do not have serious computer modeling skills.

The capabilities of this program, especially when using ready-made or developed macromodels, are comparable to professional software packages DESIGNLAB, ORCAD, PCAD2002, and AWR and have an undoubted advantage in the simplicity of describing the schematic diagrams of devices and setting the analysis conditions, starting with direct currents. The upper limit of using MicroCap in frequency is limited only by the correctness of the models used of the active components. The complete MC12 compatibility with Spice models and Spice circuits, and also a possibility of its conversion, permits MicroCap11 user to apply all developments intended for these packets. The site address, which allows the free download of the demo version of the last version of MC12, is <http://www.spectrum-soft.com/demodown.shtml> [28], which *supports the operation of all files which were created in previous versions*. The student version of MicroCap12 Evaluation permits to execute the analysis of electronic circuits including models of analog and digital integrated circuits at the number of nodes not more than 50.

The analysis in MicroCap is performed at the level of schematic or structural diagrams of analog and digital devices and allows you to use the usual conventional graphical designations for the description. Models of active and passive components have the ability to form complex structures, the components of which can change according to a given algorithm, expanding the limits of analysis. The built-in Spice library contains a large number of models of used components [27], including integrated circuits. The program allows you to calculate Spice models based on empirical or calculated data, for example, new types of transistors, to create your own four-terminal blocks that perform specified functions. We would like to note that at present, in connection with active implementation of the multifunctional integrated circuits (IC) including the large number of components of various blocks, companies-manufacturers place in their Internet sites, for instance, <http://www.analog.com>, <http://www.rfmd.com/>, <http://www.maxom-ic.com/>, etc., descriptions of textual PSPICE (Simulation Program with Integrated Circuit Emphasis) models of such devices. The analysis of electronic circuit properties is provided after construction

of electrical or functional diagrams with the help of embedded graphical editor using the library of conditional graphical designations of electronic components.

This software packet does not require large computer resources and enables even in demo version to analyze electronic devices of moderate complexity. In addition, the convenient window interface essentially facilitates a procedure of interaction with the software.

In this chapter, we consider the fundamentals of MC12 software utilization in student variant, which allows analyzing of the RF front-end stages usually after fulfilled manual calculation.. After entering the analyzed electrical circuit into the *Circuit window* and selecting the *Analysis...* command the program always performs a DC circuit analysis (in any of the possible analysis options). The results of the analysis (voltage in the circuit nodes and currents in the branches) can be displayed on the screen in the *Circuit window*, which requires enabling their display in the bottom line of the *Circuit window*. To continue analysis in the frequency or time domain, it is necessary to check their compliance with the selected values (or pre-calculated) obtained during the simulation. After eliminating the differences, varying the values of the components of the scheme, continue its analysis in the selected area. Then we perform an investigation of the corrected (specified) electrical diagram in the time or frequency domain and the optimization of obtained device indices according to chosen criteria. At device modeling containing several stages, this allows the estimation of their mutual influence, the choice coupling network parameters to provide the required integral indices.

Loading of the MicroCap 12 (MC12) software is possible in IBM PC having the Pentium II processor (and more modern) with the operation system Windows 95/98/ME or Windows NT4/2000/XP/7/8 and with the monitor SVGA and higher. The basic methods of working with the MC11 program, described in the book, can be easily repeated by the reader, since its earlier versions are given on the manufacturers' website [28]. Features of MC12 application and some differences in the interface for the students' version MC12/demo will be discussed during description of specific usage.

The MC12 multi-window interface with smooth and expanding menus greatly facilitates interaction with a program that implements simulation modeling of real circuit together with the user. After MC12 loading and entering to the MC12 system of circuit modeling, the window of diagrams (operation window) is appeared in the monitor screen (Fig. 6.1).

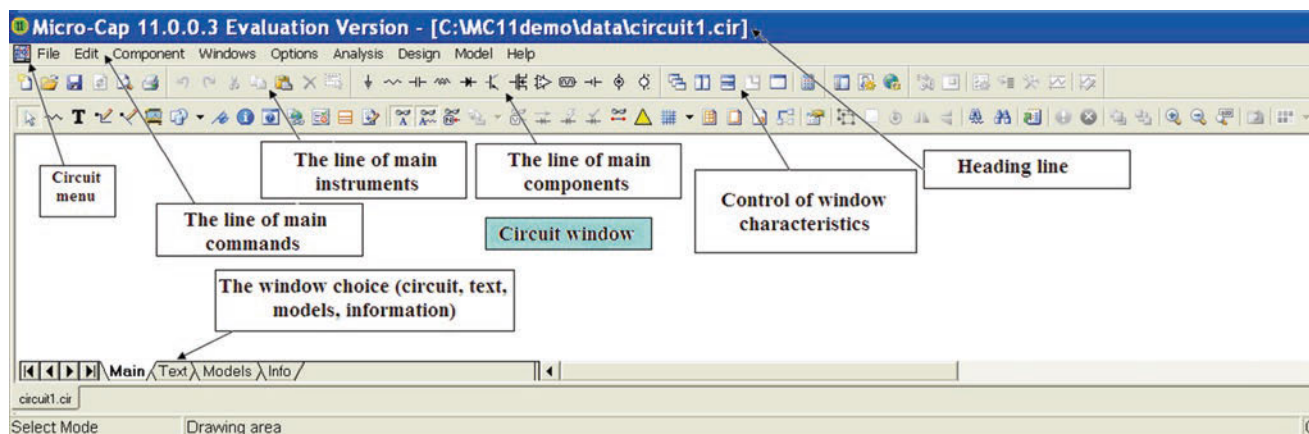


Fig. 6.1 Operation window of the MC11 system of circuit modeling

The main window where the electrical diagram of the block or device is indicated is the circuit window. In the upper line of this window (*the heading line*), the file and directory names are indicated, in which it is situated. At opened analysis window, the type of analysis is indicated.

In the left upper corner of the window, the button of the *system menu* is located, which is the control button for Windows applications. It permits to change, to move, to remove, and to close windows. Below it, the button of the *circuit menu* is located; in functions, it is similar to the button of the system menu, but it acts only in the circuit window.

The *main command line* (*File, Edit, Component*, etc.) that are executed immediately or when one of them is clicked, a submenu is formed that includes a number of options. For example, when the *Component* command is selected, an additional menu appears containing passive and active components, generators, etc., which can be used in the formation of electrical circuit.

The *line of the main instruments* realizes the commands of immediate actions according to chosen pictogram. Pictograms of mode switching provide the chosen mode before fulfillment of the following command.

The *window size control* buttons allow you to expand the *Circuit window* to the full screen, reduce its size, or close the window.

The window of *messages about errors* appears below the results window at any disturbances during the circuit analysis or at incorrectly entered conditions of an analysis, model parameters, etc.

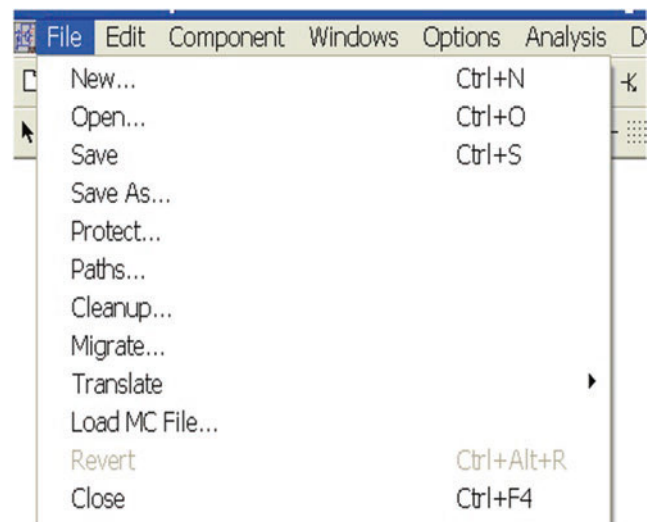
In this chapter, we consider the main approaches for operation with the software in order to enter the electrical diagram of main stages (units) of the radio front end with utilization of mathematical models of component and execution of modeling in frequency and time domain.

6.2 Menu of the Line Commands

File (Fig. 6.2) contains commands for operation with files of diagrams, textual tasks in the Spice format, and files of libraries of mathematical models.

When executing the *File* command, a list of actions describing working with files will appear in drop-down submenu (Fig. 6.2).

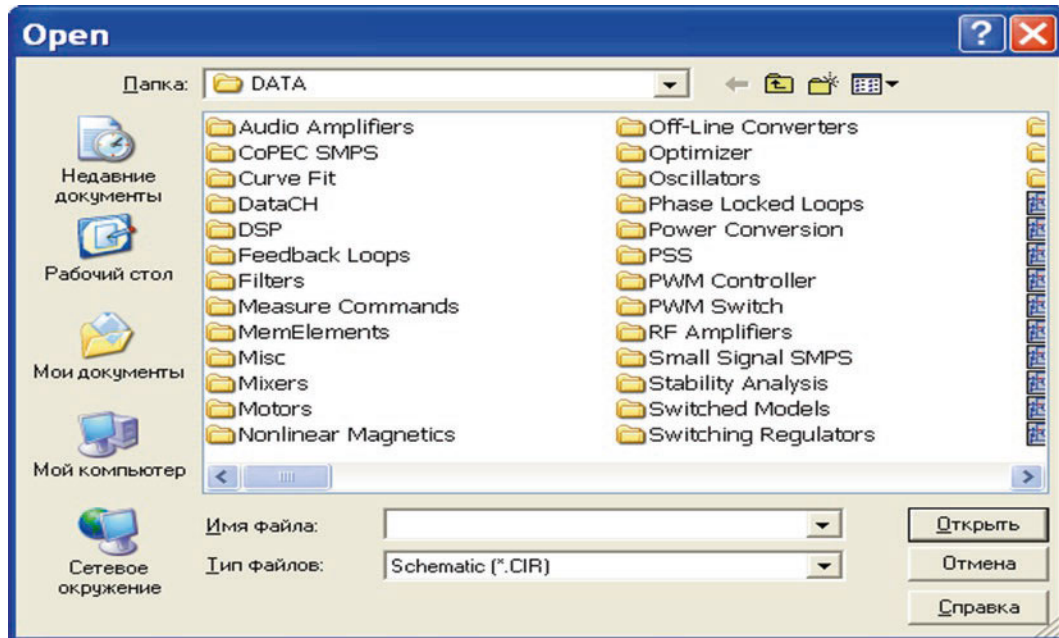
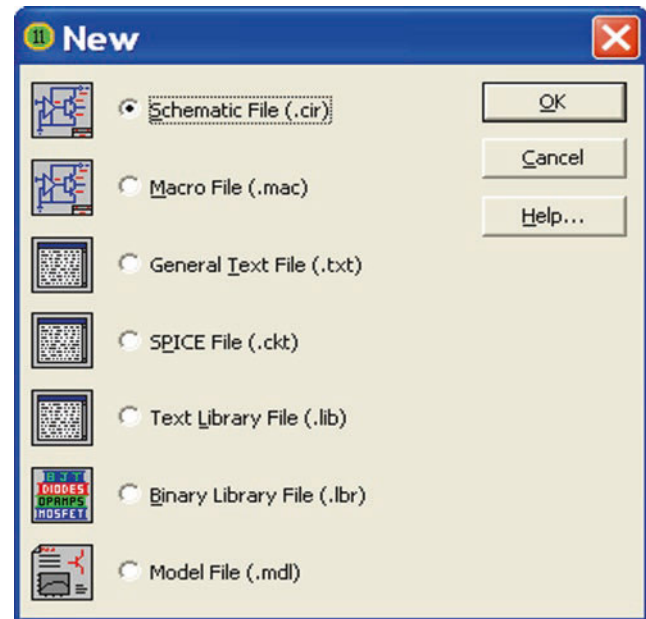
Fig. 6.2 When activating an individual line of the *File* submenu, comments appear on the implementation of specific function



New (Fig. 6.2): creation of the new file (Fig. 6.3): the diagram (Schematic.cir); the file containing macro-models (.mac); the standard textual file (.txt); the textual file Spice/Text in the format Spice (.ckt); the textual library file (.lib) of components (transistors, diodes, etc.); the binary library file (.lbr) of components or models (MDL) of the component (permitted utilization only in professional version of MC11). The type of file to be created is selected by setting a dot before the selected file type and confirming the selection by clicking OK.

Open (Fig. 6.2): opening a previously created file and stored in the *Data* folder of the MC12 program (Fig. 6.4). Other folders are formed by developers and sorted by type of models of stored devices (generators, mixers, amplifiers, etc.) as well as user-created. When using any schematic files, you must sequentially execute the commands: Open → Data → and then select a folder using the scroll bar to find the necessary file with the extension (.cir). The electrical diagram of the model will be saved in the sub-directory DATA (MC11\DATA) as a file (the second line of the circuit window ; Fig. 6.1). Besides the created model, in this folder, examples of construction and investigation of properties of various stages and devices are presented.

For the convenience of accessing the schemes you have developed or frequently used, it is advisable to create your own folder (e.g., *DataCH*) to store them. To access any of them, you need to enter the *Data* submenu and activate your folder by running the *Open* command..

Fig. 6.3 Creation of new file**Fig. 6.4** Sub-directory of schematic files and examples simulation of various devices

Save: saving the circuit from the active window.

Save as...: saving the circuit from the active window in the new file, which name is indicated on the software request and can be changed to another name corresponding to the folder format. At that, it is necessary to note extensions (.cir, .mac, .text, etc.) used at operation in the specific dialog window.

With destination of other commands and their application, we can be acquainted in [28, help file].

Translate: the conversion of formats of circuit files (text format. LIB into binary Spice file of the format. LBR) and vice versa. This is the most popular conversion, since in the catalogs of manufacturers, the parameters of active components are usually presented in form of text files, and a binary file is used in modeling.

Edit: contains commands for editing (Fig. 6.4).

Fig. 6.5 The most popular commands for editing



The most often used commands are presented by pictograms on the line of instruments (Fig. 6.1) and in Table 6.1.

Component: the menu of components contains the catalogue of libraries of analog and digital components. The catalogue has the hierarchical menu containing models of typical components located in text files with extension .LIB.

Let us consider in detail the choice and application of analog component models at analysis of electrical circuits of the radio front end of receiving devices (Fig. 6.5).

The list of main components of analog devices includes the mathematical models of different types of components from the simplest analog (Analog Primitives) and digital primitives to rather complex devices, which are models stored in the MC library or obtained from other sources. Examples of the model choice and application of the most popular components in electrical circuits of different receiver stages are presented in [73], as well as in appendixes to each chapter of this book.

If main commands are enough obvious from designation, their deciphering and methods of editing can be found in [64].

Options: the command menu of the editing mode and the mode of specifying different parameters of the MC11 software (Fig. 6.7).

Windows: contains commands for operation with windows (Fig. 6.6).

Table 6.1 The most often used commands

Pictogram	Command	Comments
	Can't Undo	Cancellation of the last editing command; repeated fulfillment of the command recovers the original variant
	Can't Redo	Cancellation of the last command
	Cut	Deleting the selected object and placing it in clipboard windows
	Paste	Copying of the exchange buffer into the current window
	Clear	Cancellation of chosen object without copying into the exchange buffer
	Select All	Choice of all objects of the current window
	Rubberbanding	«Rubber» connections
	Step Box	Copying of the circuit fragment in mentioned direction of a given number of times
	Mirror Box	Creation of the mirror image of the circuit fragment
	Rotate	Rotation to 90° of the circuit fragment counterclockwise

Fig. 6.6 Commands for operation with windows

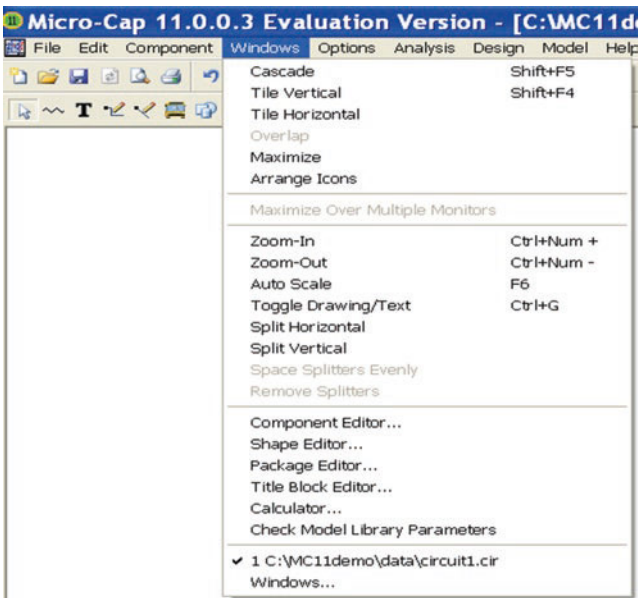
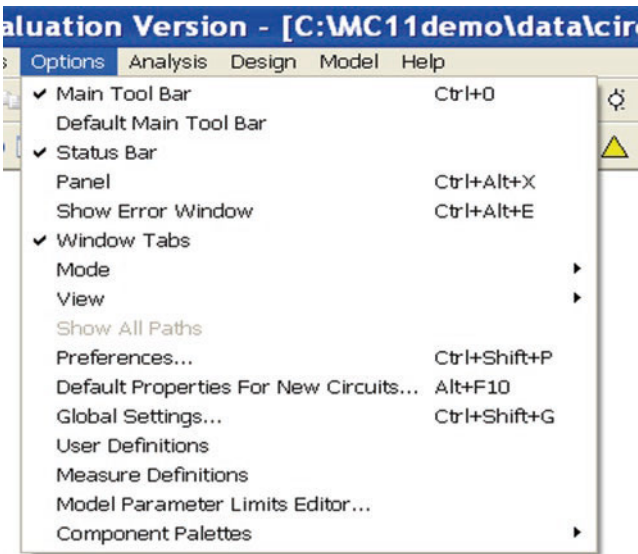


Fig. 6.7 Command menu of the editing mode



The choice of command **Options** in menu of the main window allows activation by the click of the left mouse button of the appropriate line of the bookmark:
Mode: the mode choice for operations fulfillment (Table 6.2). For details, see [28].

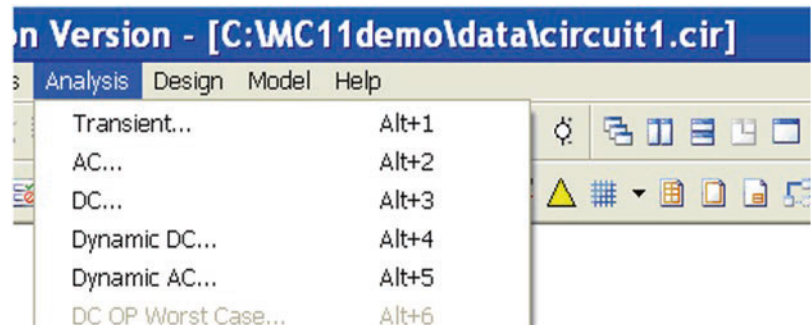
Table 6.2 The mode choice for operations fulfillment

Pictogram	Command	Comments
	Select	Choice of objects for further editing
	Component	Adding of the component into the circuit
	Text	Drawing of textual inscriptions in the diagram
	Wire	Entering of orthogonal circuits

Table 6.3 Indices and explanations, which are often used in electric diagrams

Pictogram	Command	Comments
	Grid text	Text inscriptions
	Attribute text	Designations of component positions
	Node numbers	Numbers of circuit points
	Node voltages/states	Focal points
	Current	Currents in branches

Fig. 6.8 The mode choice for operations fulfillment



View: a choice of explanations presented in the diagram (Table 6.3).

Analysis: the menu of commands providing the modeling mode (Fig. 6.8).

On falling-out sub-menu, we can choose different types of analysis of electronic circuit:

Transient...: calculation of transient.

AC...: “Alternating current,” calculation of frequency characteristics.

DC...: “Direct current,” calculation of transfer functions in DC at parameter variation of the impact sources, temperature, or parameters of component models.

Dynamic DC...: “Dynamic direct current,” calculation of DC mode with presentation in the window of voltage values in nodes, currents in branches, or the power dissipated in components of electrical circuit.

Dynamic AC...: “Dynamic alternating current,” calculation of small-signal parameters for chosen frequencies with presentation of obtained values in the circuit nodes at variation of component values.

Harmonic Distortion...: calculation of nonlinear distortion with utilization of fast Fourier transform.

Intermodulation Distortion...: calculation of intermodulation distortions.

6.3 Mathematical Models of Component in MC12 Medium

6.3.1 Mathematical Models of Components

6.3.1.1 Models of Passive Components

Pictograms of the most often used components of electronic circuits are presented in the line of *main components* (Fig. 6.1). Their designations and the model names are presented in Table 6.4.

Table 6.4 Designations and model names

Pictogram	Name of model type	Component
	G	«ground», common bus; electrical ground
	RES	Resistor (* standard of Russia)
	CAP	Capacitor
	IND	Inductor
	D	Semiconductor diode
	NPN	Bipolar <i>n-p-n</i> transistor
	nMOS	MOS transistor with embedded channel with depletion
	OPA	Operational amplifier with five terminals
	DClock	Source of digital signal
	B	Battery (source of DC voltage)
	C	Current source (the shape of formed signal is given in the sub-menu current source)
	V	EMF source (the shape of formed signal is given in the sub-menu voltage source)

Note: in parenthesis, the designations of electric diagram components are presented corresponding to Russian standards

At pressing on the pictogram of any mentioned designations, the software MC12 replaces the cursor in the operation window onto the image of component and transfers in the mode **Component**. Location of component happens at pressing on the left mouse button. Keeping the left button pressed, by the simultaneous pressing on the right button, we can rotate the component image. Release of the left mouse button fixes the position of the component model in the circuit window, and software transfers to the sub-menu of conferment of the designation, specification of parameters, etc.

6.3.1.2 Models of the Input Impact Source

The Generator of Harmonic Oscillations The generator type can be chosen using the line of main components in the main window and then we can choose from the large number of models of generators with different signal shapes, with modulation or without modulation, etc. We are limited (as an example) by inclusion in the circuit under examination of the model of the harmonic EMF source without any modulation. Choosing the command Component in the menu of the main window, we activate the main signs of the EMF source model in the sequentially falling footnotes: Analog Primitives → Waveform Sources → Voltage Source (Fig. 6.9).

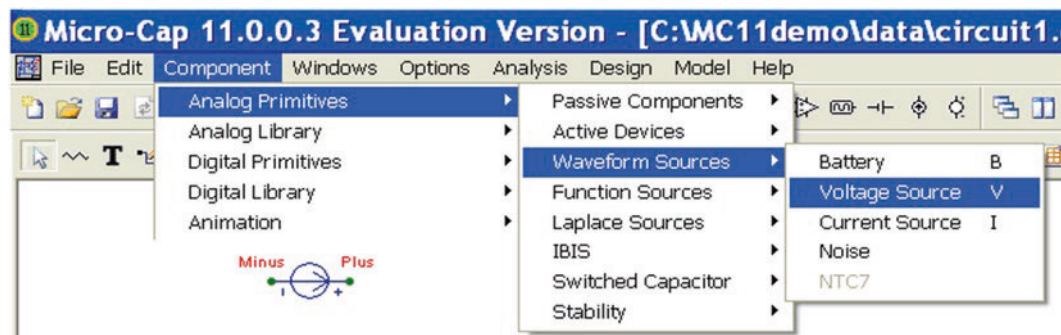


Fig. 6.9 The sequence of model formation of the harmonic EMF source

After performing these actions, the designation of the generator model and submenu of *Sine Source* parameters appear in schematics window. Depending on the task to be solved, you can select a ready-made model offered in the right window of the submenu or create a new one by clicking the *New* button (Fig. 6.10a).

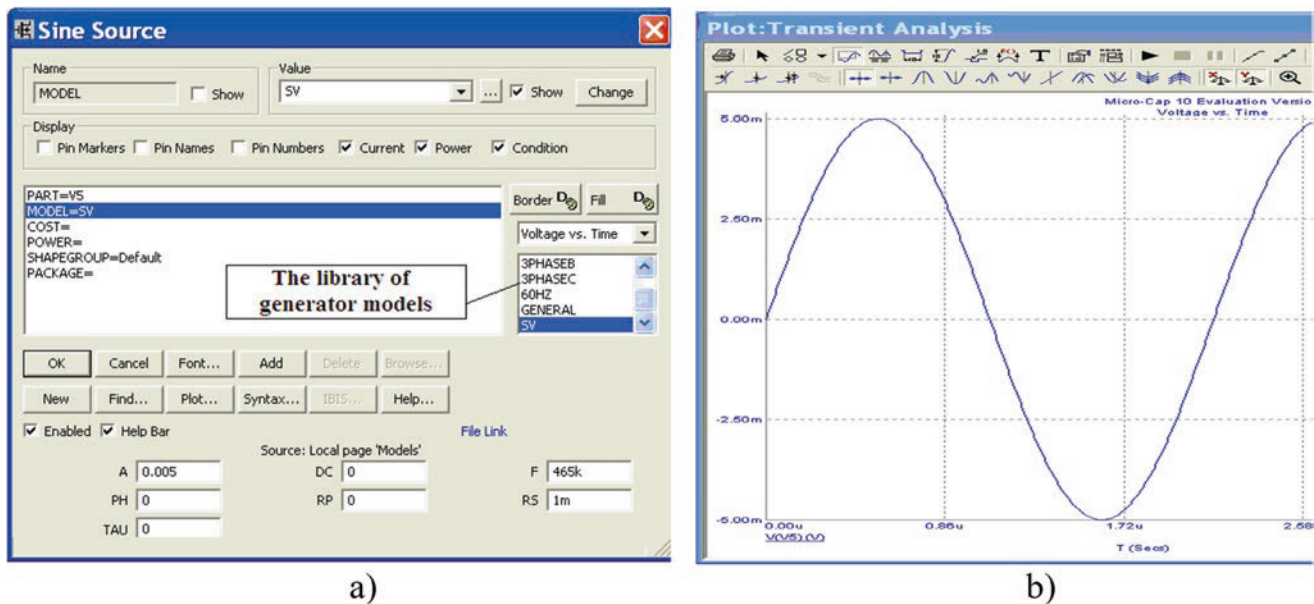


Fig. 6.10 Specification of EMF source model parameters and the signal form of formed voltage

The submenu *Sine Source* is built in the traditional form, which is used when entering basic components: resistors, capacitors, etc.

In the frame marked **Model**, data described in the left column of the left window is automatically repeated. The value of the selected component (in the *Value* frame) or the law of its change for each row of the left window can be selected in the right window or set independently. In the **Display** line, the indicators displayed on the monitor screen are marked.

The following parameters are specified in the first column of the left window:

Part – the positional designation of component assigned by the program;

Model – the name of component assigned by user;

Cost – the cost of component;

Package – the component case;

Power – the power dissipated by the component.

The buttons located below have an obvious purpose, indicated in the labels.

In the windows intended for entering the parameters of the described component, the values in the SI system are indicated.

If there are difficulties in describing more complex components, then it is necessary to refer to [28]. By hovering the cursor over an unknown attribute of the submenu, you can get brief information about it in the MC format. It should be noted that at execution of modeling in the frequency domain (AC), the voltage amplitude is always specified equal to 1 V, and the frequency changes within the limits indicated in the sub-menu AC Analysis Limits.

The generator of pulse signals is entered by the sequential choice of commands in the circuit window: Analog Primitives → Waveform Sources → Pulse Source.

The generator parameters are specified in the same manner as parameters of the harmonic signal generator. The shape and its characteristics for chosen parameters are shown in the monitor screen by pressing the button Plot in the sub-menu: Plot:Transient Analysis.

6.3.2 Models of Active Components

6.3.2.1 Models of Active Components in the MC12 Medium

The system MicroCap for circuit modeling uses the same mathematical models of semiconductor devices, which are applied in the PSPICE software, and is one of the software of family **Spice** (Simulation Program with Integrated Circuit Emphasis). At present, for the restricted number of Russian transistors, we know parameters of Spice models [121]. They can be simply enough calculated with applications of the MODEL software of the professional version of MicroCap.

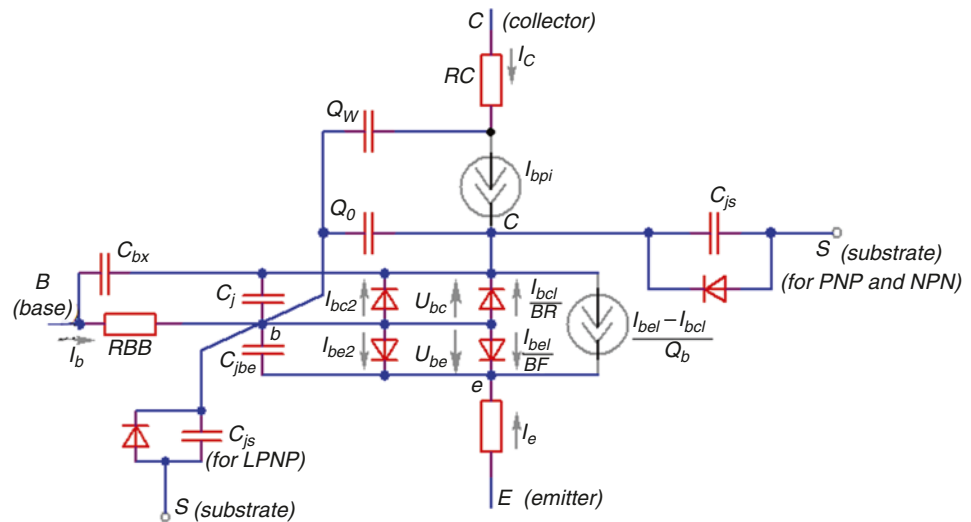
The Model of the Bipolar Transistor Application of the linear model of the ideal transistor for analysis of amplifiers properties is limited by the small level of the input signal, which practically does never correspond to the real operation conditions of active elements. Even at small levels of the useful signals in the input of the radio receiving device, transistors in first stages of the radio front end can be transferred in the saturation mode under conditions of the powerful interference (concentrated or pulse). Therefore, to obtain correct enough results at comparison of model properties and real transistors, it is necessary that the model takes into account the following:

- Variations of states of emitter and collector junctions depending on the polarity and the level of applied voltage: the transfer from the cutoff region into active region or the saturation region.
- Dependence of the diffusion capacity and the resistance of opened p - n junction upon the voltage.
- Dependence of the collector current upon the collector voltage at operation in the active area.
- Dependence of cutoff frequency f_{cut} on the collector current and voltage.

Since all mentioned effects are nonlinear, all parameters of the ideal transistor should be described by nonlinear functions. Evidently that solution of nonlinear equations describing the total model of the transistor, even the simplest unit, is possible only with the computer application.

Several models of the ideal bipolar transistor – Ebers-Moll, Gummel-Poon (Fig. 6.11) – reflect physical phenomena proceeding in the transistor: an injection of carriers by the emitter and the collector into the base area at opened p - n junctions; an accumulation of the minority charge carriers in the base area; their transfer from an emitter to a collector and from a collector to an emitter.

Fig. 6.11 The typical model of the ideal bipolar transistor reflecting physical phenomena proceeding in the transistor



This model is automatically simplified (if we omit several components of the Gummel-Poon model of the bipolar transistor of the n - p - n structure) to the Ebers-Moll model. *Remark:* model parameters are entered (are edited) in the sub-menu NPN:NPN Transistor of the circuit window of graphical editor or in the text library file with extension .lbr.

Mathematical expressions, from which the model parameters are defined, are presented in [28, 121] and depend on the material properties and the technology, which is used at manufacture of bipolar transistors.

So, for instance, models of the bipolar transistor of n - p - n type in MC11 (NPN), but having the horizontal structure (side, planar; Fig. 6.12a), include in description the isolating diode and the capacitance C_{js} (the capacitance of the collector-substrate junction), which connect the substrate unit with the internal base point, and for vertical (Fig. 6.12b) transistor structure – the substrate with the internal collector point (NPN4).

At utilization in the network of transistors or diodes from Russian manufacturers, which models are absent in the file SMALL.LBR, we must enter them independently (e.g., for 2T316D transistor; Fig. 6.13). The sub-menu NPN:NPN Transistor has the traditional attributes of description and control buttons. Parameters of the chosen transistor model are entered into appropriate windows belonging to the library SMALL.LIB and stored in the binary form.

Fig. 6.12 Models of the bipolar transistor of n - p - n type with planar (a) and vertical (b) structures

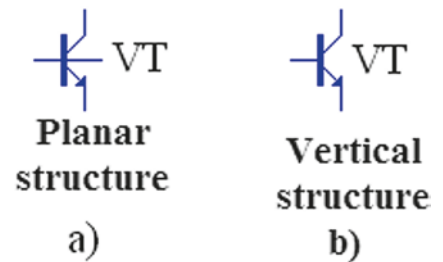
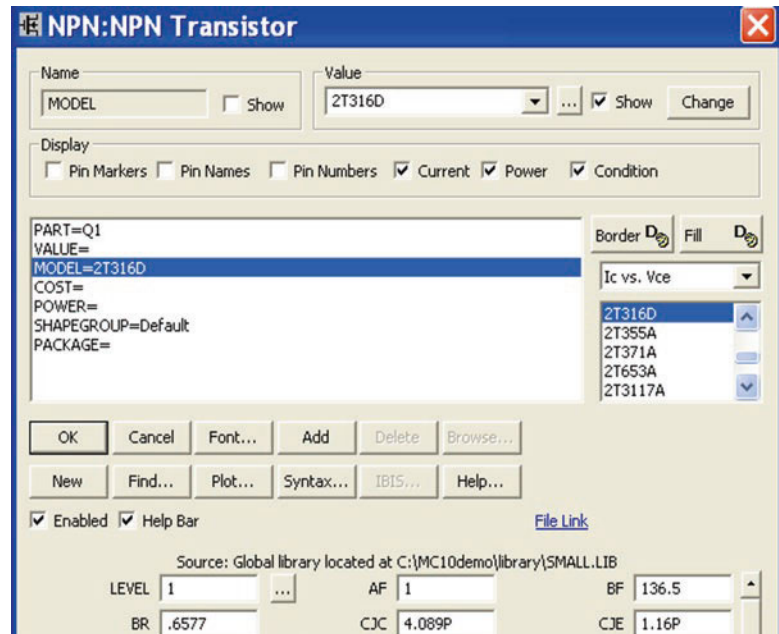


Fig. 6.13 Entering of new transistor parameters



After you have finished entering the parameters of bipolar transistor model, click OK. The transistor model can be reused in this file.

The Model of Field-Effect Transistor In FET, the portion of drain-source becomes conductive under influence of the control voltage applied across a gate and a source for absence of the gate current. In other words, in contrast to the bipolar transistor, the field-effect transistor is controlled without power expenses. The field-effect transistors can be realized with utilization of two different effects:

- In the *field-effect MOS transistor* (MOSFET) or with the isolated gate (IGFET), the gate is separated from the channel by the oxide layer from SiO_2 , which does not cause the gate current flow at voltage applying to the source and the gate. The control voltage regulates the charge carrier density in the inversion layer under the gate, which forms the conduction channel between the drain and the source. Depending on the channel doping degree, we form MOS transistors with enhancement or with depletion. In MOS transistors of the depletion type, at absence of the gate-source voltage ($U_{gs} = 0$), the drain current exists. In devices with enhancement at $U_{gs} = 0$, the current is absent.
- In the *field-effect transistor with controlling p - n junction* (JFET), the control voltage affects the structure of the barrier layer at cutoff voltage on the p - n junction, which leads to reduction of the area section, and hence, the channel conductance between the drain and the source. Since the control electrode does not separate from the channel, then at the direct-biased junction, the current flows through the gate (such a mode is not used). In the MESFET with the junction metal-semiconductor (the Schottky), the operation principle of JFET is saved and both types of transistors are used in the depletion mode, when at $U_{gs} = 0$ the drain current exists.

Transistors of MOS type and with controlling are *symmetrical*, i.e., the drain and the source are interchangeable.

Field-effect transistors with the controlling p - n junction and the channel of n -type **NJFET** (Negative **J**unction **F**ield-**E**ffect **T**ransistor; Fig. 6.14a) and of p -type (**PJFET**; Fig. 6.14b) are described by the Shichman-Hodges model with nonlinear equivalent circuit (Fig. 6.15).

Fig. 6.14 Field-effect transistors with the controlling p-n junction and the channel of *n*-type (a) and of *p*-type (b)

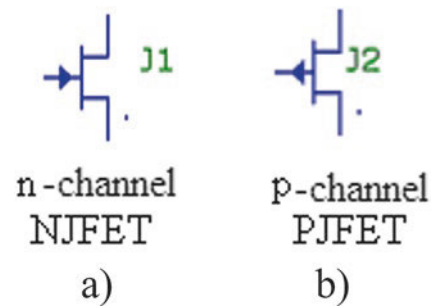
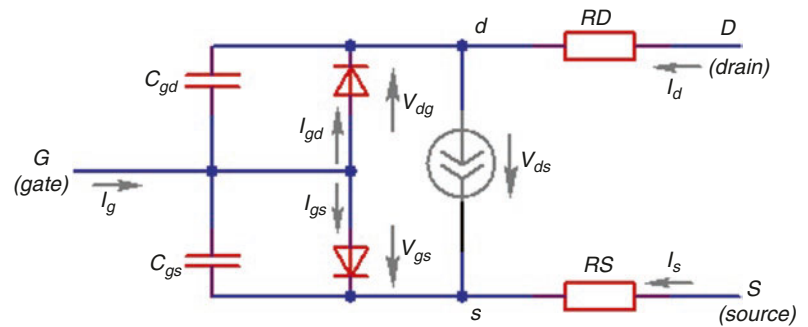
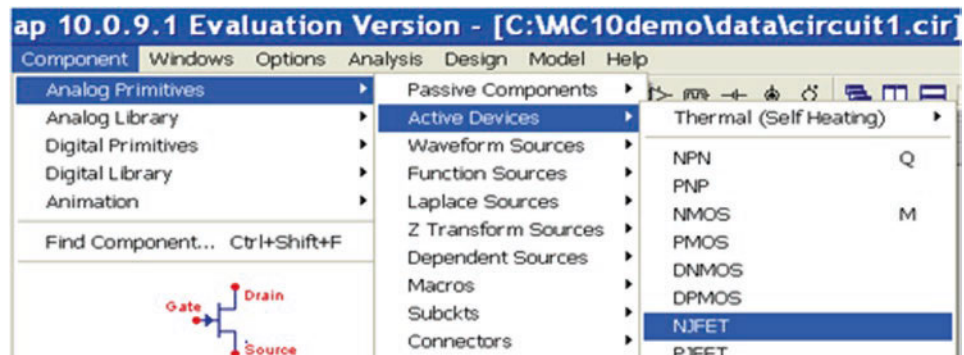


Fig. 6.15 The Shichman-Hodges model with nonlinear equivalent circuit, where $I_g = I_{gs} + I_{gd}$ is the gate current, I_{gs} , I_{gd} are leakage currents of gate-source and the gate-drain, and C_{gd} , C_{gs} are capacities gate-source and gate-drain



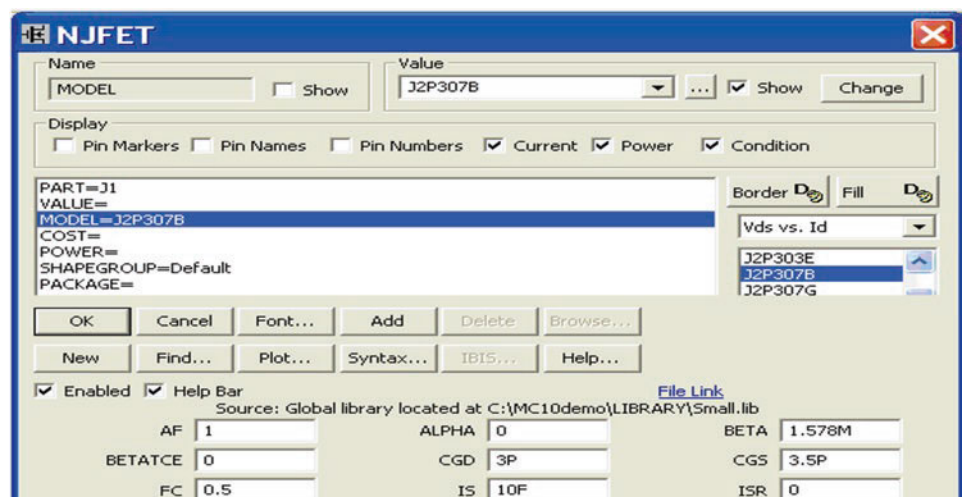
Conditionally graphical designation of FET (with the *n-p* junction and NJFET, PJFET) is absent in the line of main components; therefore, for entering of FET mathematical model parameters in the circuit window, the commands (Fig. 6.16) are entered sequentially – Component → Analog Primitives → Active Devices → NJFET – and in the appearing sub-menu NJFET (Fig. 6.17), we choose the transistor model (or we insert parameters of new model).

Fig. 6.16 Entering of FET mathematical model parameters in the circuit window



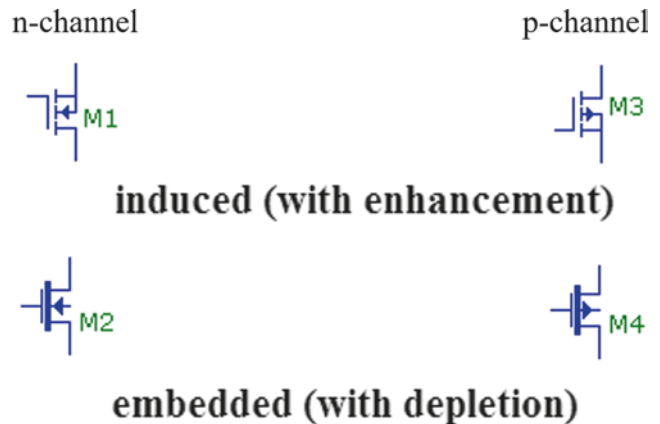
Attributes of the FET model have functions similar to bipolar transistors. The name, designation, and dimension of entering parameters of the FET model with the controlling *p-n* junction (**JFET**) can be found in [28, 121].

Fig. 6.17 The choice of the transistor model (or insertion of new model parameters)



The Mathematical Model of the MOS Transistor Depending on construction and manufacturing technology of MOS transistors, we must distinguish (Fig. 6.18) transistors with p - and n -channels subjects to the substrate type, which forms the channel.

Fig. 6.18 Depending on construction and manufacturing technology of MOS transistors, we must distinguish transistors with p - and n -channels



Among n -channel MOS transistors with isolated gate, we can distinguish:

- Self-closing (Fig. 6.19) in the enhancement mode ($U_{gs} = 0$ V, $I_d = 0$ A),
- Self-conducting (Fig. 6.20) in the depletion mode ($U_{gs} = 0$ V, I_d differs from zero A).

Fig. 6.19 The self-closing n -channel MOS transistor in the *enhancement* mode

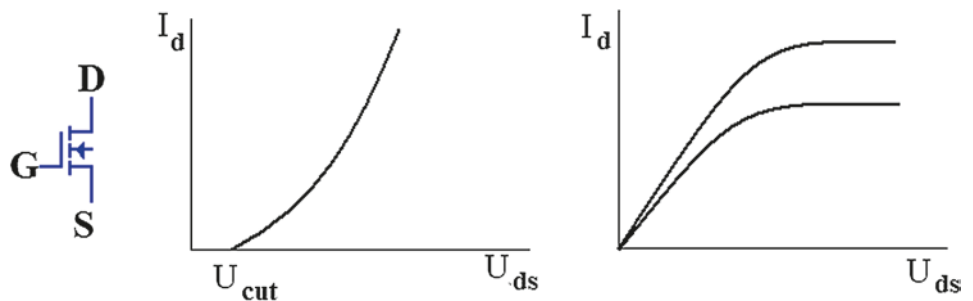
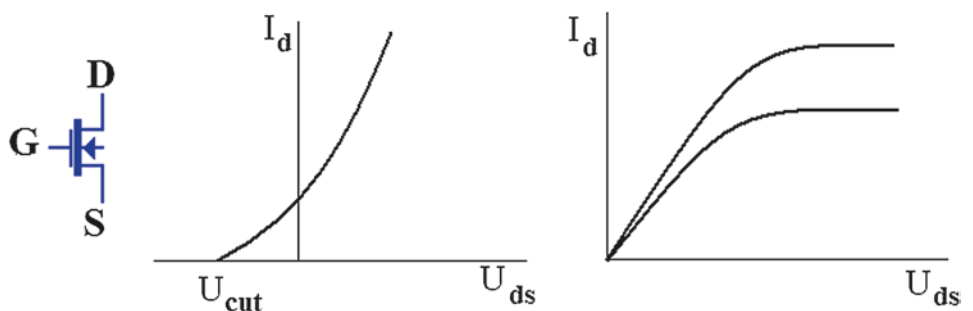


Fig. 6.20 The self-conducting n -channel MOS transistor in the *depletion* mode



Characteristics of p -channel MOS transistors with the isolated gate are similar, but they are inverted with regard to the transistor axis.

In contrast to bipolar transistors, for the FET, there is no universal model (or two models), which is suitable for any transistors and for any operation mode. The known Shichman-Hodges model usually is used for discrete MOS transistors with the large channel length and width. For modern MOSFETs with significantly smaller dimensions and short channel (especially in the integrated version), more bulky models of 2 or 3 levels or the BSIM model are used.

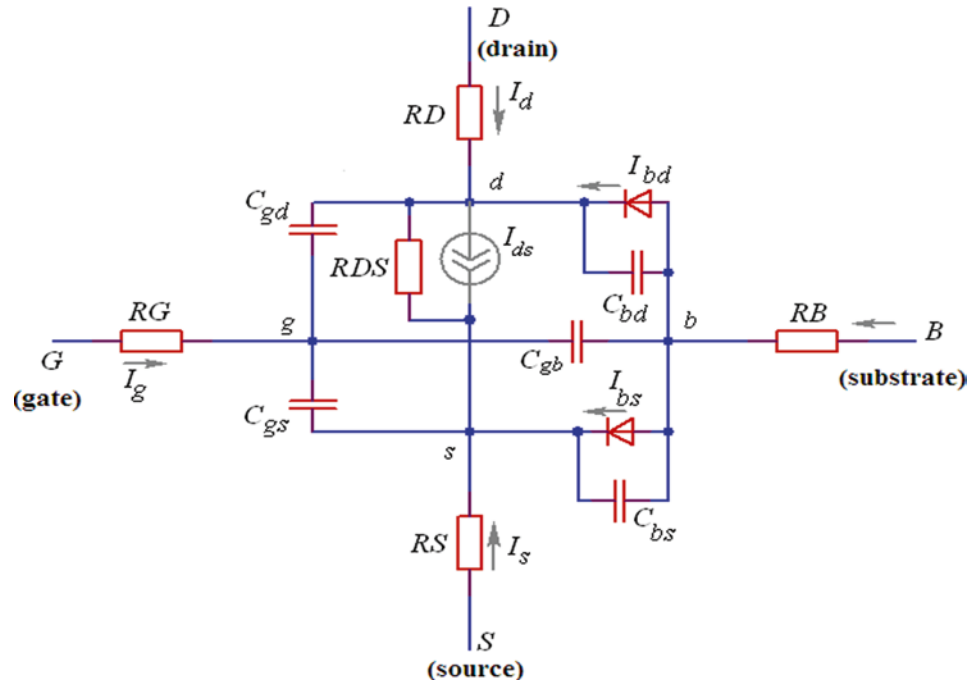
The nonlinear equivalent model of the MOS transistor with the n -type channel is shown in Fig. 6.21. On these figure resistors, RG, RD, RB, RS are the parasitic bulk resistances of the gate, the drain, the substrate, and the source.

Equivalent capacitances C_{bs} and C_{bd} take into consideration the dynamic capacitances of the depletion layer of substrate-source and substrate-drain, relatively. The spurious resistances RS, RD, RG, and RB can be also mentioned in the model parameters or to be calculated at specification of the specific resistance of the material and a number of squares of the source

and substrate regions. If these parameters are absent in the model description or are equal to zeros, they are not included into the model.

Mathematical model of isolated gate field-effect transistors (MOSFET), located in the file: Source: Global Library, located at: C:\MC12demo\library\Small. The NMOS transistor model is described by four different system of equations, the complexity of which is determined by LEVEL of the selected model.

Fig. 6.21 The nonlinear equivalent model of the MOS transistor with the n-type channel



The list of model parameters and their correspondence to components of the used model is mainly defined by technological parameters of the real transistor construction and can be found in [28, 121].

A series of MOS transistor models, which are supported by MC11 software, are known:

LEVEL Model

1. Shichman-Hodges
2. Grove-Frohmman MOS2 (Spice 3F5)
3. Empirical model MOS3 (Spice 3F5)
4. Original model for transistors with short gate (**BSIM1**-Berkeley-Short-Channel-Insulated-Gate FET-Model)
5. BSIM model of second generation **BSIM2** (Level 5)
6. BSIM model of third generation **BSIM3** (Level 8 or 49)
7. BSIM model of fourth generation **BSIM4** (Level 14)
8. Charging model of the device with sub-micron gate length EKV2.6 ($l < 0.1 \mu m$)

The simplest model of the MOS transistor of first level (LEVEL = 1) does not provide the enough accuracy in modeling and is used for rough calculation only. The model of the second level (LEVEL = 2) is based on more exact theoretical constructions; however, a series of its parameters is difficult to estimate on the base of experimental data. The semiempirical model of third generation (LEVEL = 3), at sufficient complexity of model parameter estimation, is more effective compared to the MOS transistor model of second generation, and it can be recommended for practical calculations when using devices with short channels. BSIM models permitting on the base of the small number of main transistor model parameters (the carrier mobility, the relative permittivity of a material, the specific oxide capacitance, etc.) to execute calculations of the first approximation have found the widest distribution. By default, if the LEVEL parameter is not specified at description of the MOS transistor model, then the model of the first level is used.

Mathematical models based on the calculation of their models according to performed measurements and known properties of the semiconductor structure and running physical processes in it (the doping level degree, the carrier drift speed, the transfer time, variation of the bulk charge region, etc.) can provide the high accuracy under condition of identity of the model parameters and the particular transistor parameters.

6.3.2.2 Linearized Equivalent Circuits at Investigation of Transistor Noises

The Equivalent Circuit of the Bipolar Transistor The mathematical models of active components used in the MicroCap11 software coincide with models on the PSPICE software and take into consideration thermal, shot, and low-frequency flicker noises.

At investigation of the noise properties of the bipolar transistors, the linearized equivalent circuit is applied with inclusion of the internal noise sources (Fig. 6.22).

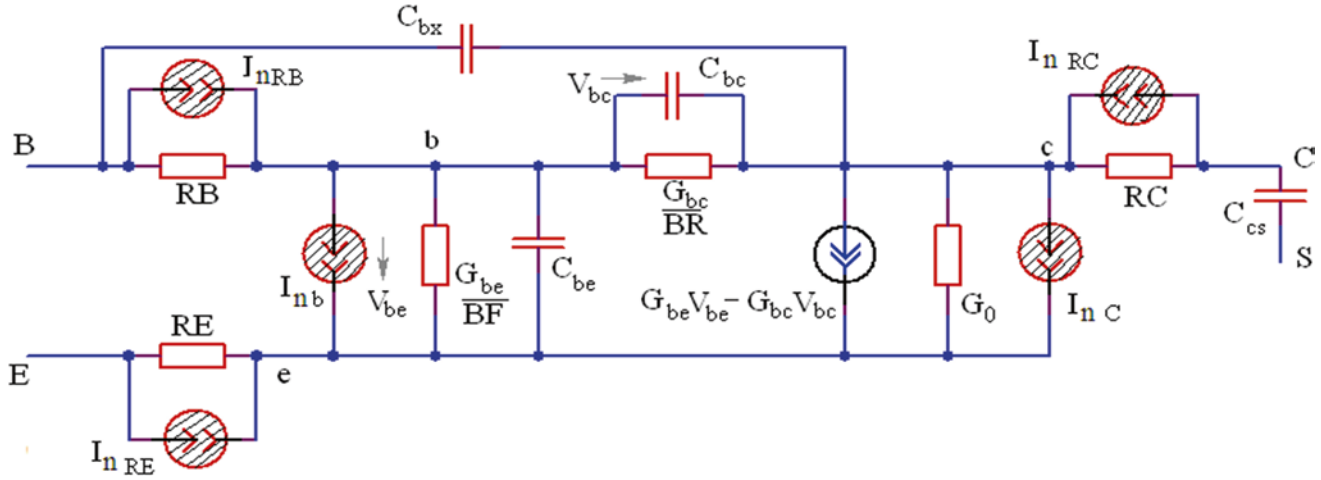


Fig. 6.22 The linearized equivalent circuit of the bipolar transistor with inclusion of the internal noise sources

As we see from Fig. 6.22, besides component of the transistor model (Fig. 5.11), the noise current sources reflecting processes running in the transistor structure are appeared. The circuit contains the fluctuation noise sources caused by the chaotic motion of charge carriers through structures having a resistance, which is reflected on the equivalent circuit of the bipolar transistors by the resistance of the structures of base R_B , collector R_C , and emitter R_E areas. Thermal noises presented in the figure by the current sources I_{nRB} , I_{nRC} , and I_{nRE} , have the noise spectral density:

$$S_{RB} = 4k \cdot T / R_B, S_{RC} = 4k \cdot T / R_C, S_{RE} = 4k \cdot T / R_E. \quad (6.1)$$

Noise sources characterizing the shot noise I_{nb} and the flicker noise I_{nC} in the base and collector circuits, relatively, have spectral densities:

$$S_b = 2qI_{nb} + KFI_{nb}^{AF} / f; S_c = 2qI_{nC}, \quad (6.2)$$

where $q = 1.6 \cdot 10^{-19}$ K is the electron charge, KF is the coefficient defining the spectral density of the flicker noise, and AF is the exponent defining the dependence of the flicker noise spectral density versus the current through the junction.

From results of spectral density calculation of the internal noise, we can easily calculate the differential noise factor of the two-port device (Fig. 6.23a).

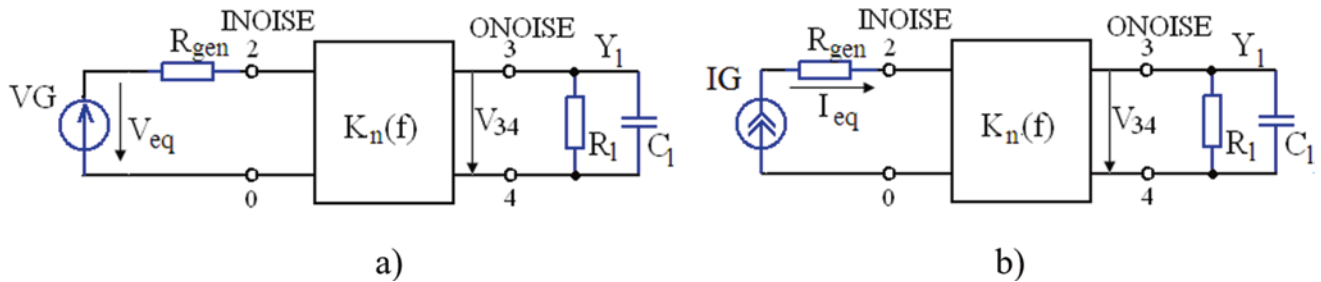


Fig. 6.23 From results of spectral density calculation of the internal noise, we can easily calculate the differential noise factor of the two-port

It is known from [118] that the differential noise factor is defined as:

$$N(f) = \frac{S_{u, in, eq}(f)}{S_{u, gen}}, \quad (6.3)$$

where $S_{u, in, eq}$ is the spectral density of the noise voltage caused by the thermal noise of the generator output resistance R_{gen} and the internal noise of the two-port reduced to its input $S_{u, in, eq}(f) = (INNOISE)^2$; $S_{u, gen} = 4kT_0R_{gen}$ is the spectral density of the noise voltage of the output generator; $T_0 = 300$ K is the nominal absolute temperature; and $k = 1.38 \cdot 10^{-23}$ J/deg. is the Boltzmann constant.

The value of the differential noise factor for mentioned conditions is:

$$N(f) = \frac{(INNOISE)^2}{1.656 \cdot 10^{-20} R_{gen}}, \quad (6.4)$$

where R_{gen} is in Ω .

In a similar manner, we calculate the differential noise factor for connection of the current source in the two-port input (Fig. 6.23b):

$$N(f) = \frac{(INNOISE)^2}{1.656 \cdot 10^{-20} / R_{gen}}. \quad (6.5)$$

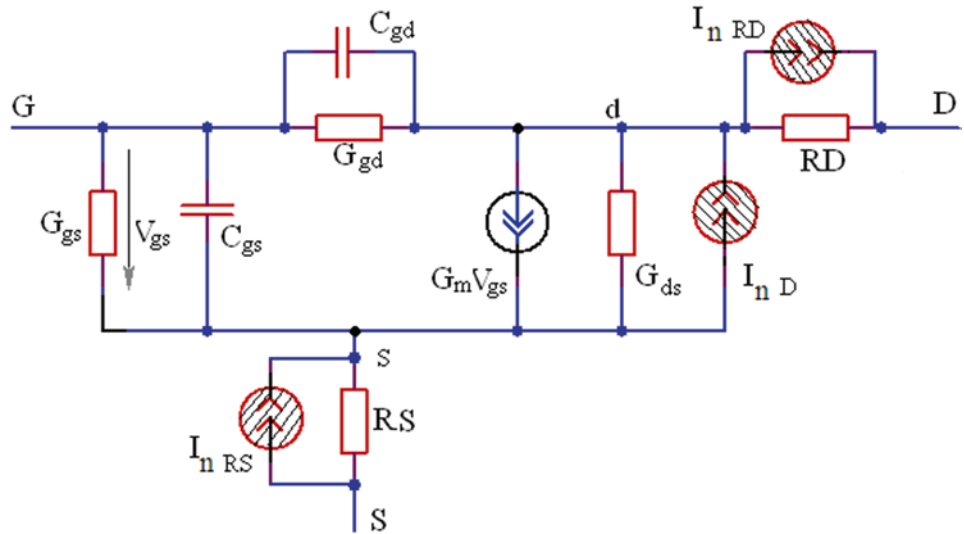
Noise properties of a stage are based on measurement of the total contribution of all noise sources acting in input and output of a circuit.

The Equivalent Circuit of FET with Controlling p - n Junction In the linear model of FET, besides components describing the nonlinear properties of the junction (Fig. 5.13), the fluctuation current sources are connected additionally (Fig. 6.24).

Thermal noises created by active components of the source (RS) and the drain (RD) are described by the current sources $I_{n, RS}$ and $I_{n, RD}$, relatively, which have the spectral density $S_{RS} = 4kT/RS$, $S_{RD} = 4kT/RD$.

The current source I_{nD} describing the shot and flicker noise has a spectral density $S_d = 8kTG_m / 3 + KF \cdot I_d^{AF} / f$, where $G_m = dI_{drain}/dV_{gs}$ is the differential conductance of the p - n junction in the operating point (in DC).

Fig. 6.24 In the linear model of FET, the fluctuation current sources are connected additionally



The Linearized Model of the MOS Transistor In the linearized model of MOS transistor (Fig. 6.21), a part of parameters is specified on the input language of the PSPICE software: L is the channel length, W is the channel width, AD is the drain area, PD is a perimeter of the diffusion drain region, PS is a perimeter of the diffusion drain region, NRS is the source area, NRG is the gate area, and NRB is the substrate area [121] are given in the attribute VALUE of the nMOS sub-menu. At that, parameters L, W, AD, and AS are setting by default with the help of directive .OPTIONS. The thermal noise sources are the

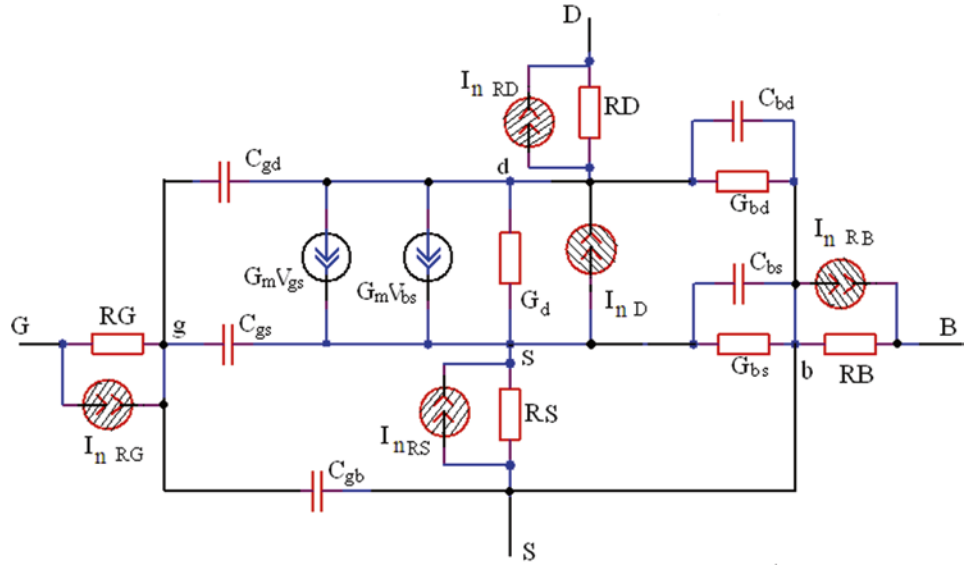
bulk resistances R_D , R_S , R_B (Fig. 6.25) are specified either in explicit form or are calculated with the help of known quantity of the specific resistance R_{SH} , for example: $R_D = R_{SH} \cdot N_{RD}$, $R_S = R_{SH} \cdot N_{RS}$, $R_B = R_{SH} \cdot N_{RB}$. Thermal noise sources created by grain, gate, and source resistances are described by current generators I_{nRD} , I_{nRS} , and I_{nRB} , relatively. Fluctuation noises created by each of the sources are characterized by the spectral density $S_{RS} = 4kT/R_S$, $S_{RG} = 4kT/R_G$, and $S_{RB} = 4kT/R_D$.

The shot noise and the flicker noise are characterized by the source I_{nD} having a spectral density for a MOSFET model below level 4:

$$S_D = 8kTD_m / 3 + KF \cdot I_D^{AF} (T \cdot K_{chan}),$$

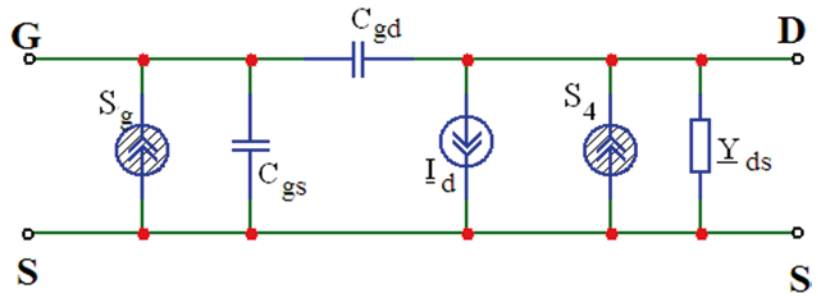
where $G_m = dI_D/dV_{gs}$ is the differential conductance in the operation point in DC; $K_{chan} = L_{eff}^2 EPS_{sil} / TOX$, TOX is the thickness of the gate oxide film, EPS_{sil} is a permittivity of silicon, $L_{eff} = L - LD$ is the effective channel length, LD is the region length of the side diffusion.

Fig. 6.25 The thermal noise sources are the bulk resistances R_D , R_S , and R_B



For analysis of the stage noise properties on MOS transistor, one often uses the simplified linearized noise model (Fig. 6.26).

Fig. 6.26 The simplified linearized noise model



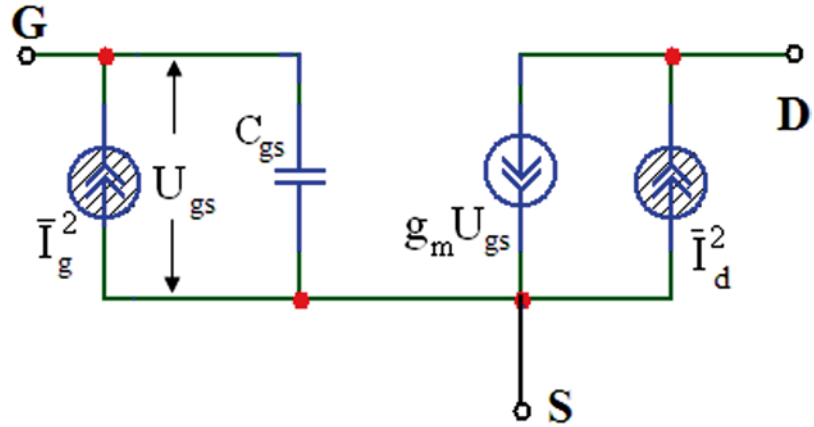
In the equivalent circuit of transistor connected in the common source circuit, besides components describing the capacitance between the gate and the source (C_{gs}) and between the gate and the drain (C_{gd}) and the controlled current source, which reflects the amplifying properties of transistor I_d , the noise sources are shown. The noise current source $S_g = S(f_1)f_1/f$ describes the spectral density of the redundant low-frequency noise. The value f_1 defines a frequency on which the redundant noise is equivalent in its spectral density to the spectral density of the channel noise. For transistors with the controlling $p-n$ junction, the frequency f_1 is usually 300 Hz...1 kHz, and for MOSFETs f_1 lies within 30...100 kHz. These values indicate to the essentially high level of the redundant noise compared to transistors with the $p-n$ junction, which restricts their application in the input stages of the receiver and defines its sensitivity. The spectral density $S_4 = 4kTg_{21}b$ reflects the thermal noises of the

channel, where g_{21} is the differential conductance of transistor direct transfer (the slope) in the operating point, and b is the constant multiplier, which characterizes the loss influence in circuits connecting to the transistor and changing the level of the thermal noise of the equivalent source. For FET with the p - n junction, $b = 1/3 \dots 2/3$, and for MOS transistors, $b = 2/3$.

In the nonlinear model of MOS transistors, we can usually neglect by noise parameters taking them as zeroes or they are simply absent in the model description.

The thermal noise model of the MOS transistor offered by Van der Zil, Shockley [118, 121], the simplified structure of which is presented in Fig. 6.27, has found the wide application.

Fig. 6.27 The simplified variant of the thermal noise model of the MOS transistor offered by Van der Zil



The model is described by the source of the thermal noise drain current $\sqrt{I_d^2}$ and also includes the source of induced noise gate current $\sqrt{I_g^2}$. The amplifying properties of the transistor model are presented by the current source, which is controlled by the voltage between a gate and a source (U_{gs}), and by the direct transfer conductance (the slope) of MOS transistor in the saturation mode.

The rms value of the drain noise current can be described as:

$$\sqrt{I_d^2} = \sqrt{4kT_0 \gamma g_0 \Delta f}, \quad (6.6)$$

where g_0 is the transistor conductivity at zero value of the drain-source voltage ($U_{sd} = 0$). The relation between transistor conductivities in the saturation mode g_m and at $U_{sd} = 0$ is expressed usually through the coefficient:

$$\nu = g_m / g_0, \quad (6.7)$$

which value depends on the channel length. The ν value is near to 1 for transistors with the long channel and decreases together with the channel length. The γ coefficient for a transistor with a long channel in saturation mode is $2/3$ and may be greater than one when the channel length decreases.

Uneven movement of charge carriers in the channel caused by thermal processes leads to the appearance of an induced current that creates random potential changes along the length of the channel. These variations, in turn, lead to fluctuations of its width and, hence, to variations of width of regions, which are depleted by charges along the channel. The existing capacitive coupling between a gate and a channel creates the gate fluctuation (noisy) current caused by random variations of the total charge of depleted regions along the channel:

$$\sqrt{I_g^2} = \sqrt{4kT_0 \delta g_s \Delta f}, g_s = \frac{\omega^2 C_{gs}^2}{5g_0}. \quad (6.8)$$

As we see from (6.8), the average value of the induced noise in the transistor gate increases with the frequency growth. The coefficient δ has the value $4/3$ for transistors with the long channel and increases with its reduction. The thermal noise of drain current causes the appearance of induced gate current that indicates the existence of a mutual correlation between them with the correlation coefficient c . Its value is calculated at description of the gate and the substrate in the form of the active distributed RC line, which permits to calculate the average value of noise squared currents $\bar{I}_d^2$, $\bar{I}_g^2$, $I_d^* I_g^*$. For MOS transistors with induced gate, the correlation coefficient is $c \approx j0.395$.

6.4 Command Menu "Analysis"

Executing the *Analysis* command in the *Schemes* window offers the user (in drop-down submenu) several analysis modes. The most popular among them are *Direct Current* (DC), DC circuit analysis, *Alternating Current* (AC) in the frequency domain, and *Transient* analysis in the time domain. Among other commands, *Dynamic direct current* allows you to calculate the model mode by direct current, *Dynamic alternating current* calculates the low signal parameters for selected frequencies, *Sensitivity* calculates the sensitivity of the circuit indicators to change the mode by direct current, and *Distortion* calculates nonlinear distortions that occur in the AE of amplifier.

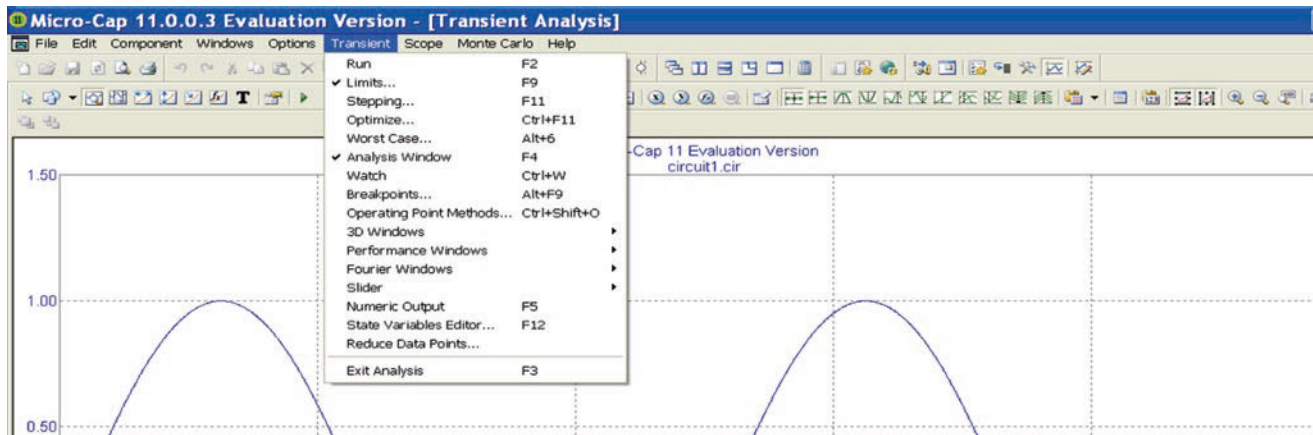


Fig. 6.28 A choice of analysis modes in DC in frequency or time domains

When selecting, for example, the *Transition* mode, a number of commands that the program offers in the drop-down submenu (Fig. 6.28) are the following:

Run – starting of the circuit analysis (on the right in the sub-menu, the “hot” buttons are shown).

Limits... – opens the Analysis Limits sub-menu.

Stepping... – opens the dialog window of the multi-variant analysis.

Worst Case... – the worst case a for chosen conditions and parameters.

Optimize... – opens the window for specifying the optimization parameters.

Analysis Window – leads to opening of the window with the image of calculated plots (*it is switched on automatically after pressing on the Run button*).

Fourier Windows – opens the **FFT (Fast Fourier Transform)**, dialog box, which specifies the boundaries of the time interval (*Upper Time Limit*, *Lower Time Limit*) and the *Number of Points* to be selected in range from 2^{10} to 2^{20} , which is multiple of base 2.

Slider – is used for direct and indirect variation of the component values or the model parameters of the circuit under investigation with the help of the cursor.

Exit Analysis – exit from the analysis mode.

6.4.1 Menu of Calculation Modes of Transients

After a choice of the *Transient...* command, the software transfers to the sub-menu *Transient Analysis Limits* (Fig. 6.29) by means of entering the analysis parameters for transient study.

Buttons in the upper sub-menu **Transient Analysis Limits** provide:

Run – enter in the analysis mode (launching of the modeling process).

Stepping – transfer in the mode of parameter varying (or the component, the model).

PSS... button (Periodic Steady State) allows enter in the dialog window PSS for calculation of nonlinear and intermodulation distortions after achievement of steady state of transients defining by parameters mentioned in the PSS window.

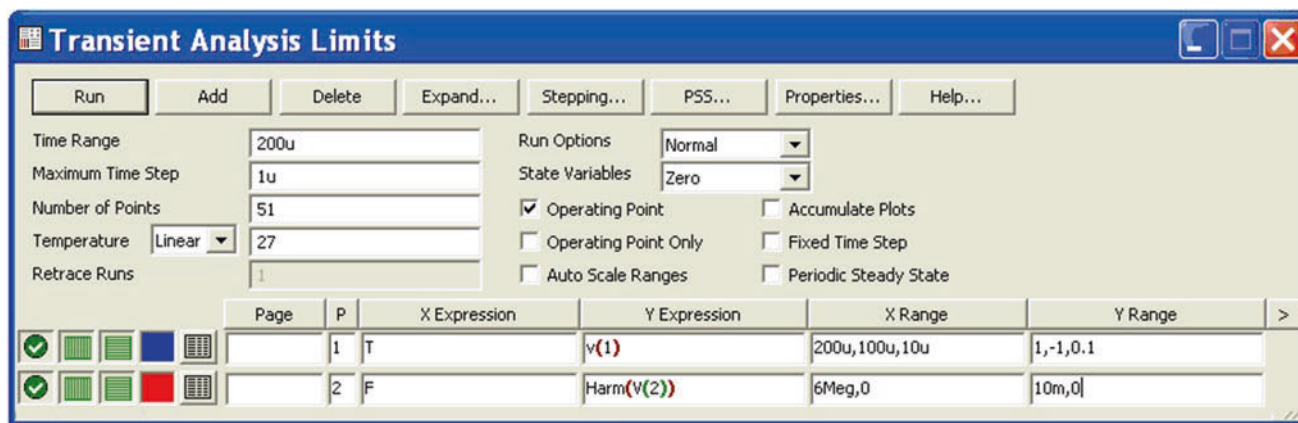


Fig. 6.29 After a choice of the Transient... command, the software transfers to the sub-menu Transient Analysis Limits

Properties... – allows parameter changing of information indicated in the monitor.

Help... – addressing to the description the system of circuit modeling MC10, which helps to be oriented with the destination of commands directly from the dialog window Transient Analysis Limits.

To the left, **under the panel line** (buttons that define the device analysis conditions in the time domain), enter the analysis limits and a graphical or numerical representation of the results obtained:

Time Range – the final and initial time for the transient calculation is indicated (by default, initial time for calculation is taken equal to zero).

Maximum Time Step – is the value of the maximal integration step. The current value of the varying integration step is defined by the value of admissible relative error.

Number of Points – a number of points indicated in the form of a table (by default, it is taken as 51).

In the column on the right, under the panel line, the **control options** of the calculation results are located.

Run Options – modes are defined:

Normal – calculation results are not stored.

Save – results are saved in the binary file of the format: <file name>.TSA.

Pressing on the button *Operation Point* provides execution of the DC calculation before each calculation of transients.

In the mode *Operating Point Only* – the calculation is executed only on DC.

The click on the button *Auto Scale Ranges* provides the automatic selection of the axis scales.

A mark in the line *Accumulate Plots* permits to store calculation results in the form of plots after each edition of the circuit. The buffer cleaning from the graphical material is performed via the command  in the line of result processing in the circuit window.

A mark opposite to *Fixed Time Step* allows providing the transient analysis with the permanent (maximal) step. This function can be turned out as usable at the initial research process although it is used rater rarely. In practice, we can use the changing step, which dimension is given by the error value defined by the throw-away terms.

The *Periodic Steady State* option uses the “target method” in order to eliminate the transients influence and to analyze the steady states only. The steady-state processes are especially important for obtaining accurate results at the harmonic analysis and at estimation of intermodulation distortions and the switching modes of the radio electronic devices.

Conditions, which define the modeling **result representation** in the time domain, are specified in the lower part of the dialog window Transient Analysis Limits. In the extremely left column of the dialog window, we can choose the one from three symbols defining the representation approach for calculation results (Table 6.5).

In the column *P* (Plot Group), the number of the graphical window is shown, where the curve will be plotted. Description of variation limits of exited curves and their character are described in the line. At simultaneous representation of several curves in the same coordinate system, the scale is chosen automatically according to the maximal value of the exited variables.

In the column *X Expression*, the variable name is presented, which corresponds to the X-axis.

When studying transients along the abscissa axis, time (T) is postponed. When calculating the signal spectrum using the Fourier transform or when calculating the amplitude frequency response along the abscissa axis, frequency (F) is postponed. In the column X Expression, the maximal (t_{max}) and minimal (t_{min}) values of the variable X, as well as the distances between the lines of the coordinate grid on the graph, are shown. (We can omit t_{min} , when $t_{min} = 0$).

Table 6.5 Graphical representation of the analysis results

Designation	Executed functions
	Exit of calculation results to the monitor screen with storage in the memory with extension .ano
	Calculation does not perform
	Calculation is performed without results' exit on the display but with results' storage
	The scale of exiting on the display screen variables on axes X and Y is given according with the logarithmic or linear laws. The choice of the scale is performed at pressing on the appropriate button
	At pressing on this button, we can choose a color of the curve, which is indicated in the screen
	This allows you to get the calculation results not only in the form of curves on the monitor screen, but also in the form of a text file containing a table of counts of the function specified in the YExpression graph. At pressing on this button, we can define a number of points which are exiting on the printer (their number is indicated in the line number of points)
	Exit of plots in the Cartesian coordinate system
	Exit of plots in the polar coordinate system
	Exit of plots on the circular diagram (the Smith diagram)

In the column **Y Expression**, the expression is presented along the Y-axis. This may be a voltage in any node, a current through any element or across any nodes, as well as the full expression. We can choose the variable type and the form of its representation by the click of the right mouse button in the falling sub-menu. In the column Y Range, we may give the maximal and minimal values of variable Y on the plot; if the minimal value is zero, we are able not to mention it.

The transfer to the mode of properties analysis of some circuit in the time domain is provided by pressing on the button RUN, and for conditions of graphical representation of results indicated in the sub-menu (Fig. 6.29), it will lead to construction of the voltage plot in the node V(1) in the time interval 100–200 ns (the process in the device characterizes the steady-state mode). In this case, the plot of transients in the node will not represent the voltage transient process from the moment of “switching on” of the circuit power supply ($t_{\min} = 0$).

To extend the software possibilities (for instance, the amplitude spectrum calculation in the node V(1)), we must use the button Properties (Fig. 6.29).

Pressing on the button Properties... opens the dialog window and transfers the operation in the sub-menu **Properties for Transient Analysis** containing eight bookmarks, and the most important among them are (Fig. 6.30):

Plot – the control to plot exiting onto a screen and the control to the coordinate system.

Scales and Formats – the scale choice in coordinate axes.

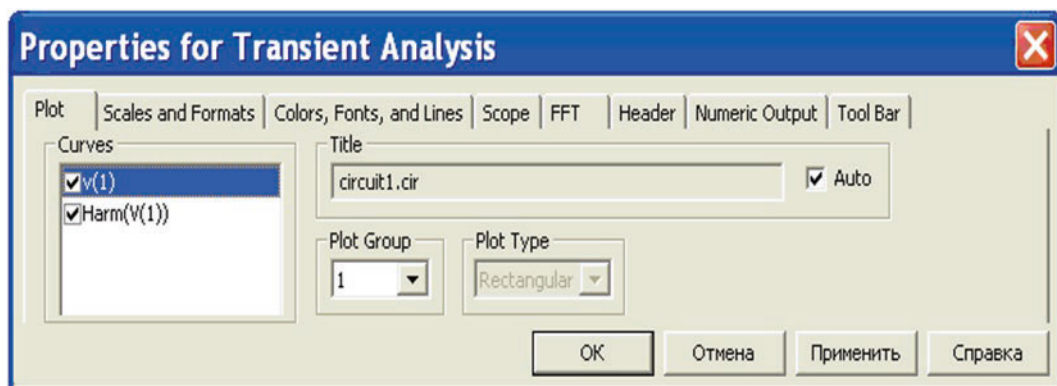
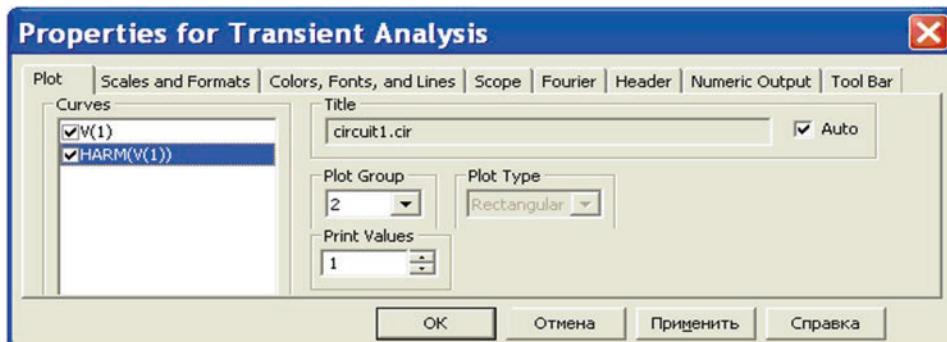


Fig. 6.30 Pressing on the button Properties... opens the dialog window and transfers the operation in the sub-menu Properties for Transient Analysis

Being in the sub-menu **Properties for Transient Analysis**, besides obtaining the time curves, we can perform the calculation of nonlinear distortions and the spectrum in any point of the circuit under investigation (Fig. 6.29) with the help of fast Fourier transform.

Let us consider (as an example) the actions' sequence, which should be performed for construction of the voltage spectrum in some node V(1) with utilization of FFT (Fig. 6.31; Harm (V(1))).

Fig. 6.31 Construction of the voltage spectrum in some node V(1) with utilization of FFT



Very often, for correct calculation of the spectrum of the harmonic coefficient, we must reject the initial part of the transient mode in active components including the connecting generators. For utilization for the spectral analysis of the steady-state mode only, we must mark on the bookmark *Fourier* in the sub-menu **Properties for Transient Analysis** in the line *Lower Time Limit* the part only (finite, several periods) of analyzed time-process (Fig. 6.32). For example, to analyze the spectral properties of a source of harmonic vibrations with an amplitude of 1 volt and a frequency of 1 MHz, the Fourier analysis was performed.

The essence of other parameter values, which are given on the *Fourier* bookmark, may be easily understood from explaining notes.

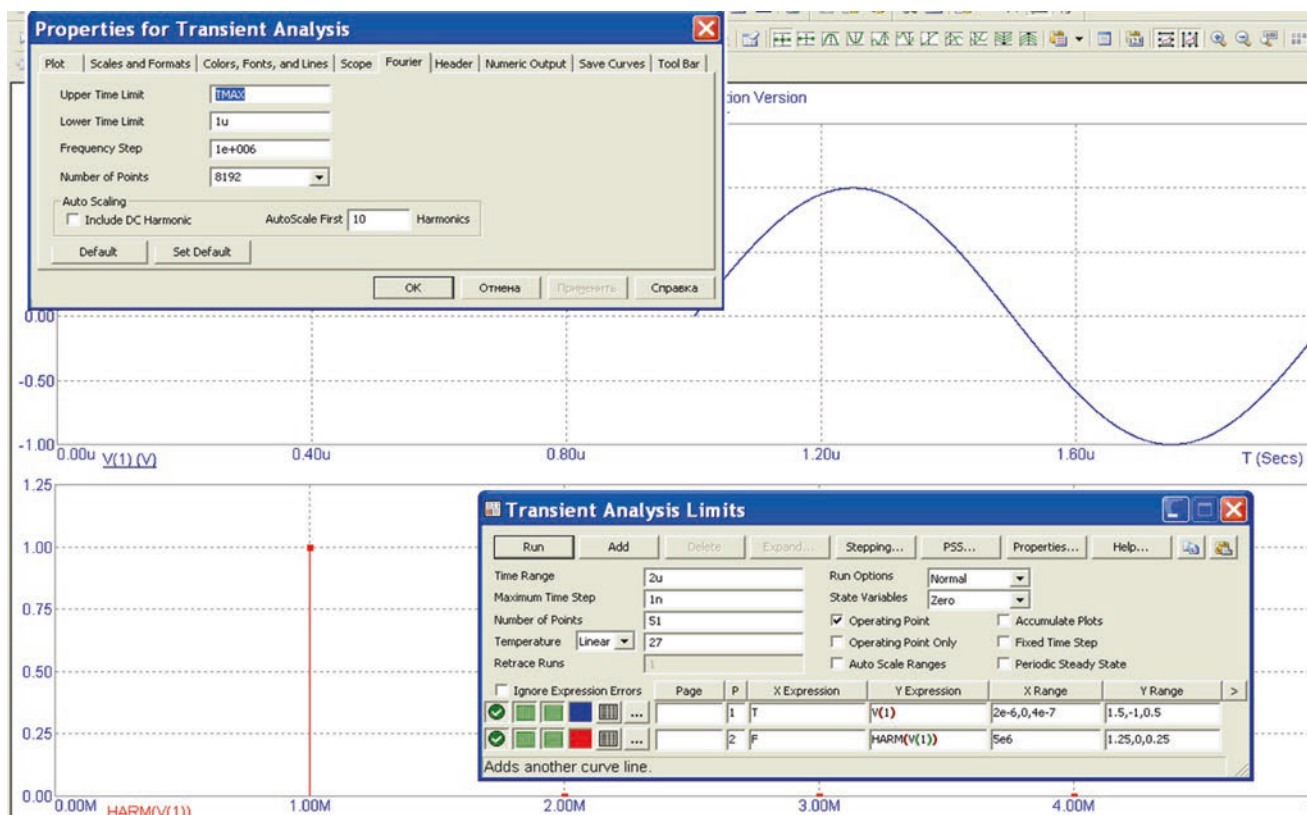
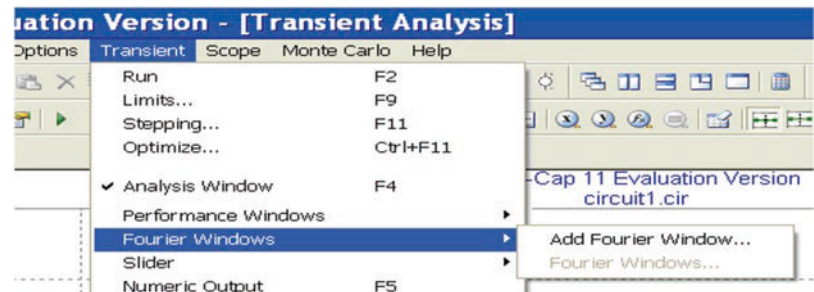


Fig. 6.32 An example of modeling of the time plot and the amplitude spectrum for the voltage steady-state mode in the node V(1)

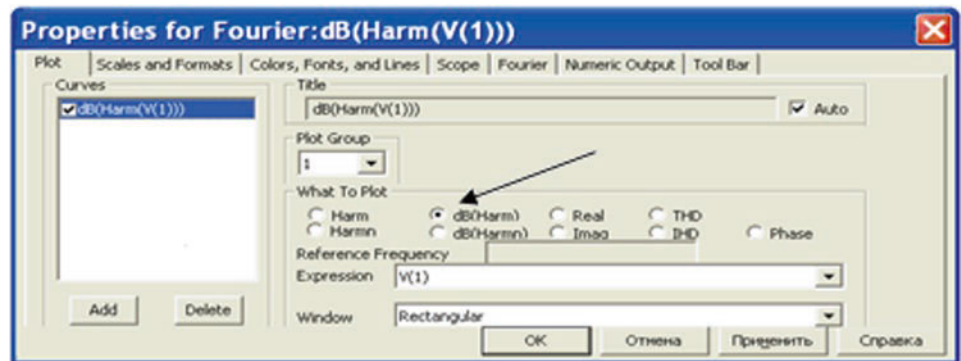
We can calculate the spectrum in the node V(1) on the base of the calculated time function using the Fourier Windows procedure (Fig. 6.33). To do this, in the schematics window, select sequentially: Transient → Fourier Windows → Add Fourier Window.... Values of spectral components in log-scale dB(Harm(V(1))) or as the amplitude spectrum can be shown in the monitor screen specifying conditions in the sub-menu *Add Fourier Windows*.

Fig. 6.33 Switching to setting parameters for the Fourier transform



In the bookmark Plot of the sub-menu **Properties for Fourier: dB(Harm(V(1)))** by pressing the button *dB(Harm)* (Fig. 6.34), we choose the logarithm form of the spectrum components representation.

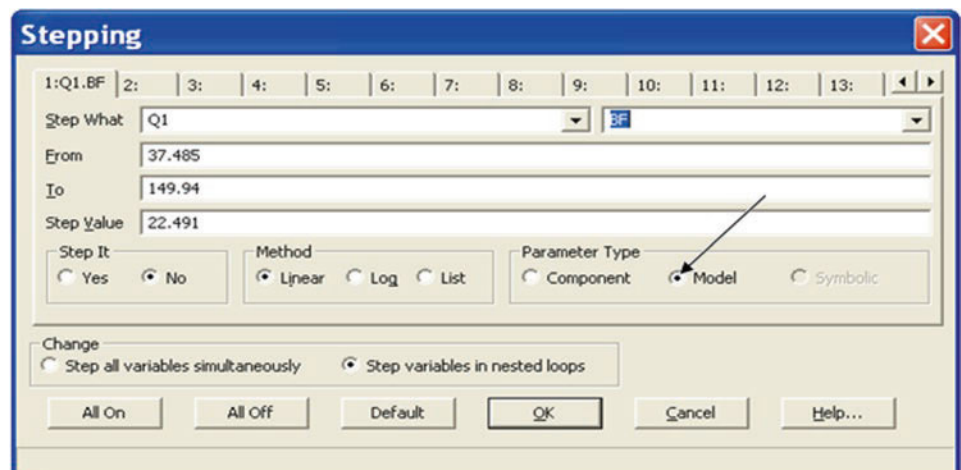
Fig. 6.34 The choice of the logarithm form of the spectrum components representation



Representation form of the amplitude spectrum plot is chosen in the bookmark *Colors, Fonts, and Lines* (Fig. 6.34).

One of the points of the sub-menu Transient Analysis Limits (Fig. 6.29) is the dialog window **Stepping** similar for all illuminating commands Analysis of the circuit window, which provides the variation of components or its parameters according to chosen conditions (Fig. 6.35).

Fig. 6.35 Specification of conditions for investigation of the effect of the BF parameter variation on output circuit parameters in the time domain



At first, we choose the bookmark (for instance, 1); then the component, which will be varied (e.g., the transistor Q1), is chosen and the parameter BF-Beta Forward. In this case, it is necessary to mention in the bookmark Parameter Type that it is a model (to set the radio button near the note Model). Having indicated the limits of parameter (component) variations, we choose the variation law (Method) and confirm the chosen mode for stepping (Step it → Yes).

6.4.2 The Menu of the Calculation Mode of Frequency Response

The analysis of electronic circuit characteristics in the frequency domain assumes the utilization of simplified (true at small signals) linearized models of all circuit components. If concerning to linear components (R, L, C) this is always true in the first approximation, then the active components (diodes, bipolar transistors, FETs, MOS transistors) are replaced by linear models. For instance, the nonlinear Gummel-Poon model of the bipolar transistor, which is true for any levels of input signals, is replaced by the linear model. In this model, the nonlinear components of transistor junctions described by diodes (Fig. 6.11) are replaced by constant values of a capacitance and a resistance of junctions, which are defined by voltages created by the DC (I_{b0} , I_{c0}) on the appropriate p - n junction. It is clear that this is true only under condition of amplitude smallness of the state variables (components of a current and a voltage) on appropriate junction.

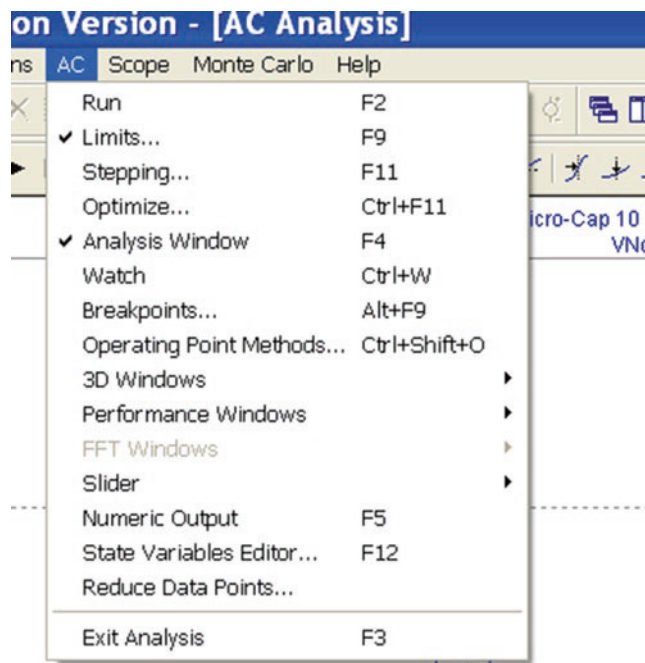
Such a linearized system is described by the system of algebraic equations with complex coefficient, which depend on the frequency. The complex character of coefficients is caused by inertial components (capacitors, inductors) including in electronic devices.

The indicators of electrical model of DC circuit (voltages in nodes and currents in chains) precede any other calculations and allow, first of all, to make sure that the required position of working point of the active elements is reached. The program does this automatically when the Analysis command is selected, and tests of real laboratory layouts (“smoke check”) always begin with this. When an analysis in the frequency domain is performed with utilization of the values of state variables, which are preliminary recorded in the hard disk and obtained from the previous calculations, then the calculation in DC was already performed in the beginning of the previous analysis. The list of assumptions at application of the small-signal analysis can be found in [28, 121].

The command menu used to calculate the frequency characteristics of an electrical circuit contains a list that is presented in drop-down submenu (Fig. 6.36) and appears when selecting the AC command of the circuit window is selected.

A transfer to the mode of frequency response analysis is fulfilled at the sequential command execution from the circuit window Analysis → AC... (Fig. 6.1), after which the software checks the correctness of the circuit formation. If errors in the circuit are absent, the software forms its topological description, reloads models of used components, performs its linearization, and opens the window for parameters specifying for AC modeling Analysis Limits (Fig. 6.37).

Fig. 6.36 The list of functions executing in the sub-menu AC of the command Analysis in the circuit window



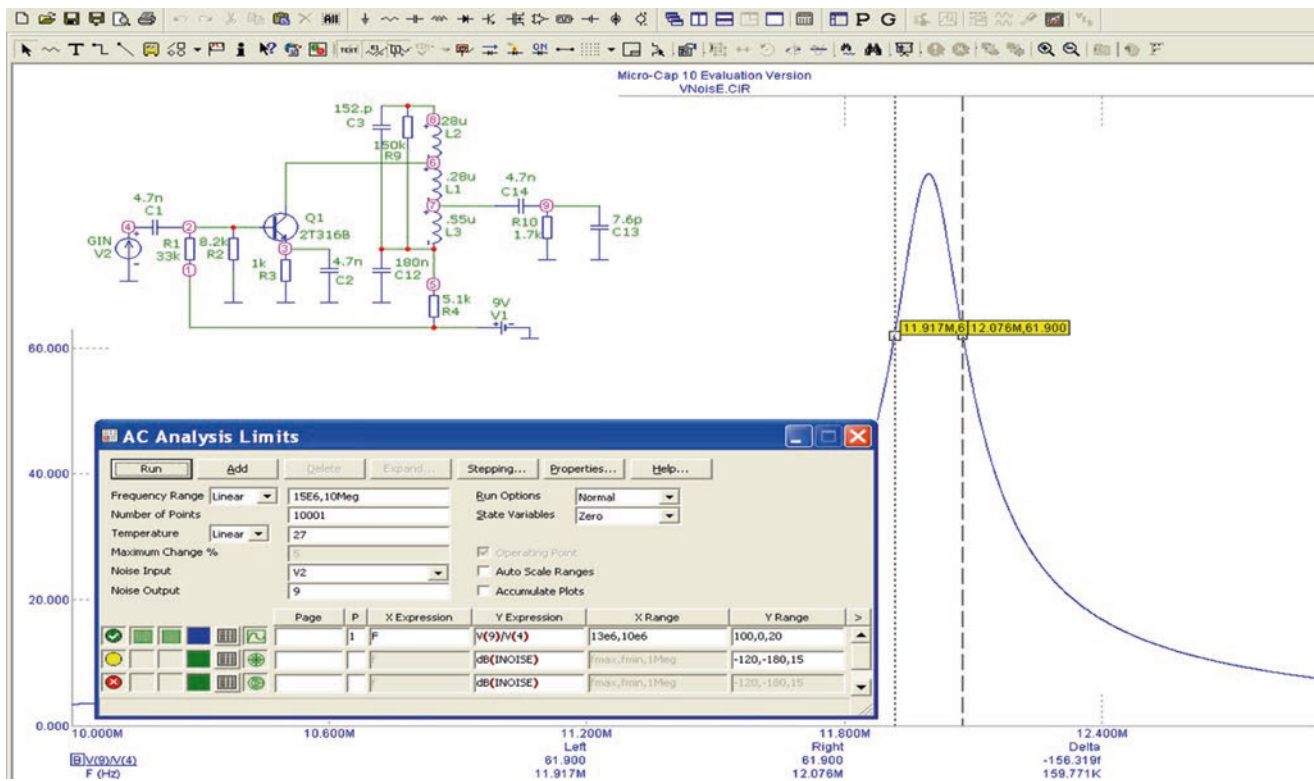


Fig. 6.37 The window for parameters specifying for AC modeling

Functions of buttons in the upper line **AC Analysis Limits** are the same as buttons in the sub-menu, which are used in the time domain.

To transfer into calculation mode, we can by a click at the pictogram ► in the instruments line of the schematic window or by pressing on the key F2, when you are in the result window (Fig. 6.37). To stop modeling, we should press the pictogram ■ or the button Esc being in the result window; to hold up calculation, we must press on pictogram ||. The second pressing on this pictogram continues the interrupted calculation.

Frequency Range in the left column of the expressions field of the dialog. **AC Analysis Limits** sets the low of frequency change (a small field on the left: *Auto*, *Linear*, *Log*, *List* – according to the list). The analysis limits (right column) are described in format: F_{max} – maximal frequency value, F_{min} – minimal frequency value in the analyzed range.

With automatic (*Auto*) selection of the step of splitting the interval of the analyzed frequency band, the next step is selected based on the results of calculations obtained when solving the equations of state for the first graph. If the calculation results result in error exceeding the percentage change specified in the <Maximum Change,%> line, then the step (frequency interval change) is reduced – and vice versa.. The boundary in the next interval frequency is defined by interpolation, and the number of steps in the next interval is fixed: 51.

Number of Points specifies the approaches of frequency interval division and a number of points in frequency, in which the calculation of frequency response is performed at the choice of the frequency variation law in the window *Frequency Range*:

Linear – with the uniform scale; *Log* – with uniform logarithm scale; *List* – according to the list.

Temperature – the temperature of environment, at which the analysis is performed.

List – the list of temperatures, at which the analysis is performed; the list is mentioned in the right column of the line Temperature.

Maximum Change – is the limited value of variation speed of the fundamental harmonic in percents (used only at the choice of the automatic step variation (*Auto*) in the line Frequency Range). It is necessary to obtain the smooth curve.

Noise Input – in this field, the source name is indicated, where the spectral noise is determined. If this field is not marked in the list of expressions as exiting into the monitor screen (the symbol ☐ in the result line), it ignores.

Noise Output – in this field, the node number is shown, in which the noise spectral density will be calculated. If variables Noise Input and Noise Output are not marked as exiting onto the monitor screen in the column YExpression, then commands from these fields ignore.

Options of the AC Analysis Limits window are located on the right from the field of numerical limits (Frequency Range).

Run Options – the control of calculation results registration. By pressing on the arrow of the scroll bar, we can choose the mode:

Normal – launching of modeling without storage results on the hard disk.

State Variables – shows which values are conferred by the state variables before the circuit analysis in DC (if this calculation is anticipated) from the previous session, which are taken as initial values of the state variables.

Operating Point – the mark settled in the line shows on unconditional fulfillment of calculation in DC without taken into account the option selected in the line State Variable.

Auto Scale Ranges – the choice of automatic scaling on the X- and Y-axes for all plots exiting in all windows.

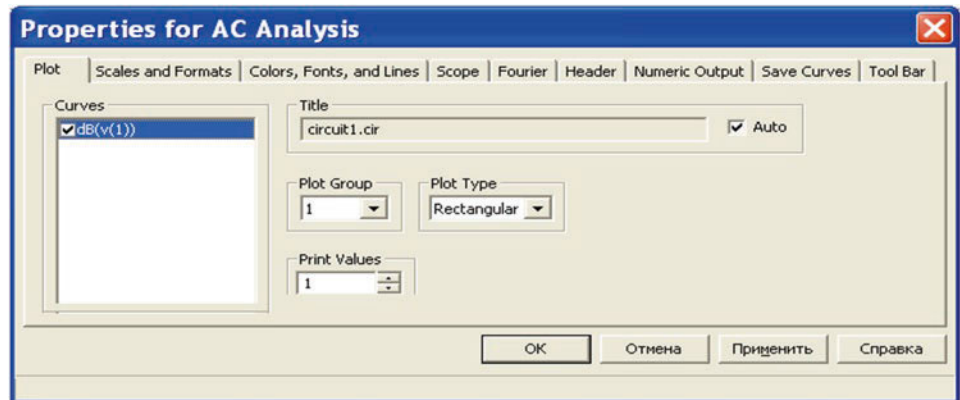
Accumulated plots – the mark settled in this line forms a file in which calculated curves are stored, which obtained after each circuit editing.

Content of the dialog window *Properties* in some often used bookmarks in the sub-menu **Properties for AC Analysis** has a form (Fig. 6.38):

Plot – the control of plot exit onto the screen and coordinate system.

Scales and Formats – the scale choice in coordinate axes is chosen in the same manner as for analysis in the time domain.

Fig. 6.38 Content of the dialog window *Properties* in some often used bookmarks in the sub-menu **Properties for AC Analysis**



At necessity to calculate the current or voltage spectrum in any circuit or the unit, we can use the bookmark (Fig. 6.39) **Fourier** (fast Fourier transform) in the sub-menu **Properties for AC Analysis**.

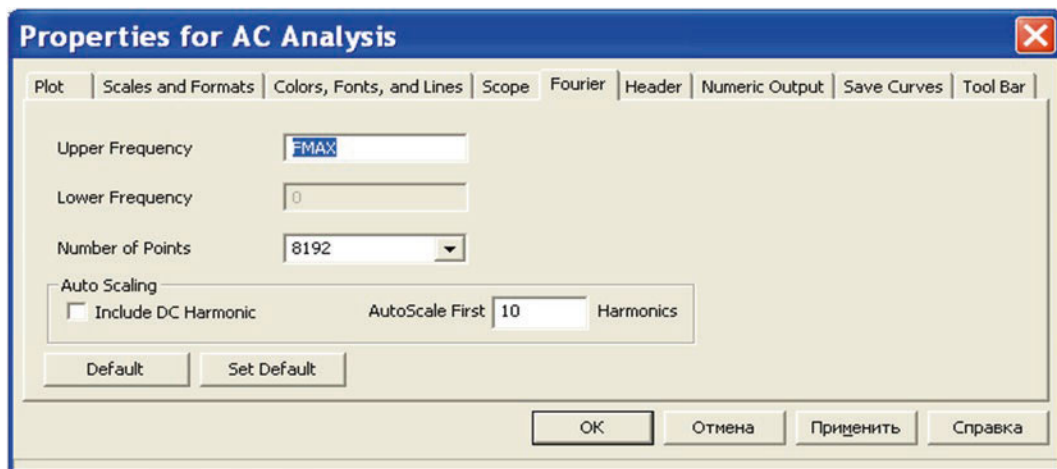


Fig. 6.39 The bookmark *Fourier* (fast Fourier transform) in the sub-menu **Properties for AC Analysis**

On the bookmarks, we can edit:

Upper Frequency – the value of the upper interval frequency included in analysis.

Lower Frequency – the lower boundary frequency (usually chosen as zero, if another value is not shown in the line Frequency Range in the sub-menu AC Analysis Limits).

Number of Points – shows a number of intermediate points in the analysis interval. A choice of the number of points can be arbitrary, but usually we can choose 1024, 2048, or 4096 points.

Auto Scaling – (automatic choice of point number) in the analysis interval can include:

Include DC Harmonic (marked) – the DC amplitude of the process expansion in the Fourier series together with calculated harmonics (multiple to the fundamental frequency).

AutoScale First – only first harmonics (for instance, 10 harmonics).

The calculation procedure can be formalized (by pressing the button *Default*) by a choice of standard conditions or *Set Default* – the refusal from them and the choice of necessary number of harmonics defined by the required accuracy of calculation.

For information extension about possibilities of the MC11 software, you must call to the Help file in the main window using command HELP or in any of sub-menu: AC Analysis Limits, Transient Analysis Limits, DC Analysis Limits.

6.4.3 Analysis of Transfer Functions in DC

Transfer into the analysis mode in DC is performed by execution of commands: Analysis → DC (Fig. 6.8). It should be noted that in the latest versions of the MicroCap program, in the sub-menu for setting the parameters of active components, it is provided to display the static I-V characteristic of the transistor on the monitor screen (e.g., the family of output characteristics of the BT: I_c vs. V_{be} ; Fig. 6.13). However, such characteristics are illustrative and do not reflect the actually used BT modes at low collector currents, which are typical for the receiver input stages. Therefore, the study of the properties of a cascade on a direct current is a very important task to exclude the transition of AE to a nonlinear mode.

One or two independent DC voltage or current sources (Fig. 6.40) are connected to inputs of the circuit under consideration. The difference of the node voltages or the current through the branch with added resistor can be considered as the output signal.

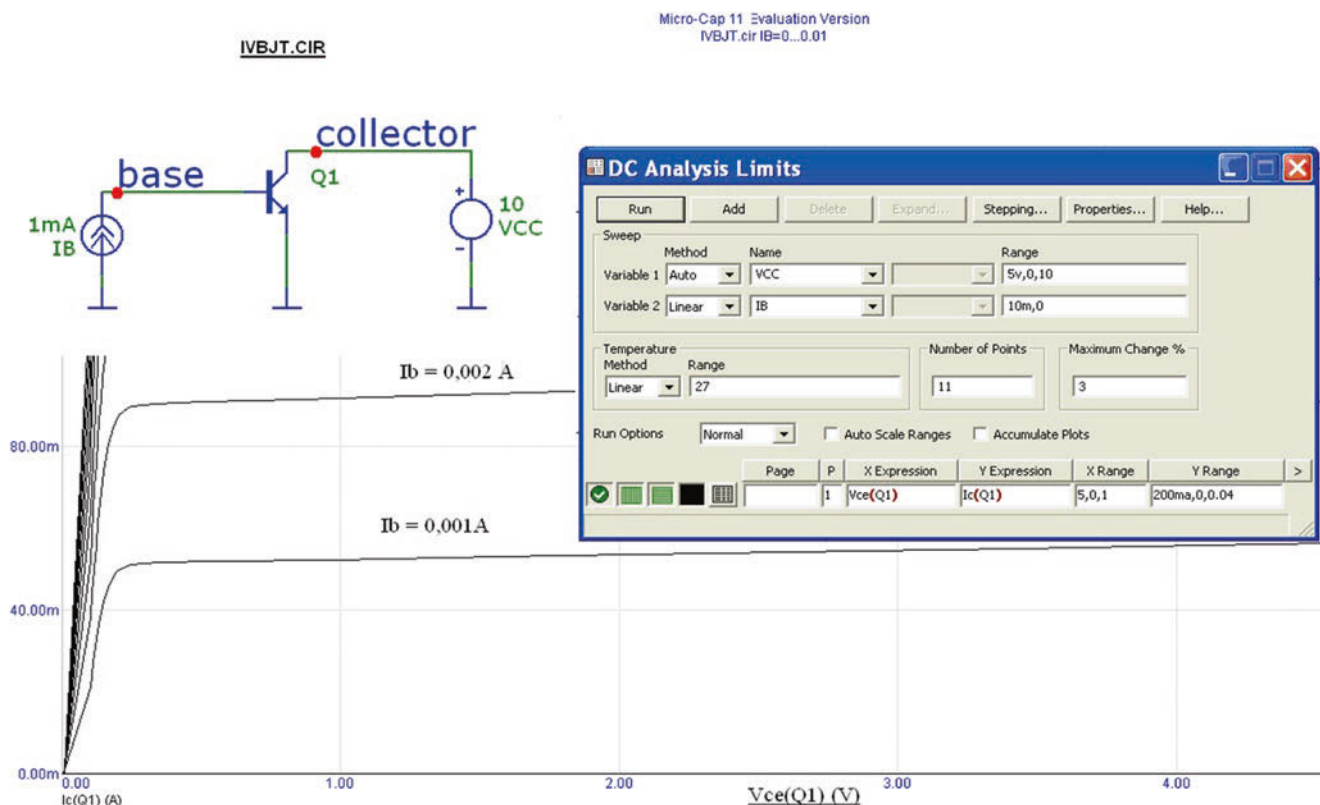


Fig. 6.40 The construction of the output characteristics' family

In the DC mode, the software short-circuits all inductors and eliminates all capacitors, and only after this, the DC mode is calculated at several values of the input impacts. Connection on the one DC source in the device input allows calculation of the transient in DC for devices, which do not use the reactive components, while two DC sources allow obtaining the family of output static characteristics.

After transfer into the DC mode, the software checks the correctness of electronic circuit formation, and if errors are absent, it forms the topological description and also provides the preparation to calculation of nonlinear equations by the Newton-Raphson iteration method. As the example of performing the analysis of the bipolar transistor model in DC, we consider the construction of the output characteristics' family with utilization of the circuit in Fig. 6.40.

Numerical parameters in the sub-menu **DC Analysis Limits** are given in several windows allocated by frames: *Sweep*.

In the Method column, the methods of variables variation are given:

Variable 1 with the name VCC (the power supply battery is designated) can change the value of EMF source with the help of the scroll bar:

Auto – the automatic step of measurement is chosen providing accuracy not more than required in % in the frame *Maximum Change%*.

Linear – the linear step given for parameter variation, which is defined in the column Range in the format $\langle \text{end} \rangle [\langle \text{start} \rangle [\langle \text{step} \rangle]]$. If there is no specified information about the *Step* parameter, its size will be set to $(\text{start} - \text{end})/50$.

Log – the logarithmic step specified in the field Range. By default, the initial value is assumed equal, $\langle \text{start} \rangle = \langle \text{end} \rangle / 10$, and the step value, $\text{step} = \exp(\ln(\text{end}/\text{start})/10)$.

List – in the form of the list of values divided by commas.

Name – the name of the first variable is indicated (the value of the DC voltage (or current) source; temperature; the value of one of components of the model described by the mathematical model). The name of variable being varied is selected from the list using the scroll bar. *Variable 2* – variation of the second variable obeys the same rules that are described for the first variable.

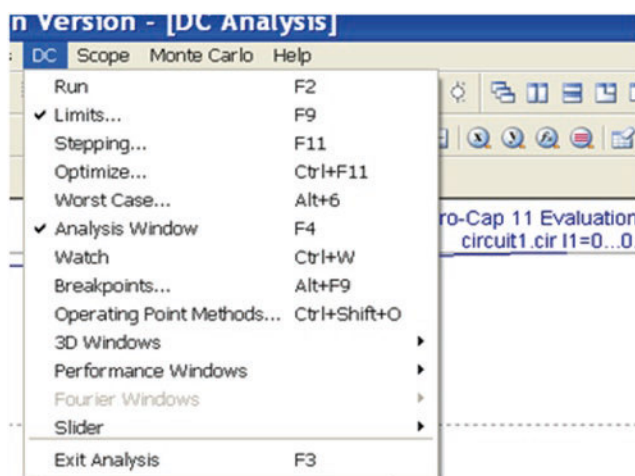
In the column *Method*, in contrast to Variable 1, the approach of automatic variation of the step is for Variable 2. In this case, for this variable, the approach of linear variation of the step or the option None is chosen, when only one variable changes.

Figure 6.40 shows an example of output characteristics of the bipolar transistor. To output blocks, the voltage source is connected, which changes with the automatic step. The current source I_b is connected to the base with the selected (*linear*) low of variation (Variable 2) from zero to 10 mA. The collector current (I_b) is selected as indicator displayed on the monitor screen, as a function of collector-emitter voltage $I_c = f(U_{ce})$. The value of the variable parameter (I_b) corresponding to each curve is determined in the result window by hovering the cursor over corresponding dependence (its value can be displayed using the T (*text*) function). The MC12 version provides a function for calculating the output characteristics of the transistor, but the analysis limits do not correspond to the real BT modes used in receivers.

Commands executed at temperature variation are similar as described in Sect. 6.4.2. This is concerned to description of representation conditions of circuit modeling in DC.

Being in the result window (after performing the calculation, for instance, of transfer functions), we can enter into the sub-menu DC in the menu line (Fig. 6.41) for editing the exit procedure for repeated launch of software.

Fig. 6.41 The list of commands expanding possibilities of the DC sub-menu

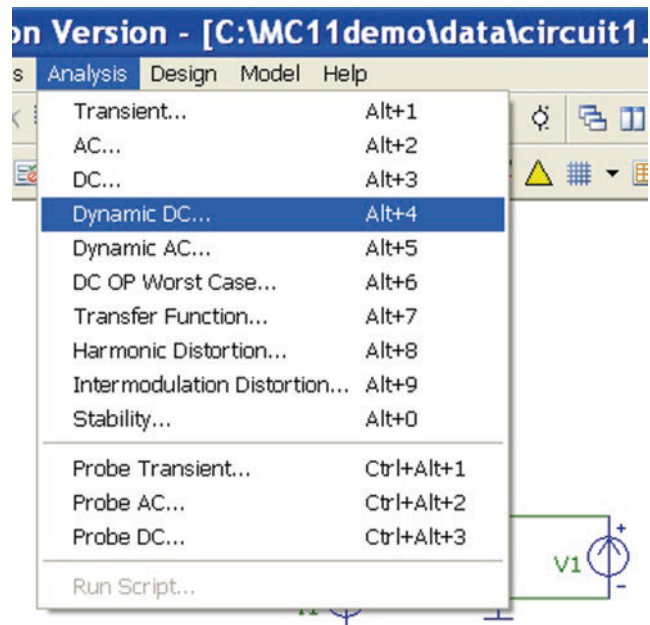


The list of commands in the sub-menu DC is the same for all types of analysis (on the right in each line of the menu, the "hot buttons" are indicated, which provide its execution).

6.4.4 Dynamic Mode of DC Analysis

Transfer in the dynamic range is performed by execution of command sequence: **Analysis** → **Dynamic DC** (Fig. 6.42).

Fig. 6.42 Transfer (by the command Dynamic DC) in the sub-menu of calculation and optimization of the active component mode in DC



A choice in the sub-menu **Analysis** of the line **Dynamic DC Limits** realizes the interactive process, at which for chosen in Fig. 6.42 variant of the task, on the example of the tuned amplifier circuit analysis, the software calculates DC voltages in the circuit nodes, currents in branches, etc. and results are represented in the circuit window.

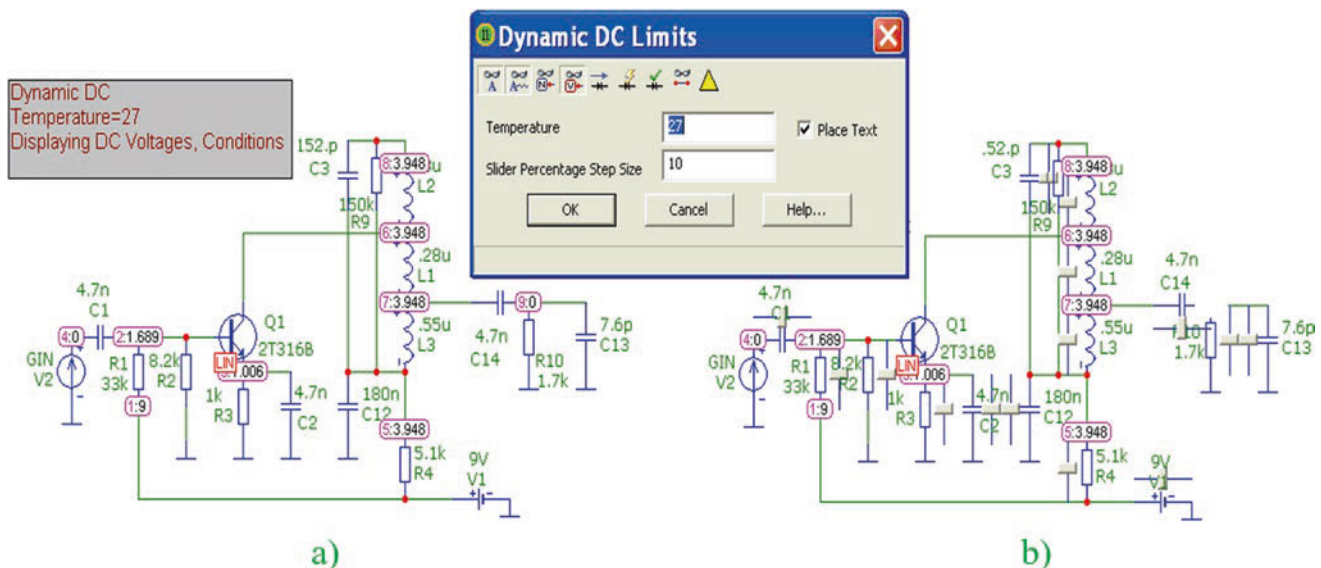


Fig. 6.43 An example of exit on the monitor screen of the calculated DC voltages in the nodes of the circuit (a) and possibilities of variation of component values (b) for changing the AE mode and optimization of the circuit components

After starting of **Dynamic DC** analysis mode, the software activates the node potential  representation button on the monitor screen in the circuit window by default. Under the line of the main components (Fig. 6.1), there are other buttons duplicated in the **Dynamic DC Limits** submenu (Fig. 6.43). First of all, these are *display buttons* that allow you to control the composition of information about the current DC mode displayed on the screen. The most commonly used of them is shown in Table 6.6.

Table 6.6 The most important indices representing in the circuit window

Pictogram	Executing function
	Display text messages on monitor screen: operating mode, temperature, information about measured parameters, etc.
	Show component designations in the circuit under analysis
	Execute a numeration of circuit nodes
	Show voltages in circuit nodes
	Show currents in circuit branches

Temperature – in the dialog window, the temperature value, at which the circuit analysis is executed.

Place Text – the settled mark informs about the exit to the monitor screen of text messages.

Slider Percentage Step Size – indicates on the percent variation of the chosen component value at each pressing of buttons
 ↑ (Up Arrow), ↓ (Down Arrow).

The command buttons:

OK – closing of the dialog window and transfer to analysis execution according to instructions in the sub-menu **Dynamic DC Limits**.

Cancel – closing of the dialog window and transfer to analysis execution without execution of entering variations.

Help – call the file of help.

Pressing on the **OK** button transfers the software in the calculation mode and results are presented in the circuit.

Make sure that transistors (e.g., Q1 in Fig. 6.43a) is in active mode and the pre-calculated operating point.

To achieve the values of current and voltages in transistor electrodes, which correspond to the required mode, we must change components of the circuit under investigation. It can be done by application of the variation mode of the chosen parameter (**Stepping**) or via the **Slider** sub-menu (Fig. 6.43b). A choice of varying parameters is provided with the help of the scroll bar.

6.4.5 Dynamic Small-Signal Analysis of Transfer Functions in Frequency Domain

The small-signal analysis mode is called by the command (Fig. 6.9) **Analysis → Dynamic AC** and allows investigation of circuit frequency properties at small level of the input impact for some frequencies and values of circuit parameters.

An interface for this mode is close to the **Dynamic DC** interface. The main difference consists in the fact that in the small-signal mode, the value of the *transfer coefficient* (module and phase) in voltage, which is the ratio of the voltage in the arbitrary node to the voltage in the input of the circuit under investigation, represents in nodes. This information may be supplemented by transfer coefficients in current and in power for different frequencies and temperature values at variations of components.

The transfer in the mode of the dynamic small-signal analysis is provided from the circuit window by the sequential choice: **Analysis → Dynamic AC**. This transfers the software in the mode of parameters specification in the dialog window **Dynamic AC Limits** (Fig. 6.44).

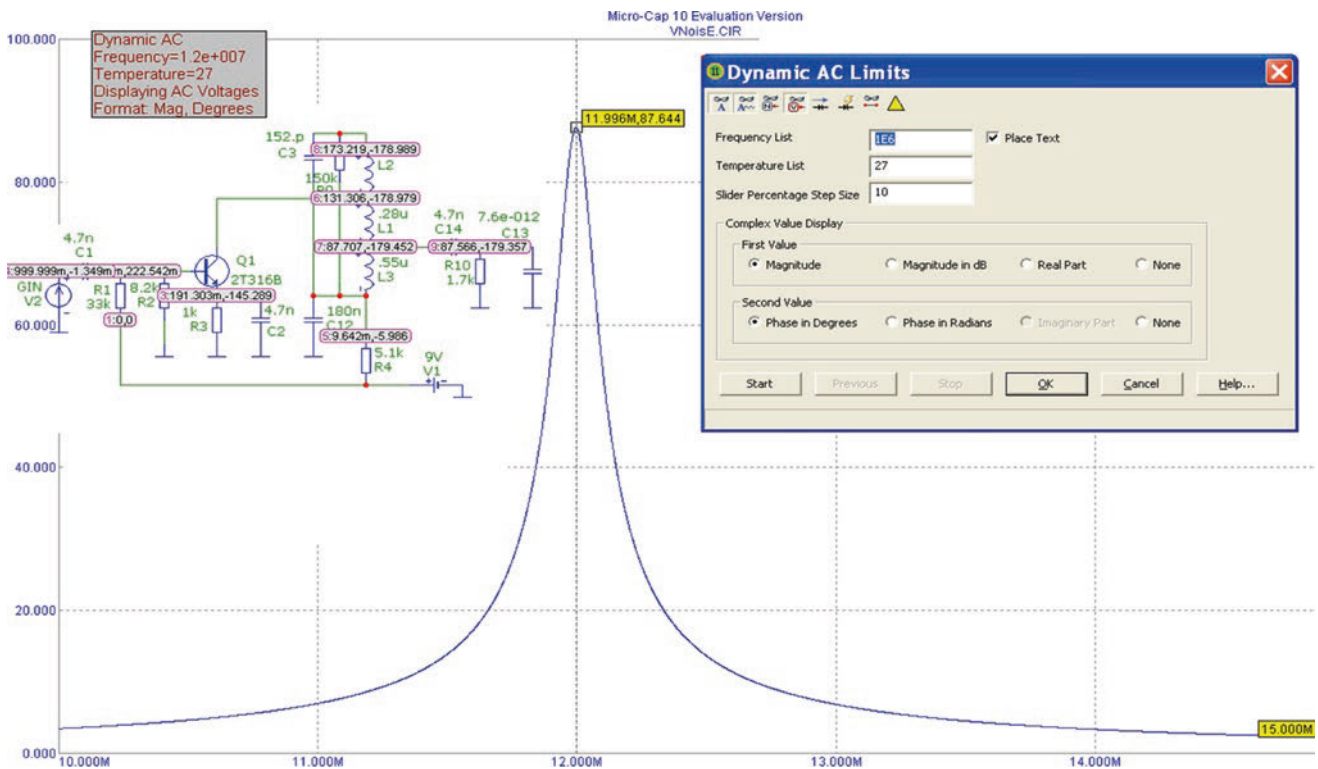


Fig. 6.44 The method of properties investigation of the tuned amplifier at small signals

In the bookmark **Dynamic AC Limits** (Fig. 6.44), buttons are located under the heading, which by name and purpose coincide with those presented in Table 6.6. In addition, commands appear:

Frequency List – the list of frequency values, in which the calculation of the small-signal transfer coefficient is executed.

Temperature List – temperature values.

Slider Percentage Step Size – variation of component value in percent.

The frame **Complex Value Display** integrates the auxiliary panels defining the method of presentation of calculated complex quantities in the circuit, where:

First Value – the first quantity presenting the complex number in the form:

Magnitude – the module of the complex quantity represented in exponent form $\text{Mag} = \sqrt{\text{Re}^2 + \text{Im}^2}$; *Real Part* – the real part of the complex quantity.

Second Value – the second exited part of the complex number:

Phase in Degrees – the phase (argument) of the complex quantity in degrees.

Phase in Radians – in radians.

Imaginary Part – the imaginary part of the complex quantity.

Start – the starting of the Dynamic AC analysis on the first frequency (according to the list) (in Fig. 6.44, 12 MHz). After starting, this button is switched off and the buttons Previous and Next are activated, which allow repeating the calculation on the previous in the list (Previous) or next (Next) frequency.

After closing the dialog window, functions of information control (exiting in the circuit) are transferred to the instrument panel.

6.4.6 Analysis of Intermodulation Distortions

Enter in the mode of intermodulation distortion analysis can be provided by sequential execution commands from the circuit window: **Analysis** → **Intermodulation Distortion Analysis Limits**.

The dialog window **Intermodulation Distortion Analysis Limits** (Fig. 6.45) contains (besides the command) the following buttons:

Run – the command for software transfer in the analysis mode.

Add – this command leads to appearance of the additional field (line) below the line, where a cursor will be placed.

Delete – the command allowing deleting of the line.

Expand – pressing on this button leads to execution of the field expansion for text editing or for its looking.

Stepping – pressing on this button leads to the sub-menu appearance (Fig. 6.46).

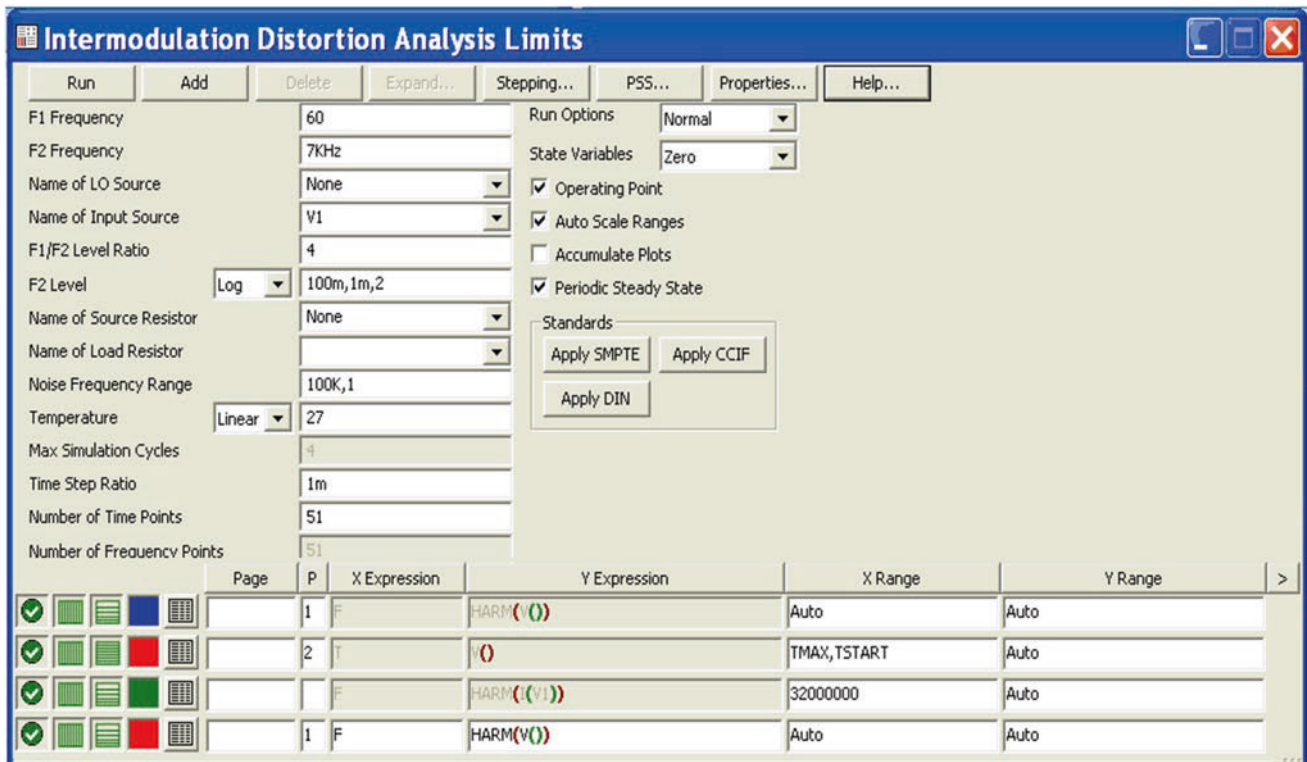


Fig. 6.45 The dialog window Intermodulation Distortion Analysis Limits

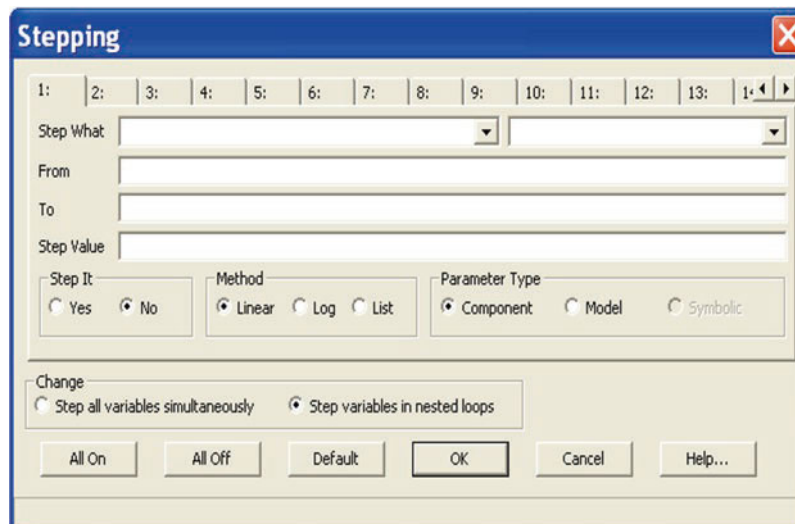


Fig. 6.46 Sub-menu for step-by-step component changes The *Stepping* submenu allows you to explore the properties of the model while changing the values of one or more components of the scheme. To do this, on one of the bookmarks of submenu, specify the limits and the method of setting the step of changes to the selected component of the scheme (From, To, Step value). You can simultaneously investigate the effect on the properties of the model of up to 14 components. If the component is complex (e.g., BT), then it is possible to vary one or more parameters of the transistor, stimulating a change in temperature in active region, aging, and disruption of the structure of semiconductors

Depending on the solved task, the different conditions of the stepping variation of the component variations can be chosen. Limits of component parameters' variation or AE models indicated in bookmarks of the sub-menu Stepping and also a choice of its measurement methods are described in Sect. 6.4.1.

PSS (Periodic Steady State) – pressing on the PSS button (Fig. 6.47) transfers the software operation in the mode of time interval calculation, in which the estimation of intermodulation distortions will be executed corresponding to the steady-state mode.

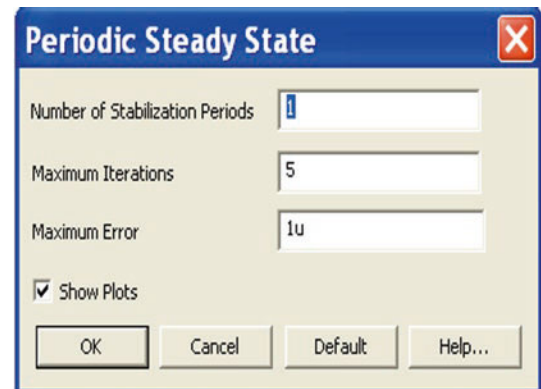
In the sub-menu, the number of input signal periods (*Number of Stabilization Periods*) is specified, before the software will be able to transfer to calculation of distortions in the time domain (nonlinear and intermodulation).

Periodic Steady State – this option uses the "pointing method" of transient elimination to provide the steady-state mode for all processes. The **PSS** option is the best variant, especially if circuit analysis having the small level of distortion is executed. This almost always provides the high accuracy of presenting curves and lesser calculation time.

Then the maximal number of iterations *Maximum Iteration* is specified.

Maximum error – the maximal value of calculation error in each iteration.

Fig. 6.47 The sub-menu describing conditions of the steady-state mode achievement



Properties – this command permits to transfer in the sub-menu **Properties for Intermodulation Distortion Analysis**, which defines parameters of intermodulation distortion analysis.

A mark in the line *Show Plots* allows exiting in the monitor screen the process at each iteration.

Help – at this command, the file help is called to specify the analysis limits in the mode **Transient Analysis Limits**.

The frequency range, in which the intermodulation distortion analysis is executed, is indicated in dialog windows of the sub-menu **Intermodulation Distortion Analysis Limits** (Fig. 6.45).

F1 Frequency – the value of the lower used frequency.

F2 Frequency – the value of the higher used frequency.

Name of LO Source – the name of the harmonic oscillation source, which is chosen with the help of the Component sub-menu in the command menu of the circuit window: Sine Source or Voltage Source. This is a necessary condition because at analysis of radio electronic circuits, the generator frequency can be added or subtracted from frequencies H1, IM2, and IM3. Frequency addition or subtraction is specified in the window Intermodulation Properties....

Name of Input Source – the name of input signal sources is specified. It must also be the harmonic voltage source of Sine Source or Voltage Source types. In order to execute the distortion analysis, at least, one of these sources connected to the circuit input must be in the circuit.

F1/F2 Level Ratio – the ratio of the generator voltage *amplitude* in low frequency (*F1*) to the voltage amplitude in higher frequency (*F2*).

F2 Level – the voltage amplitude of the signal source with higher frequency (*F2*). This may be one value or the list of amplitudes (usually used in Stepping mode).

Name of Source Resistor – the name of source resistor (internal resistance) in the circuit input. This resistance is used for calculation of the input power of the circuit under investigation ($PIN = (V/2) \cdot (V/2) / R_{SOURCE}$). If the option **None** is chosen, for input power estimation in the audio range, the $PIN = VIN \cdot IIN$ is usually used.

Name of Load Resistor – the name (title) of load resistor (load resistance) used for distortion measurement (for instance, RL, RLOAD, R23). The output power in the load is defined as $POUT = V \cdot V / R_{LOAD}$.

Noise Frequency Range – the noise frequency band; this is a band, above which the noise power is measured. Noises are calculated in the mode AC... at execution of the command `SQRT(SD(ONoise*ONoise))` by integration in the mentioned frequency band. We note that the noise contribution is estimated only when the **Include Noise** option is marked in the dialog window **Properties** from the sub-menu **Distortion**.

Temperature – the operation temperature chosen at modeling. It is usually used as the measuring parameter in Stepping mode.

Max Simulation Cycles is the multiplication factor of the time required to complete the entire process by the time spent on one simulating cycle..

Time Step Ratio – this value defines the maximal value of the time step admissible at analysis.

Number of Time Points – the number of time points exiting in the digital form at analysis of the time characteristics (Transient).

Number of Frequency Points – the number of digital values at frequency analysis (AC Analysis) used for noise calculation.

Run Options – variants of modeling modes:

Normal – usual mode when modeling executes without storage of modeling results on the hard disk.

Save – with storage.

Retrieve – with recovering.

State Variables – these may be zero values or created by the editor of steady states of variables (**State Variables Editor**).

Leave – permission; in this mode, exclusively current values of variables are stored.

Operating Point – the only modes in DC are calculated.

Auto Scale Ranges – this mode provides the automatic choice “AUTO” of the scaling along the X- and Y-axes at each following input in the modeling process.

Accumulate Plots – a choice of such a mode permits to accumulate (store) modeling results.

To show in the monitor screen of the two-port amplitude characteristic, for example, LNA, taking into account its nonlinear properties, being in the sub-menu Intermodulation Distortion, we must execute commands Intermodulation Distortion Windows → Add Intermodulation Distortion Windows... (Fig. 6.48).

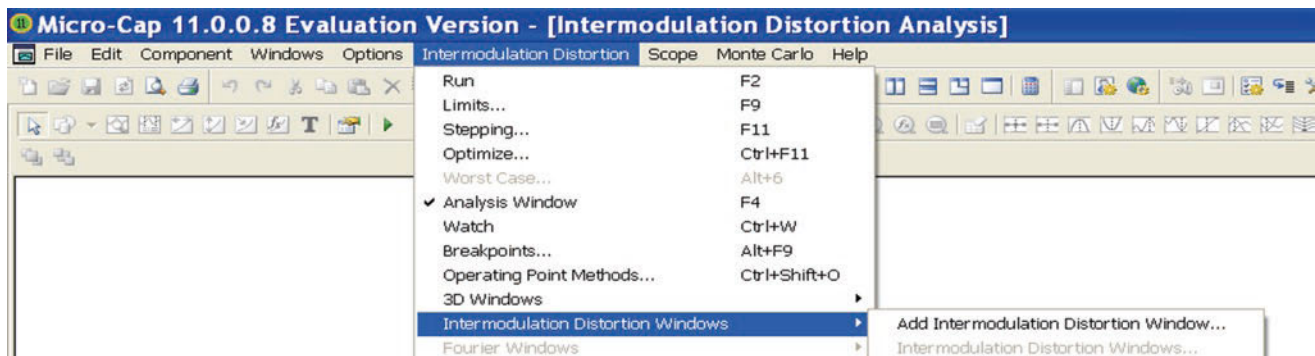
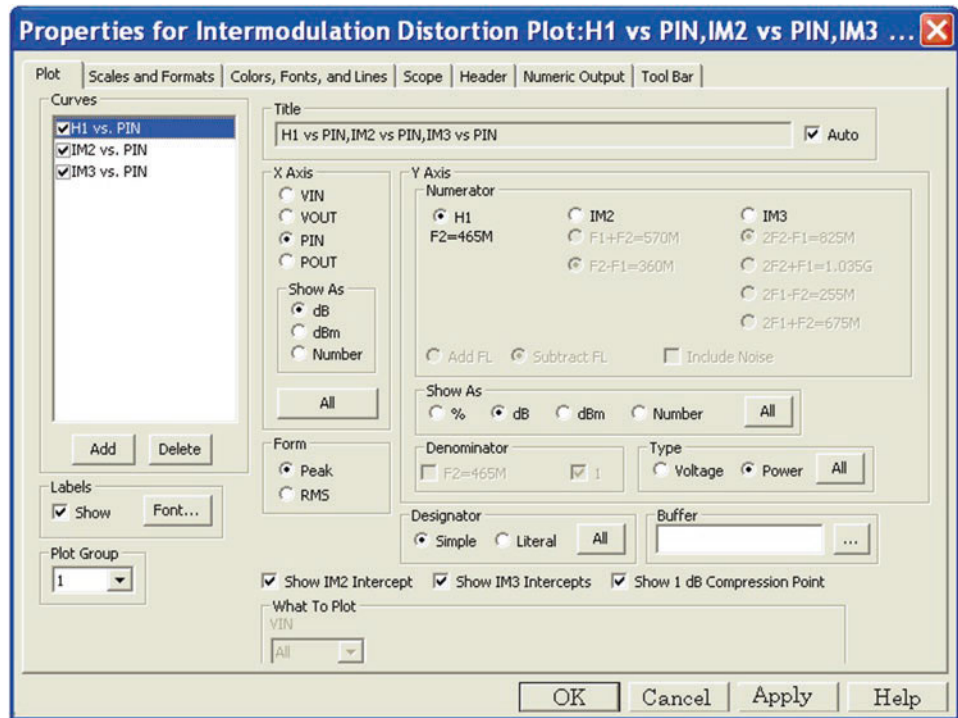


Fig. 6.48 The command sequence for exit in the screen of amplitude characteristic

Now we transfer to the sub-menu Properties for Intermodulation Distortion Plot: H1 vsPIN, IM2vsPIN, IM3... (Fig. 6.49), which provides the exit to the monitor of any of the mentioned dependences [64] or all together.

Fig. 6.49 The sub-menu of plots presenting on the screen



Confirmation of derivation of these functions (by OK) launches the modeling process.

This chapter describes the most common options for analyzing typical electronic circuits used in the RF front end of radio receiving devices implemented using modern AE. The capabilities of the professional version of MicroCap are much wider than those described in this chapter and allow you to investigate circuitry devices of a complex structure, covered by feedback, using nonlinear components (inductors, with magnetoelectric cores with hysteresis, losses in capacitors, mutual induction between circuit components, etc.). All this is beyond the scope of this book, but indicates the versatility of the MicroCap program, a friendly interface, and the use of modern component model libraries. Variants of its use are given in each of the chapters of the book offered to your attention.

Checking Questions and Numerical Examples

1. Denote the main differences between professional and student versions of the software MicroCap for electronic circuit modeling.
2. How can we execute the editing of the model inserted in the circuit window (to replace the components, to change its connection)?
3. List the action sequence at modeling execution in the frequency domain (AC) for several values of the component or a series of values of the AE model, for example, BF.
4. What is the principal difference of the Π -shaped model of the bipolar transistor from the mathematical Gummel-Poon model?
5. What are the features of the amplifier model realization on the self-closing and self-conducting nMOS transistor?
6. Draw a linearized Π -shaped model of the bipolar transistor that allows you to study its noise characteristics. Name the reasons for the appearance and properties of equivalent noise generators.
7. How can we take into account the non-ideality of reactive components of the oscillating circuit in the tuned amplifier?
8. Which command must we execute to obtain the signal in input/output of the tuned amplifier?
9. What are the principal differences in the modeling modes DC and Dynamic DC?
10. What are the differences in analysis conditions for nonlinear distortions in the software MC Harmonic Distortion and Intermodulation Distortion (results of modeling)?
11. Which variations must be included in the sub-menu Intermodulation Distortion Analysis Limits (Fig. 6.45) in order to calculate the intermodulation distortions for the mixer?
12. Insert variations in the sub-menu DC Analysis Limits (Fig. 6.40) in order to calculate the volt-ampere characteristic of the bipolar transistor at variation of the base current within the limits: $I_b = 20 \mu A \dots 1 \text{ mA}$.
13. How can we calculate in MicroCap medium the small-signal Y-parameters of the equivalent circuit of the bipolar transistor?

Appendix

Application 6.1 Calculation of small-signal Y-parameters of an active four-port network using the MicroCap program

The task of the designers of the radio receiving equipment (amplifiers, frequency converters), regardless of the degree of integration and the operating frequency range, is to achieve the required indicators, the most important of which are a given operating frequency band, gain, permissible nonuniformity of frequency response in the operating band, and nonlinear distortion coefficient.

Thus, the task of designing, for example, a resistor cascade on a BT, includes two stages: ensuring the operation of the AE in direct current and the main characteristics of the cascade.

1. Checking the AE mode for direct current.

The low level of input signals allows the use of A mode for AE. The choice of the BT mode in receivers is mainly determined by the place where the cascade is switched on in the RF front end and, for example, for its first stages it should provide $I_{c0} = 2 \text{ mA}$ at $U_{ce} = 2.5 \text{ V}$.

2. Calculation of small-signal parameters of BT to estimate the gain and other parameters of the stage.

If in the first case it is necessary to use a real circuit of stage, then to calculate the gain, it is necessary to know the small-signal parameters of the BT for the selected transistor manner to turn on. In the equivalent circuit of the two-port by the system of Y-parameters, coefficients of the equation system, defining its own and mutual conductance, are calculated at short circuit of input and output terminals in AC.

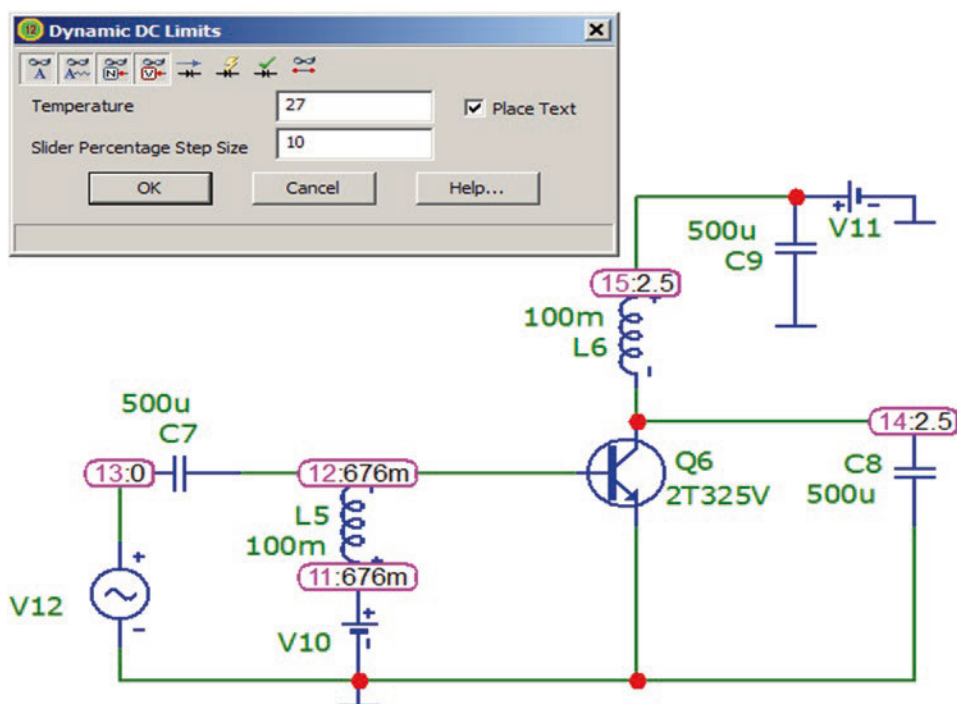
As an example, we will calculate the input conductivity of a BT small signal, switched on according to the scheme with OE, using the electrical circuit in the figure Appendix 6.1 in the MicroCap environment.

The “short circuit” mode for alternating current at the BT output is ensured by switching on a large capacitor C8. Internal resistance in power supply models (V11 and V10) is selected to be zero (ideal EMF sources). Inductors L5 and L6 exclude the influence of external circuits connected to the transistor on its performance, as well as the signal source (V12). BT properties are described by its PSPICE model:

```
.MODEL 2T325V NPN (BF = 321.5 BR = 1.78 CJC = 2.958p CJE = 3.42p IKF = 87.77 m IKR = 606.8 m IS = 9.164f
ISC = 100f ISE = 87.74f ITF = 300 m NC = 1.744 NE=1.473 RB=31 RC=0.2997 TF=122.2p TR=8.891n VAF=87
VAR=45 VTF=25 XTB=1.5 XTF=2).
```

After numbering the nodes and executing the commands Analysis → Dynamic DC Limits, select the V icon from the drop-down sub-menu, and display the voltage values in the circuit nodes on the monitor screen (Fig. 6.50).

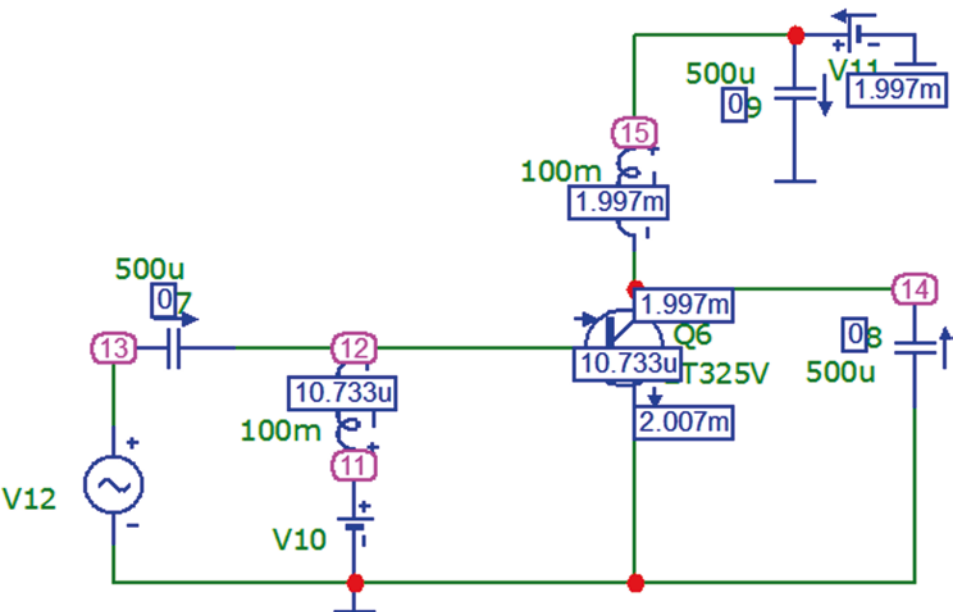
Fig. 6.50 Auxiliary circuit for calculating small-signal Y-parameters of BT (input conductivity, capacitance and direct transmission conductivity)



If they differ from the analysis conditions, it is necessary to correct the source voltage V11.

For a fixed value of $U_{ce} = 2.5 \text{ V}$, the current at the operating point is set by varying the voltage of the bias source (V10) connected to the BT base. For a source voltage of $V10 = 0.676 \text{ mV}$, the required collector current is provided: $I_{c0} = 2 \text{ mA}$ (Fig. 6.51).

Fig. 6.51 Distribution of direct current (DC) in the measuring circuit



To perform calculations of the input and conductivity of the direct transmission, it is necessary to execute in the circuit window: Analysis → AC... → Limits, go to the menu for setting the conditions of the circuit analysis: AC Analysis Limits (Fig. 6.52).

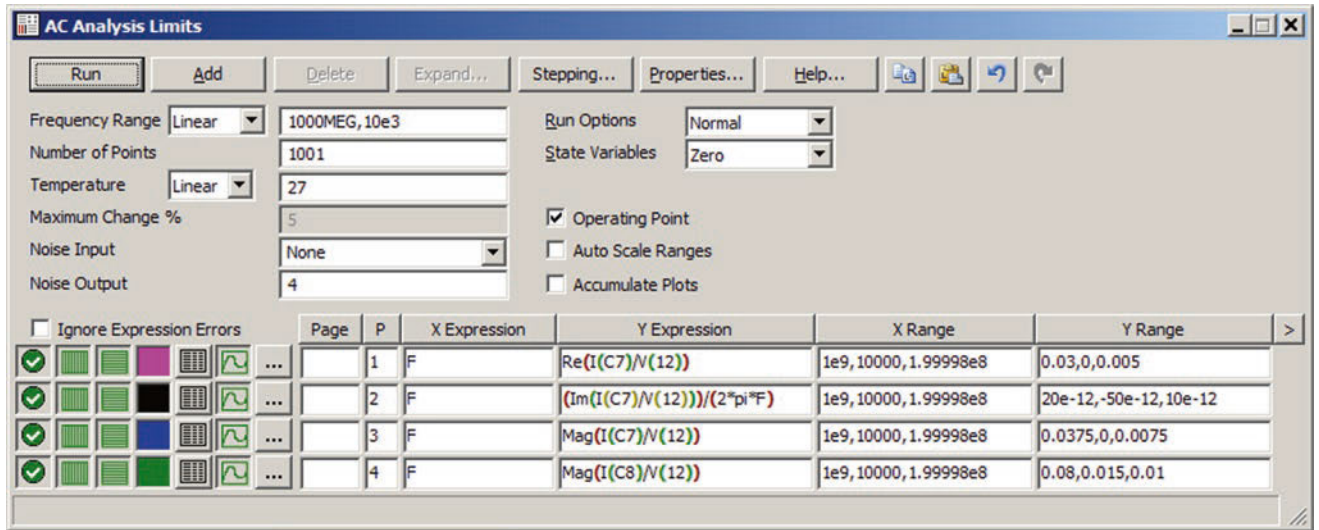


Fig. 6.52 Analysis conditions, methods, and limits of presentation of obtained results

As follows from the description of Dynamic AC limits submenu (Fig. 6.44), the output parameters of the simulation are the module and phase of voltages and current of the circuit components. This allows you to find the actual part of the conductivity of the corresponding ports of the circuit. Considering that the reactive component of the input (and output) conductivity of BT has a capacitive character, therefore, in the line (Fig. 6.53b) there is C11, its input capacitance. In the selected example, in the frequency range from 10 kHz to 1 GHz, the real part of the input conductivity $\text{Re}(Y_{11})$ BT (denoted as g_{11} in Fig. 6.53a), the input capacitance of the transistor (Fig. 6.53b), the input conductivity module $|Y_{11}|$ (Fig. 6.53c), and the direct transmission conductivity module $|Y_{21}|$ (Fig. 6.53d) are calculated.

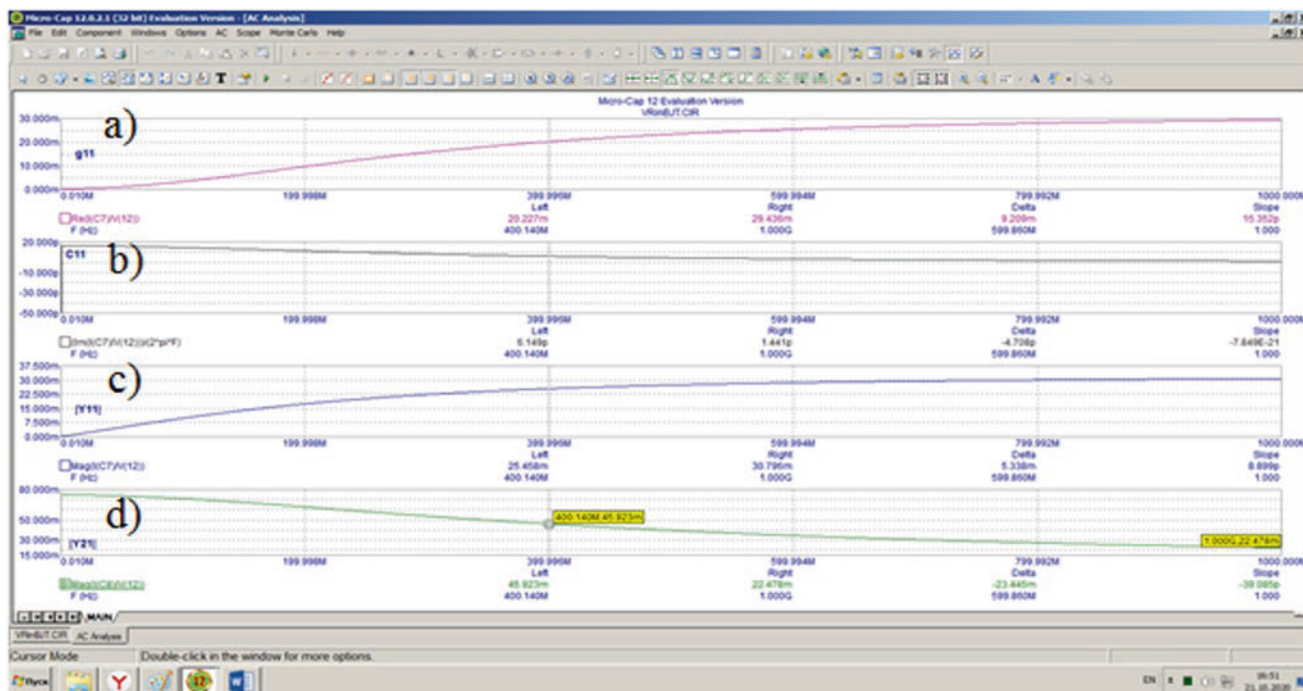


Fig. 6.53 Calculation of the components of the Y-parameter matrix of the BT. (a) the real part of the input conductivity, (b) the input capacitance, (c) the modulus of the input conductivity, (d) the modulus of the direct transmission

Calculation in the MicroCap environment of the input conductivity and input capacitance of the DC allows you to obtain values that are often used when manually calculating the parameters of the amplifying stages for the selected transistor DC mode and a specific operating frequency.

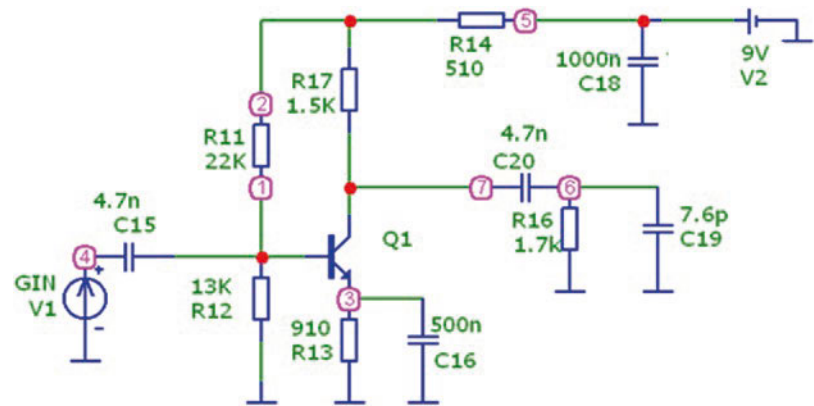
The value of any indices on the operation frequency is defined with the help of the cursor by extracting the interested frequency region in the sub-menu AC Analysis or by reduction of the region of analyzed parameters in columns Limits: XRange and YRange.

Application 6.2 Calculate the mean-square value of the noise voltage in input and output of the resistor stage (Fig. 6.54)

The electrical diagram of the stage includes the bipolar transistor of NPN type, parameters of which nonlinear PSPICE model includes:

.MODEL KT316D NPN (BF = 136.5 BR = 0.6577 CJC = 4.089p CJE = 1.16p IKF = 97.23 m IKR = 120 m EG = 1.11 FC=0.5 IS = 2.753f ISC = 15.5p ISE = 12.8p ITF = 151 m NE=2.496 RB = 70.6 RBM = 70.6 RC = 8.35 TF = 78.97p NC = 2 MJC = 0.33 MJE = 0.33 TR = 27.84n VAF = 96 VAR = 55 VJC = 650 m VJE = 690 m VTF = 25 XTB = 1.5 XTI = 3 XTF = 2).

Fig. 6.54 Schematic diagram of a resistor stage on BT



The mode of operation of the KT316D transistor, switched on according to the scheme with common emitter ($I_{c0} = 2 \text{ mA}$, $U_{ce0} = 2.7 \text{ V}$), must be provided by the previously calculated or selected value of components. At the first stage of modeling, the correspondence of the real position of the working point of the BT and the one obtained in the preliminary calculation is checked. To do this, in the schematics window, the following is performed sequentially: Analysis > Dynamic DC...

and on the Dynamic DC Limits submenu, icons are selected that display the currents in circuits and voltages in nodes. By clicking OK, we get the values of the collector current and the voltage calculated by the program. If the difference between the selected working point and the one obtained during modeling is more than 20%, the correction of the circuit components is performed. The total source V2 provides the power supply for collector circuit and with the help of a divider on R11 and R12 the bias voltage on the base of the bipolar transistor. The resistor R13 is the component of the negative feedback circuit in DC, which improves the stability of operation mode of the bipolar transistor and its gain.

In the circuit input, the generator of the harmonic oscillation V1 (GIN) connects, which is described by the model. MODEL GIN SIN ($A = 0.01$ F = 5MEG).

Noise indices of the amplifying stage take into account: thermal noises, flicker and shot noises. The circuit resistors, which play the auxiliary role, and the spurious resistors of the bipolar transistor are the sources of thermal noises.

Flickering noises are caused by processes occurring in bipolar transistors caused by defects in the crystal lattice in the semiconductor structure, causing random generation and recombination of charge carriers, the presence of traps in the base structure and

surface states. A distinctive feature of such noises is their spectral characteristic of the form $\sim 1/f^n$, where $n = 2$, when their intensity increases significantly near zero frequency. The noise of the shot usually reflects random changes in the injection and recombination current in semiconductors. The model of the source of such noise caused by the uneven flow of charge carriers through the collector-base junction is a current generator controlled by the base current, which, in turn, is determined by the degradation of the semiconductor structure.

The noise analysis is based on the measurement of the total contribution of all sources in the output of the circuit for the node (6), indicated in the line Noise output of the sub-menu AC Analysis Limits (Fig. 6.55). The spectral density of the mean-square value of the noise voltage (RMS, root mean square, dimension $\text{V}/\sqrt{\text{Hz}}$) is defined through the noise power spectral density in this node (V^2/Hz).

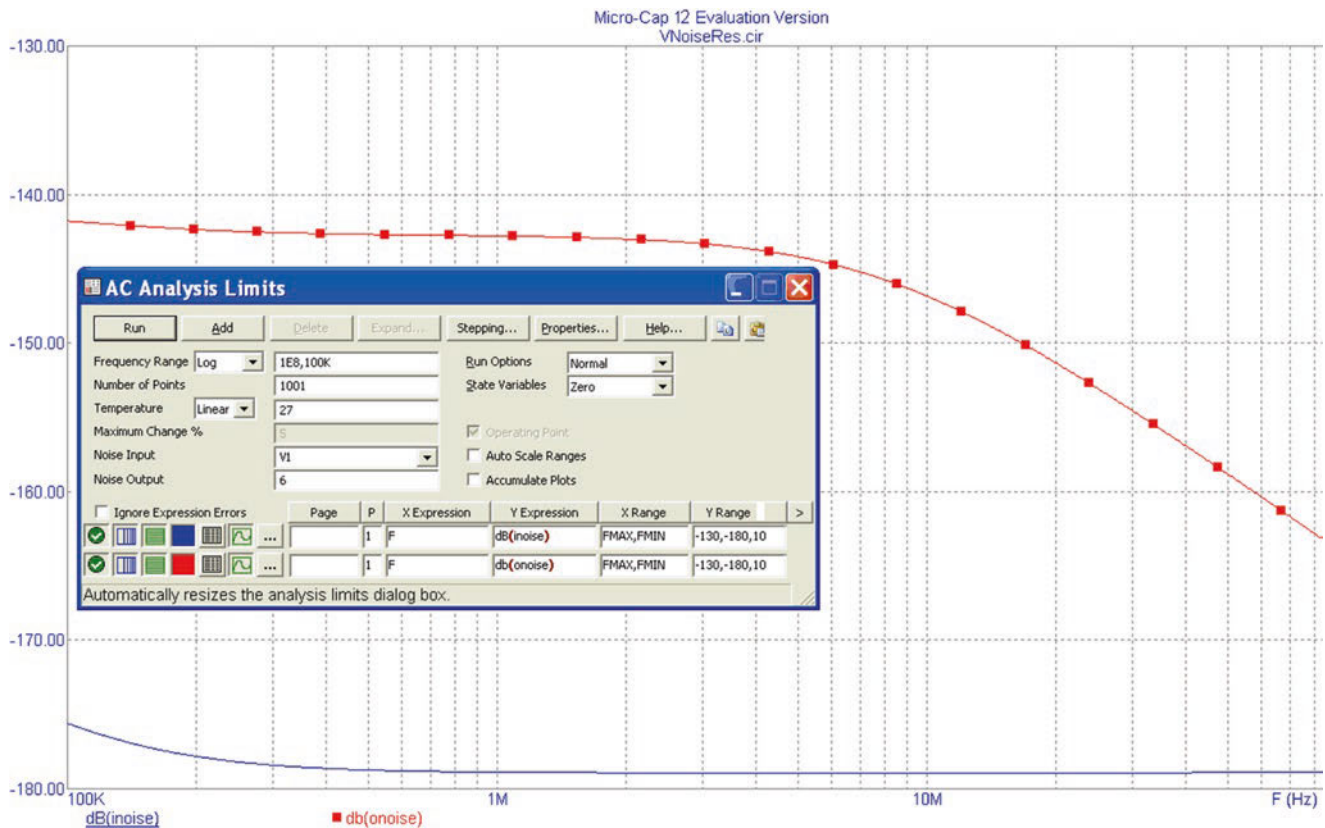


Fig. 6.55 Calculated noise characteristics at the amplifier input (solid line) and the total noise at its output (marked curve)

This plot presents the spectral density of the mean-square EMF in dB in the input (inoise, solid curve) and in output (onoise) of the circuit marked by points.

The input noise is calculated through the power value of the output noise by division on the frequency-dependent transfer function from the input (mentioned in the line Noise input, V1) to the output.

A feature of the calculation of noise indicators in the frequency domain is the use of a low-signal model of the circuit, which does not allow to simultaneously perform traditional analysis by the method of variable states when a harmonic signal of a much higher level is used as source.

Conclusion

The first book introduces the reader to the existing systems of fixed and mobile communications. Particular attention is paid to changes in the structure of wireless multi-station radio access networks and how it affected the architecture of radio receivers. The introduction of digital modulation techniques and optimal signal processing techniques on the receiving side has led to changes in the traditional receiver structures of analog modulation modes and the emergence of new ones based on the advantages of digital signal processing methods.

The classical structure of radio engineering systems has radically changed with the organization of communication between mobile users and the division of the service area into cells included in the terrestrial radio access network. This led to an increase in the role of the physical channel between the subscriber and the BS, which determines the performance indicators of the system, and the organization of packet data transmission over cable (copper and optical) and radio lines. Such a complex configuration of networks and their interaction made it possible to organize handover of subscribers within their own network and in the networks of other operators due to the use of compatible interfaces at various levels of interaction of open OSI systems.

All processes at the physical layer are controlled from the upper OSI layers, forming at the output of the BS transmitter of each cell a set of commands to MS characterizing the data transmission rate, modulation type, frequency band, clock signals, MS transmitter power control, etc. All this indicates the complexity signal processing algorithms in the receivers of each BS and MS, not taking into account the algorithms introduced in the system to increase the noise immunity of signals, antenna control, methods of coding, and signal detection. All this indicates a close relationship between the structure of radio access networks and the architecture of the receiver.

Open space, as a signal transmission guiding system, is affected by the external environment and highly dependent on the occupied bandwidth allocated to system, operating frequency, and signal propagation features. To reduce the influence of the external environment, the transmission of information about the state of the channel in the “up” direction is organized, which are included in the instructions for the MS, which determine the type of modulation and the data transmission rate for the transmitter.

A separate problem solved in the receiver is to build a non-tunable RF front-end when operating in the frequency ranges of systems of different standards and the occupied bandwidth of operating frequencies. This becomes especially relevant when mastering the frequency band allocated to the 5G standard. The RF front-end and its components must be designed to meet the stringent standards of each interoperating wireless communication system.

It is proposed to assess the properties of individual stages of the receiver and the RF front-end as a whole, with the possibility of analysis in the frequency domain and in the time interval in the MicroCap environment, which has a friendly interface and allows the use of conventional graphic designations adopted for the circuit diagrams of radio engineering devices.

The book describes the interconnection of systems, networks, and radio receivers as an integral part of such architectures. Possible ways of constructing receivers at the level of structural diagrams are shown, and influencing effects are described, starting with the properties of the radio channel and ending with the components used, as well as ways to assess their properties using circuit simulation.

The designs of the individual stages and possible ways of their implementation will be described in the second book “Elements of radio receivers of digital radio electronic systems” (in press).

Abbreviations

AAA	Authentication, authorization, accounting
AAC	Advanced audio coding
AAS	Adaptive antenna system
AC	Adjacent channel
ACF	Auto-correlation function
ACI	Adjacent channel interference, adaptive compensator of interferences
ACK/NACK	Acknowledge/not acknowledge
ADC	Analog to digital conversion
ADSL	Integrated digital subscriber line
AE	Active element
AES	Automatic error correction
AFC	Amplitude-frequency characteristic, automatic frequency control
AFR	Amplitude-frequency response
AGC	Automatic gain control
AGCN	Access grant channel
AIS	Analyzer of interference situation
AM	Amplitude modulation
AMPS	Advanced mobile personal system
AMR	Adaptive multi-rate
AMS	Adaptive MIMO switching
AP	Access point
APC	Automatic power control
ARN	Active remote node
ASK	Amplitude shift keying
ASM	Antenna switch module, analog signal multiplier
ASN	Access service network
ATM	Asynchronous transfer mode
ATS	Automatic telephone station
ATSC	Advanced television systems committee
AUC	Authentication center
AWGN	Additive white Gaussian noise
AWPM	Amplitude-width pulse modulation
BB	Base band
BBU	Base band unit
BCCH	Broadcast control channel
BER	Bit error rate
BLAST	Bell laboratories layered space-time
BM	Base model
BPF	Band pass filter
BPSK	Binary phase shift keying
BS	Base station
BSC	Base station controller
BSIC	Base station identity code

BSS	Basic service set, base station subsystem
BT	Bipolar transistor
BTS	Base transceiver station
BOC	Binary offset code
BQM	Balanced quadrature mixers
CA	Collision avoidance, carrier aggregation
CAT	Cable television
CC	Component carrier
CCCH	Common control channel
CCI	Co-channel interference
CCK	Complementary code keying
CCN	Cognitive control network
CCS	Control-correcting station
CD	Compact disk
CDMA	Code division multiple access
CDMA-DS	Code division multiple access direct spread
CDMA-FH	Code division multiple access frequency hopping
CDMA-TH	Code division multiple access time hopping
CEPT	European Conference of Postal and Telecommunication Administrations
CMN	Cognitive mesh network
CMOS	Complementary metal-oxide semiconductor
C-NMS	Cognitive network management system
CO	Central office
C-OFDM	Code OFDM
CoMP	Coordinated multi-point
CP	Central port
CPE	Customer premises equipment
CPC	Cognitive pilot channel
CR	Cognitive radio
C-RAN	Cloud radio access network
CRC	Cycling redundancy check, cyclic redundancy code
CSMA	Carrier-sense multiple access
CSN	Circuit switched network
CSP	Central system processor
CSPDN	Circuit-switched public data network
CSS	Customer switching system
CT	Cordless telephony
CTA	Cordless terminal adapter
CTM	Cordless terminal mobility
CTV	Color TV
CWN	Cognitive wireless network, composite wireless network
DAA	Data access arrangement
DAB	Digital audio broadcasting
DAMA	Demand assigned multiply access
DAMPS	Digital advanced mobile phone system
DBPSK	Differential double-position phase shift keying
DC	Directional coupler, direct current, data center
DCI	Downlink control information
DCMS	Differential correction and monitoring system
DCS	Digital cellular system
DDS	Direct digital frequency synthesizer
Det	Detector
DECT	Digital enhanced cordless telecommunication
DFS	Dynamic frequency selection
DRiVE	Dynamic radio for IP services in vehicular environments

DRM	Digital radio mondiale
D-RAN	Distributed RAN
DRSF	Double-circuit pass-band filter
DQPSK	Differential quadrature phase shift keying
DSP	Dynamic spectrum allocation
DSMA	Digital sense multiple access
DS-CDMA	Direct spectrum CDMA
DSP	Digital signal processor
DSSS	Direct sequent spread spectrum
DSL	Digital subscriber line
DSS-UWB	Direct spread spectrum UWB
DTV	Digital television
DVB	Digital video broadcasting
DVB-T	Digital video broadcasting terrestrial
DWDM	Dense wavelength-division multiplexing
E-UTRAN	Enhanced UMTS terrestrial radio access network
ECC	Electronic communications committee (of Europe)
RMF	Electromotive force
EDACS	Enhanced digital access system
EDGE	Enhanced data rate for global evolution
EGSM	Extended Global System for Mobile Communications),
EIR	Equipment identification register
eMBB	Enhanced Mobile Broadband
EMC	Electromagnetic compatibility
EMF	Electromotive force
EMI	Electromagnetic interference
EPC	Electronic product code
eNB	Evolution node B, base station
EPIC	Electrical-physical information converter
ET	Ephemerid time
ETSI	European Telecommunications Standards Institute
FBC	Feedback circuit
FBMC	Filter bank multi carrier
FC	Frequency converter
FCC	Federal Commission of Communication
FCCH	Frequency correction channel
FCD	Frequency channel division
FDD	Frequency division duplex, full function device
FDM	Frequency division multiple
FDMA	Frequency division multiple access
FEC	Forward error correction
FEM	Front-end module
FET	Field-effect transistor
FFD	Full function device
FFT	Fast Fourier transform
FIC	Fast information channel
FH	Frequency hopping
FHDC	Frequency hopping diversity coding
FHSS	Frequency hopping spread spectrum
FHSS-UWB	Frequency hopping spread spectrum UWB
FI	Fluctuation interference
FIC	Fast information channel
FIFO	First in, first out
FLS	Filter of lumped selectivity
FM	Frequency modulation

F-OFDM	Filtered orthogonal division multiplexing
FS	Frequency synthesizer
FSA	Field spectrum allocation
FSK	Frequency shift keying
FSS	Federal communication commission (of USA)
FUSC	Fully used sub-carrier
GBCMC	Ground-based complex for monitoring and control
GBS	Ground-based station
GERAN	GSM EDGE radio access network
GFDM	Generalized frequency division multiplexing
GMSC	Gateway mobile switching center
GMSK	Gaussian minimum shift keying
GNSS	Global navigational satellite system
GoS	Grade of service
GPIO	General purpose input/output
GPRS	General packet radio service
GPON	Gigabit passive optical network
CP	Compression point, cyclic prefix
CR	Cognitive radio
GSO	Geostationary orbit
GTP	GPRS tunneling protocol
HARQ	Hybrid automatic repeat request
HBT	Heterojunction bipolar transistor
HD	High definition
HDTV	High-definition television
HDR	High data rate
HEMT	High emitted mobility transit
HF	High frequency
HI	HARQ indicator
HIPERLAN	High performance local area network
HIPERMAN	High performance metropolitan area network
HLR	Home location register
HoA	Home address
HSAGC	High-speed AGC
HSCSD	High-speed circuit switched data
HSDPA	High-speed downlink packet access
HSPA	High-speed packet access
HSUPA	High-speed uplink packet access
I	Inphase
IASA	Inter access spectrum anchor
IBOC	In band on channel
IC	Integrated circuit, input circuit
ICI	Inter-channel interferences
iDEN	Interactive data entry network, integrated digital enhanced network
IGSL	Integrated digital subscriber line
IEEE	Institute of Electrical and Electronics Engineers
IEC	International Electrotechnical Commission
IF	Intermediate frequency
IFA	Intermediate frequency amplifier
IMA2	Intermodulations-apart-product 2
IMD	Intermodulation distortion
IMP2	Intermodulations-product 2
IMP3	Intermodulations-product 3
IMS	IP multimedia subsystem
IMSI	International mobile subscriber identity

IMT/UMTS	International mobile telecommunication/universal mobile telecommunications system
IN	Input circuit
IOS	International organization on standardization
IP	Internet protocol
IP2	Intercept point on 2 harmonic
IP3	Intercept point on 3 harmonic
IPD	Incumbent profile detection
IPv4	Internet protocol version 4
iPSDN	Integrated packet switched digital network
IRI	Industrial radio interference
IRFM	Input RF module
ISDN	Integrated switched digital network
ISM	Industry science medicine (range)
ISO	Open system interconnection
ITU	International Telecommunications Union
JRRM	Joint radio resource management
LAI	Location area identity
LAN	Local area network
LDSS	Local differential sub-system
LFM	Linear frequency modulation
LNA	Low-noise amplifier
LO	Local oscillator, low-orbit
LSI	Large-scale IC
LTE	Long-term evolution
LTE/E-UTRAN	Long-term evolution/evolved-UMTS-terrestrial radio access network
MAC	Media access control
MAHO	Mobile-assisted handover
MAI	Multi-access interference
MAN	Metropolitan area network
MB-OFDM	Multiband OFDM
MBS	Multicast and broadcast service
MBWA	Mobile broadband wireless access
MC	Multicarrier
MCS	Mobile communication system
MDHO	Macro diversity handover
MESFET	Metalized semiconductor field-effect transistor)
METIS	Mobile and wireless communications enablers for the twenty-twenty information society
MF	Matched filter
MFA	Amplifier of the modulation frequency
MI	Multiplicative interference
MIMO-OFDM multi-channel input	multi-channel output
MISO	Multiple input single output
MLH	Maximal load hour
MME	Mobility management entity
MMSE	Minimum mean squared error
mMTC	Massive machine type communications
MO	Middle orbit
MPEG	Motion pictures experts group
M-PSK	Multi-phase shift keying
MP2MP	Multipoint-to-multipoint
MS	Mobile station
MSC	Main service channel, mobile switching center
M2M	Machine to machine

MTC	Machine type communications
MUE	Multiuser equipment
MU MIMO	Multiuser MIMO
N	Noise factor
NAP	Network access provider
NCO	Numerically controlled oscillator
NE	Nonlinear element
NF	Noise factor, dB
NGN	Next-generation network
NG-PON2	New generation of passive optical network
NLS	Noise-like signal
NMS	Network management system
NMT	Nordic mobile telecommunication
NRZ	Not return to zero
NSA	Navigational space apparatus
NSP	Network service provider
NSS	Noise shape signal
NVF	Network function virtualization
OATS	Office automatic telephone station
OFDM	Orthogonal frequency division multiplexing
OFDMA	Orthogonal frequency division multiple access
OLT	Optical line terminal
OMC	Operating and maintenance center
ONU	Optical network units
OOB	Of-of-band
OOK	On-off keying
OS	Open service
OSI	Open system interconnection
OSM	Operator spectrum management
PaLFI	Passive low-frequency interface
PAA	Phased antenna array
PAC	Perceptual audio coding
PAN	Personal area network
PBCC	Packet binary convolution coding
PCH	Paging channel
PCM	Pulse-code modulation
PCS	Personal communication system, personal communication services
PDC	Personal digital cellular
PDN	Public data network
PDP	Packet data protocol
RFA	Radio-frequency amplifier
PFC	Phase-frequency characteristic
PG	Processing gain
PHS	Personal handyphone system
PHY	Physical layer protocol
PI	Pulse interference
PLL	Phase locked loop
PMP	Point-to-multipoint
PNF	Physical network function
PNI	Pseudo-noise in-phase
PON	Passive optical network
PPM	Pulse position modulation
PNQ	Pseudo-noise quadrature
PRS	Pseudo-random sequence, public regulated service
PRCS	Personal radio call systems

RPE-LTP	Regular pulse excitation long-term prediction
PRMA	Packet reservation multiply access
PRP	Pseudo-random process
PRS	Pseudo-random sequence, public regulated service
PRS	Public regulated service
PSTN	Public switched telephone network
PT	Potable termination
PTN	Public telephone network
PUSC	Partially used sub-carrier,
Q	Quadrature
QC	Quadratic couplers
QPSK	Quadrature phase shift keying
QoS	Quality of service
RACE	Research in advanced communications equipment
RACH	Random access channel
RAM	Radio access method, random access memory
RAN	Radio access network
RBER	Residual bit error ratio
RDSS	Regional differential sub-system
RFA	Radio-frequency amplifier
RF front-end	Radio-frequency front-end
RFID	Radio-frequency identification
RFN	Reduced frame number
RLL	Radio local loop
RNS	Radio navigation system
RM	Reversible modulator
RO	Reference oscillator
ROADM	Reconfigurable optical add/drop multiplexer
ROG	Reference oscillation generator
RR	Radio receiver
RRL	Radio relay line
RRNP	Reference radio navigation point
RRS	Radio relay system, reconfigurable radio system, remote radio heads
RS	Relay station
RSSI	Receive signal strength indicator
SAE	System architecture evolution
SB	Single band
SBR	Spectral band replication
SC	Single carrier
SC-FDMA	Single carrier frequency division multiple access
SCH	Synchronization channel
SD	Subtraction device
SDM	Space division multiplexing
SDMA	Space division multiplex access
SDN	Software defined networking
SDR	Soft defined radio
SE	Subscriber equipment
SF	Spreading factor
SIM	Subscriber identity module
SISO	Single input single output
SMS	Short message service
SMR	Specialized mobile radio (system)
SNR	Signal noise ratio
SoC	System-on-chip
SOC	Single oscillating circuit

S-OFDMA	Scaling-OFDMA
SPDT	Single pole double throw
SPI	System programming interface
SPRC	System of the personal radio call
SR	Soft radio
ST	Subscriber terminal
STBC	Space timing block code
STC	Space-time coding
SU MIMO	Single user MIMO
SV	State vector
TCP	Transmission control protocol
TDD	Time division duplex
TDMA	Time division multiple access
TD-SCDMA	Time division-synchronous CDMA
TH-UWB	Time hopping UWB
3G1X EV-DO	Enhanced version data only
TLL	Time of the largest load
TETRA	Terrestrial trunked radio
TH-SS	Time hopping spread spectrum
TMSI	Temporal identifier of the mobile station
TPC	Transmit power control
UCN	Unified communication network
UE	User equipment
UFMC	Universal filtered multicarrier
UHD	Ultra-high definition
ULP	Ultra-low power
UMB	Ultra-mobile broadband
UMTS	Universal mobile telecommunications system
UPE	User plane entity
URLLC	Ultra-reliable and low latency communications
UTRA	UMTS terrestrial radio access
UTRA-FDD	UMTS terrestrial radio access, frequency division duplex
UTRA-TDD	UMTS terrestrial radio access, time division duplex
UTRAN	UMTS terrestrial radio access network
UWB	Ultra-wideband
VAC	Volt-ampere characteristic
VC	Virtual cell
VCN	Virtual cellular network
VCCO	Voltage controlled crystal oscillator
VCO	Voltage controlled oscillator
VHE	Virtual home environment
VHF	Very high frequency
VLR	Visitor location register
VOFDM	Vector OFDM
VoIP	Voice over IP
VoLTE	Voice over LTE
VPN	Virtual private network
VSAT	Very small aperture terminal
WBS	Wideband signals
WCDMA	Wideband CDMA
WCDMA/UMTS	Universal mobile telecommunication services in WCDMA
WDM	Wavelength-division multiplexing
WDSS	Wide-zone differential sub-system wide-zone differential sub-systems
WiFi	Wireless fidelity
WiFiG	Wireless fidelity gigabit

WiMAX	Worldwide interoperability for microwave access
WLAN	Wideband local-area network
WLD	Wideband-limiter-decision
WLL	Wireless local loop
WLN	Wideband-limiter-narrowband
WMAN	Wireless local area network
WP	Wireless port
WPAN	Wireless personal area network
WRS	Wireless relay station
WRZ	Without the return to zero
WSD	White space devices
WSN	Wireless sensor network
WWAN	Wireless world area network
ZigBee	Zigzag bee
ZIF	Zero intermediate frequency

References

1. K. Schwab The Fourth Industrial Revolution. World Economic Forum, 2016.
2. ITU-T Rec. Y.2060 (06/2012): Overview of the Internet of Things.
3. S.R. Saunders. Antennas and Propagations for Wireless Communication Systems, John Wiley& Sons, Ltd., Chichester, 1999.
4. W.C.Y. Lee. Mobile Cellular Telecommunication, Analog and Digital Systems, 2nd Edition, McGraw-Hill, New York, 1995.
5. A. Mehrota, GSM System Engineering, Artech House Publishers, Boston, 1997.
6. G. Heine GSM Networks: Protocols, Terminology and Implementation, Artech House, Boston, London, 1999.
7. N.J. Booucher The Trunked Radio and Enhanced PRM Radio Handbook, John Wiley & Sons, Inc., New York, 2000.
8. Draft ETS 3000 392-2, Radio Equipment and Systems (RES); Trans-European Trunked Radio (TETRA); Voice plus Data (V+D); Part 2: Air Interface (AI), Second Edition, December 1999.
9. H. Holma, A. Toskala WCDMA for UMTS. Radio Access for Third Generation Mobile Communications. Third Edition – John Wiley & Sons, Inc. 2004.
10. B. Martin UMTS par Satellite: Adaptation du Standard 3GPP UTRA FDD W-CDMA. – Alcatel Space. 2006.
11. Analysis of Potential Interface from WiFi to 3G UTRA-FDD Femto BS. – ECC PTI(09)007. February. 2009.
12. 3GPP TR 25.913. Requirements for Evolved UTRA (E-UTRA) and Evolved UTRAN (E-UTRAN (Release 9).
13. 3GPP TS 36.311 Evolved Universal Terrestrial Radio Access (E-UTRA); Physical Channels and modulation (Release 9).
14. F.H. Harmuth Nonsinusoidal waves for Radar and Radio Communications. Academic Press. N-Y. 1991. – 474 p.
15. F.H. Harmuth Squelch Theory. N-Y. 1977. – 574 p.
16. C.A. Shannon Mathematical Theory of Communication // Bell System Technical Journal. Vol.27, 1948, pp. 379–423 and pp. 623–656.
17. H. Zhou The Internet of Things in the Cloud: A Middleware Perspective. – CRC Press, 2012. – 2012. 365 p.
18. W.C.Y. Lee Mobile Cellular Communication Systems. – McGraw-Hill. New York, 1989
19. A roadmap for RFID Applications and Technologies/ Final Report. CE RFID, 2008. – 343 p.
20. K. Feher Wireless Digital Communications. Modulation and Speed Spectrum Applications. – Prentice_Hall PTR, 1995. – 400 p.
21. Proakis J.D. Digital Communications. – McGraw-Hill. Englewood Cliffs, 1995.
22. A. Abdulla Characterization of MIMO Chanel Capacity in Urban Microcellular Environment. Technical Report T55. Helsinki University of Technology Communications Laboratory. 2007.
23. 3GPP TS 36.101. v8.7.9 (2009-09). Technical Specification Group Radio Access Network; Evolved Universal Terrestrial Radio Access (E-UTRA); User Equipment (UE) radio transmission and reception. Technical Specification (Release 8). 2009.
24. R. Becker LTE and Evolution to 4G. Wireless Design and Measurement Challenges. –Agilent Book. June 2009.
25. B. Razavi Design of Analog CMOS integrated Circuits, McGraw Hill, 2000.
26. S.W Wedge, R. Compton, D. Rutledge Puff. Computer Aided Design for Microwave Integrated Circuits. Version 2.0. – Passadena: Caltech, 1991. – 61 p.
27. P. Antognetti, G. Massobrio Semiconductor device modeling with SPICE. McGraw-Hill, Inc. – New York, 1988. – 391 p.
28. Micro-Cap 8.0 Electronic Circuit Analysis Program Reference Manual Copyright 1982–2005 by Spectrum Software 1021 South Wolfe Road Sunnyvale CA 94086 Internet <http://www.spectrum-soft.com>
29. http://ru.wikipedia.org/wiki/IEEE_802.15.4
30. ETS 300 175 – 1; Radio Equipment and Systems (RES); Digital European Cordless Telecommunications (DECT) Common Interface Part 1: Overview.
31. 3GPP TS 37.104 V9.4.0 3rd Generation Partnership Project; Technical Specification Group Radio Access Network; Evolved Universal Terrestrial Radio Access (E-UTRA); User Equipment (EU) radio transmission and reception (Release 9), 2010-06.
32. 3GPP Long Term Evolution: System Overview, Product Development and Test Challenges. – Agilent Technologies.
33. GP2010. GPS Receiver RF Front End. /Supersedes edition in August 1996 Global Positioning Products Handbook, HB4305-1.0 DS4056-3.4. October 1996.
34. C. Oestges, B. Clerckx MIMO Communications: From Real-world Propagation to Spice –time Code Design. – Academic, 2007.
35. U. Tietze, K. Schenk, E. Gamm Halbleiter-Schaltungs Technik, 15 Auflage. Springer-Verlag, Berlin, Heidelberg, 2016.
36. H. Chen, M. Guizani Next Generation Wireless Systems and Networks. – John Wiley & Sons Ltd. 2006
37. Report ITU-R SM.2152. Definition of Software Defined Radio (SDR) and Cognitive Radio System (CRS). 2009.

38. Recommendation ITU-R M.1641–1. A methodology for co-channel evaluation to determine separation form a system using high-altitude platform stations to cellular system to provide IMT-2000 service. 2003–2006.
39. ESTI TR 102 802 v0.1/2/ (2009–12). Reconfigurable Radio Systems (RRS); Cognitive Radio System Concept. – Draft Technical Report. 2009.
40. Draft ECC Report 159. Technical and Operational Requirements for Possible Operation of Cognitive Radio Systems in the “White Space” of the Frequency Band 470–790 MHz. 2010.
41. 3GPP TS 36.311 Evolved Universal Terrestrial Radio Access (E-UTRA); Physical channels and modulation (Release 9).
42. S. Clint, C. Daniel. 3G Wireless Networks McGraw-Hill Companies, Ltd., 2002.
43. Bluetooth Specification. Version 1.1. Part A. Radio Specification.
44. http://ru.wikipedia.org/wiki/IEEE_802.15.4
45. ITU-R Document 6/282-E. Working Party 6A. Draft new Recommendation ITU-R BS.[MULT-FM]. Requirements for enhanced multimedia services for digital terrestrial broadcasting in VHF Bands I and II. 26 October 2010.– 2p.
46. The Global Positioning System. Interface Control Document ICD-GPS-200-C (v. IRINC200C 2004) <http://www.arinc.com/downloads/is-gps-200mar06.pdf>
47. 3GPP TS 36.201 v 8.2.0 (2008–12) LTE Physical Layer – General description
48. http://www.3gpp.org/ftp/workshop/2009-12-17_ITU-R_IMT_Adveval/docs/pdf/REV-090006.pdf
49. Juniper Networks White Paper, Machine-to-Machine (M2M) – The Rise of the Machines, 2011, [Online Available: <http://www.juniper.net/us/en/local/pdf/whitepapers/2000416-en.pdf>]
50. F. Schaich, T. Wild “UFMC OFDM Transceiver Chain”, <http://www.5gnow.eu/node/63>
51. N. Michailow, M. Matth  , I.S. Gaspar, A.N. Caldeilla, L.L. Mendes, A. Festag, G. Fettweis Generalized frequency division multiplexing for 5th generation cellular networks. Transactions on Communications, 2014
52. GS1 «EPC™ Radio-Frequency Identity Protocols Generation-2 UHF RFID Specification for RFID Air Interface Protocol for Communications at 860 MHz – 960 MHz Version 2.0.0 Ratified» URL: https://www.gs1.org/sites/default/files/docs/epc/Gen2_Protocol_Standard.pdf
53. RF and Microwave Circuit Design for Wireless Communications, Larson Lowrence E., editor. Artech House. London. 1996.
54. HSPA to LTE-Advanced: 3GPP Broadband Evolution to IMT Advanced (4G) – 3GAmericas, Rysavy Research, September 2009.
55. GPP TS 23.122. Technical Specification Group Core Network and Terminals; Non-Access-Stratum (NAS) functions related to Mobile Station (MS) in idle mode (Release 9).
56. 3GPP TR 25.913. Requirements for Evolved UTRA (E-UTRA) and Evolved UTRAN (E-UTRAN) (Release 9).
57. 3GPP TS 36.211. Evolved Universal Terrestrial Radio Access Network; Physical channels and Modulation (Release 9).
58. C. Oetges, B. Clerckx MIMO Wireless Communications: From Real-world Propagation to Space-time Code Design– Academic, 2007.
59. H. Kaaranen., A.Ahtiainen, L. Laitinen, S.Naghian, V.Nienu UMTS Networks: Architecture, Mobility and Service. (Second Edition). John Wiley & Sons, Ltd.,2005.
60. ESTI EN 302 304. Digital Video Broadcasting (DVB); Transmission System for Handheld Terminals (DVB-H).
61. W. Dargie, V. Poclabauer Fundamentals of Wireless Sensor Networks. Theory and Practice. John Wiley & Sons Ltd.. 2010.
62. V. K Ingle., J. G. Proakis Digital Signal Processing Using MATLAB. Brooks/Cole Publishing Company, 2000.
63. V. Garg Wireless Communications and Networking. Morgan Kaufman Publishers, Elsevier. 2007
64. V.V. Logvinov Receivers of Fixed and Mobile Communications (in Russian). Solon-Press. 2016.
65. 3GPP TS 38 141-2. 3rd Generation Partnership Project; Technical Specification Group Radio Access Network; NR; User Equipment (UE) radio transmission and reception; Part 1: Range 1 Standalone (Release 15).
66. 3GPP TS 38.104. NR; Base Station (BS) radio transmission and reception
67. 3GPP TS 38.211. NR; Physical channels and modulation.
68. 3GPP TS 38.300. NR; NR and NG-RAN Overall Description; Stage 2
69. .ETS 300 175 – 1; Radio Equipment and System (RES); Digital European Cordless Telecommunications (DECT) Common Interface Part 1: Overview
70. 3GPP TS 45.005 V9.02 3rd Generation Partnership Project; Technical Specification Group GSM/ADGE Radio Access Network; Radio Transmission and Reception (Release 9), 2010-03.
71. R.A Scholtz. et al. UWB radio deployment challenges. In Proc. IEEE PIMC, vol. 1 Sept. 2000.
72. Full LTE architecture and components, <https://yatebts.com/documentation/concepts/lte-concepts/>
73. V. V. Logvinov Elements of Radio Receivers of Digital Radio Electronic Systems (is printed)
74. EU FP7 METIS project, “Deliverable 6.4 – Final report on architecture”, January 2015.
75. Networks Functions Virtualization (NFV); Management and Orchestration, ETSI GS NFV MAN 001 V1.1.1 (2014-12)
76. P. Rost, A. Banchs, I. Berberana, M. Breitbach, M. Doll, H. Droste, C. Mannweiler, M.A. Puente, K. Samdanis and B. Sayadi, “Mobile Network Architecture Evolution towards 5G,” in IEEE Communications Magazine, vol. 54, no. 6, June 2016.
77. 3GPP TR 38.913 v.14.2.0 Release 14 Study an Scenarios and Requirements for Next Generation Access Technologies
78. EU PROJECT METIS-II, 5G RAN Architecture and Functional Design White Paper, <https://metis-ii.5g-ppp.eu/wp-content/uploads/5G-PPP-METIS-II-5G-RAN-Architecture-White-Paper.pdf>
79. J. J. Benedetto, C. Heil, and D. F. Walnut, “Gabor Systems and the Balian-Low Theorem,” in Gabor Analysis and Algorithms, H. G. Feichtinger and T. Strohmer, Eds. Birkh  user, 1998, ch. 2, pp. 85–122
80. V.K. Garg IS-95 and cdma2000. Cellular/PCS Systems Implementation, Prentice-Hall, Upper Saddle River, N.J., 2000
81. Opportunities in 5G Networks: A Research and Development Perspective / Edited Fei Hu, CRC Press Taylor&Francis Group, 2016
82. M. Kottkomp 5G New Radio. Fundamentals, procedures testing aspects. Rhode & Shwarz., 2016
83. 3GGP2 S.R0005-B v.2.0Network Reference Model for cdma2000Spred Spectrum Systems – Revision 13.06.2007.
84. IEEE Std 802.16-2004. IEEE Standard for Local and metropolitan area networks/ Part1b: Air Interface for fixed Broadband Wireless Access Systems. – IEEE, 1 October. 2004
85. ESTI TR 102 802 v0.1.2. (2009–12). Reconfigurable Radio System (RRS); Cognitive Radio System Concept. – Draft Technical Report. 2009

86. ETSI EN 302 208-1. Radio Frequency Identification Equipment operating in the band 865 MHz to 868 MHz with power levels up to 2 W, Part 1: Technical requirements and methods of measurement. v.1.3.1; July 2009. URL: http://webapp.etsi.org/action/V/V20100206/en_30220801v010301v.pdf
87. ETSI EN300 744 v1.6.1 (2009–11) Digital Video Broadcasting (DVB); Framing structure channel coding and modulating for digital terrestrial television. – European Telecommunication Standards Institute. 2009.
88. ETSI 300 401. Radio Broadcasting Systems; Digital Audio Broadcasting (DAB) to mobile, portable and fixed receivers. – European Telecomm. Standards Institute, 2001.
89. IEEE Std 802.16e-2005 and IEEE Std 802.16-2004/Cor 1-2005, Part 16: Air Interface for fixed and mobile Broadcasting Wireless Access Systems. Amendment 2: Physical and medium Access Control Layers for Combined fixed and Mobile Operated in Licensed Band – IEEE, 28 February, 2006.
90. P.A. Laurent Exact and Approximate Construction of Digital Phase Modulations by Superposition of Amplitude Modulated Pulses (AMP). IEEE Trans. Commun., Vol. COM-34, 1986. P. 150–160.
91. J. Crols, M.S.L. Steyaert Low-IF topologies for High-performance analog front ends of fully integrated receivers // IEEE Trans. Circuits Syst.-II. March 1998. Vol.45, №3. P. 269–282
92. J. Strange, S. Atkinson. A direct conversion transceiver for multiband GSM application // IEEE RFIC Digest Papers. 2000. P. 25–28.
93. A.A. Emira, E.A. Sanches-Sinenci. Pseudo Differential Complex Filter for Bluetooth with Frequency Tuning // IEEE Trans Circuits and Syst. Pt. II. Oct.2003. Vol.50, №10. P. 742–754.
94. 3GPP TS 45.005 V9.2.0 3rd Generation Partnership Project; Technical Specification Group GSM/EDGE Radio Access Network; Radio transmission and reception (Release 9), 2010-03.
95. J.L. Bohorquez, A.P.Chandrakasan, J.L. Dawson A 350uW CMOS MSK transmitter and 400 uW OOK super-regenerative receiver for medical implant communications, IEEE J Solid –State Circuits 44 (2009), 1248–1259.
96. Applied Wavw Ressearch, <http://www.awr.com>
97. J. Crols, M. S.J. Steyaert Low-IF topologies for High-performance analog front ends of fully integrated receivers // IEEE Trans. Circuits Syst.-II. March 1998. Vol.45, №3. P. 269–282
98. A World of Wireless Technology. Our World of Wireless Products. Designer's Handbook. – Greensboro: RF Micro Devices, 1999
99. J. Warren, N. Sargoloon PRISM NITRO Introduction.– Intersil. 2003.
100. WWW.ti.com/sc/device/CC2530.
101. M. Ha, J. Kim, Y. Eo A 6 – 10 GHz Noncoherent IR-UWB CMOS Receiver// Microwave and Optical Technology Letters/ 2012. Vol. 54, No.9. P. 2007–2010.
102. GLONASS. Principles of construction and operation / Edited by A. I. Perov, V. N. Kharisov, M.: radio engineering, 2010 (in Russian).
103. J.L. Bohorquez, A.P. Chandrakasan, J.L. Dawson A 350uW CMOS MSK transmitter and 400 uW OOK super-regenerative receiver for medical implant communications, IEEE J Solid –State Circuits 44 (2009), 1248–1259.
104. H-K Cha, X. Yuan, M.K. Raj, M. Je A CMOS Medradio Receiver RF front-end with complementary current-reuse for LNA for Biomedical Applications// Proceedings of IEEE Asian Solid-State Circuits Conference, San Francisco, CA, 2010, pp.53–56.
105. R. Zanbaghi, M. Atarodi, A. Tadjalli An ultra low power Gm-C complex band-pass filter for low-IF wireless PAN applications, In: Proceedings of IEEE TENCON, 2006.
106. H.C. Chen, T. Wang, S.S. Lu A monolithic 5.9 GHz CMOS I/Q direct-down converter utilizing a quadrature coupler and transformer-coupler subharmonic mixers //IEEE Microwave Communication Letters 16, (2006), P.197–199.
107. H.C. Chen, C-W Tsou, Y-S Lin A Single-Chip 24-GHz Differential I/Q Receiver in 0.18- μ m CMOS Technology// Microwave and Optical Technology Letters/ 2011. Vol. 53, No.11. P.2593–2600.
108. S. Haykin Digital communication – John Wiley&Sons, Inc., New York, 1994.
109. P.H. Lehne, M. Pettersen An Overview of smart antenna technology for mobile communication system, IEEE Communication Surveys and Tutorials, <http://www.cosmoc.org/publicatins/surveys/>, vol.2, №4, Fourth Quarter 1999
110. A. Viterbi The Orthogonal-Random Waveform Dichotomy for Digital Mobile Personal Communications// IEEE Personal Communications, First Quarter 1994, pp. 18–24.
111. Li Chunjian Efficient Antenna Patterns for Three-Sector WCDMA Systems. – Master of Science Thesis. Chalmers University of Technology, Ericsson AB. Sweden. 2003
112. MIMO Transmission Schemes for LTE and HSPA Networks – 3G Americas. June, 2009.
113. C. Oestegs, B. Clerckx MIMO Wireless Communications: From Real-world Propagation to Space –time Code Design. – Academic, 2007.
114. 3GPP TS25.101. User Equipment (UE) radio transmission and reception (FDD).
115. K.Wesolovsky Introduction to digital communication systems. – John Wiley&Sons. 2009.
116. “In the Matter of Revision of Part of the Commission's Rules Regarding Ultra-Wideband Transmission Systems” (ET Docket 98-153) Federal Communications Commission. Memorandum Opinion and Further Notice of Proposed Rule Making. Adopted: February 13, 2003, Released: March 12, 2003.
117. U.L. Rode Eight ways to better radio receiver design. – Electronics, 1975, №4, p.87–91.
118. A. Van der Ziel, K.M.H.F. Van Vliet Thermal Noise in Space-charge Limited Solid-state Diodes. – Solid State Electron., 1968, v.11, № 4, p.508–509.
119. K.Kurokawa Power waves and the scattering matrix // IEEE Trans. Microwave Theory and Tech. March 1965. Vol.MTT-13, №2. P.194–202.
120. B.A. Gilbert A precise four quadrant multiplier with subnanosecond response // IEEE J. Solid-State Circuits. Dec. 1968. Vol. SC-3, №4. P.365–373.
121. PSipce User's guide. MicroSim Corporation. – La Cadena Drive, Laguna Hills, 1989.
122. W. Shockley, J.A. Copeland, R.P. James The Impedance Field method of Noise Calculations in Active Semiconductor Devices// In: Quantum Theory of Atoms, Molecules and the Solid State, N-Y: Academic Press, 1966.

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