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COMMUNICATION SYSTEMS THROUGH ARTIFICIAL EARTH SATELLITES

(Selected Pages)

by

N. I. Kalashnikov

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MICROFICHE NR: FTD-87-C-000120

COMMUNICATION SYSTEMS THROUGH ARTIFICIAL EARTH SATELLITES (Selected Pages)

By: N. I. Kalashnikov

English pages: 616

Source: Sistemy Svyazi Cherez Iskusstvennyye Sputniki Zemli, Publishing House "Svyaz'", Moscow, 1969, pp. 1-151; 159-286; 307-313; 345-383

Country of origin: USSR
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>
> Page 51.
>
> However, in recent years it has been shown that with the hereditary pigment degenerations in animals the biochemical changes are observed in all celluar elements of the retina.

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Block	Italic	Transliteration	Block	Italic	Transliteration
А а	*А а*	A, a	Р р	*Р р*	R, r
Б б	*Б б*	B, b	С с	*С с*	S, s
В в	*В в*	V, v	Т т	*Т т*	T, t
Г г	*Г г*	G, g	У у	*У у*	U, u
Д д	*Д д*	D, d	Ф ф	*Ф ф*	F, f
Е е	*Е е*	Ye, ye; E, e*	Х х	*Х х*	Kh, kh
Ж ж	*Ж ж*	Zh, zh	Ц ц	*Ц ц*	Ts, ts
З з	*З з*	Z, z	Ч ч	*Ч ч*	Ch, ch
И и	*И и*	I, i	Ш ш	*Ш ш*	Sh, sh
Й й	*Й й*	Y, y	Щ щ	*Щ щ*	Shch, shch
К к	*К к*	K, k	Ъ ъ	*Ъ ъ*	"
Л л	*Л л*	L, l	Ы ы	*Ы ы*	Y, y
М м	*М м*	M, m	Ь ь	*Ь ь*	'
Н н	*Н н*	N, n	Э э	*Э э*	E, e
О о	*О о*	O, o	Ю ю	*Ю ю*	Yu, yu
П п	*П п*	P, p	Я я	*Я я*	Ya, ya

*ye initially, after vowels, and after ъ, ь; e elsewhere.
When written as ё in Russian, transliterate as yё or ё.

RUSSIAN AND ENGLISH TRIGONOMETRIC FUNCTIONS

Russian	English	Russian	English	Russian	English
sin	sin	sh	sinh	arc sh	$\sinh^{-1}$
cos	cos	ch	cosh	arc ch	$\cosh^{-1}$
tg	tan	th	tanh	arc th	$\tanh^{-1}$
ctg	cot	cth	coth	arc cth	$\coth^{-1}$
sec	sec	sch	sech	arc sch	sech^{-1}
cosec	csc	csch	csch	arc csch	csch^{-1}

Russian	English
rot	curl
lg	log

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COMMUNICATION SYSTEMS THROUGH ARTIFICIAL EARTH SATELLITES.

N. I. Kalashnikov.

Year of publication 1969

In monograph are examined the bases of calculation and principles of construction of communication systems through active Earth satellites, intended for transmission of large number of telephone conversations and television programs.

In book are presented questions, which relate to selection of orbits of satellites, to calculation of levels of signals and qualitative indices of communication channels with different methods of modulation. At the same time the analysis of the special features of the joint operation of communication systems through ISZ [AES - artificial earth satellite] and radio relay lines of sight in the overall frequency band is given.

Monograph can be considered as introduction to complicated problem of construction of contemporary communication systems through ISZ. It is designed for the radio specialists and can be useful for the students of the old courses of radio engineering departments and graduate students.

Preface.

Starting/launching by Soviet Union on 4 October, 1957, first artificial Earth satellite (ISZ) was indicated real possibility of using Earth satellites as relay stations. This accelerated the work of Soviet and foreign scientists and engineers in the field of the creation of new radiolink systems - communication systems through ISZ.

On 23 April, 1965, in Soviet Union were launched communication satellite "Molniya-1", intended for transmission of large number of telephone conversations, television programs (image and sonic accompaniment), telegraph and phototelegraphic signals [1].

Results of investigation and testing communication systems through ISZ "Molniya-1" and American communication satellites Telstar, Syncom and Intelsat showed that similar communication systems make it possible to ensure high-quality transmission of television programs and large number of telephone conversations up to considerable distances. Therefore in recent years attention given to the communication systems through ISZ grew considerably: the number of stations, which by 1970 will be established/installed in different countries, according to the data of the International Union of Electrical Communication, will equal 94 [2].

To special features of communication systems through ISZ is

devoted considerable quantity of literature, for example, work [1, 3, 4, 5, 6, 7, 8, 9]. However, the information, available in those enumerated and in other publications, proved to be separated and presented insufficiently in detail, which hindered/hampered their use.

Interest in communication systems through ISZ showed advisability of publishing book, in which would be examined and analyzed bases of construction and characteristic features of such systems. At the same time it was desirably systematize information along the communication systems through ISZ. The proposed monograph is dedicated to the solution of the enumerated problems.

Page 4.

During writing of book were taken into consideration work of many Soviet and foreign authors, materials of International Radio Consultative Committee (MKKR). The author considers it necessary to note the high value of the labor of S. V. Borodich [10; 11; 12], V. A. Smirnov [13] and B. R. Levin [14]. In the book some published work of the author and materials set forth by the author when taking courses at the Moscow electrical Institute of Communication are included.

Many-sidedness of questions, which relate to communication systems through ISZ, impedes their examination. In this work are examined questions, in essence relating to the principles of calculation and construction of the communication systems through ISZ

during the active relaying of signals; reducing, relating to the selection of diagrams and nodes of apparatus and equipment, they cannot pretend to the completeness of presentation.

Book is intended for scientific and technical-engineering workers, who carry out development and operation of communication systems, and also to students of schools of higher education, which specialize in region of connection/communication.

I consider it my pleasant duty to express appreciation to A. L. Badalov and to Honored Scientist and Technologist, Doctor of Technical Sciences Prof. B. P. Terent'yev for his advice, which was considered during the writing of monograph. The author is very grateful to Candidate of Tech. Sciences Docent M. V. Nazarov for the valuable observations, made with the survey of the manuscript.

Author expresses sincere gratitude to reviewer of the book K. K. Ogorodnikov for his extremely thorough examination of manuscript and are important advice and observations.

Critical observations and wishes should be sent to address: Moskva-tsentr, Chistoprudnyy bul'var, d. 2, izd. "Svyaz'".

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Chapter 1.

Principles of the construction of communication systems through artificial Earth satellites.

1.1. General information.

In communication systems through artificial Earth satellites (ISZ), satellites perform role of relay station. Let us examine the diagram of connection/communication, given in Fig. 1.1.1.

Here A and Б - points/items, between which is established communication, and straight lines AA′ and ББ′ - tangents to surface of Earth, at points A and Б they are lines of horizon of these points/items. From the figure it is evident that ISZ-1, moving/driving along orbit MN, simultaneously can be observed in points/items A and Б only in the section of orbit A'-Б'. During the motion in this section the electromagnetic vibrations, emitted by the antenna system of point/item A toward ISZ-1, are received as the onboard radio equipment of satellite, and then immediately, after amplification, they are headed in the direction of the Earth and they can be accepted in point/item Б.

Radiolink system in presence of onboard equipment is called system with active relaying of signal or system with active satellite.

System with active relaying of signal depending on orbit altitude and distance between correspondents can be carried out as system with instantaneous relaying (system on real time) and as system with delayed relaying. During the transmission of identical with respect to space communications/reports, for example, during the transmission of television program, the communication system with the delayed relaying will have the more elaborate complex of installed equipment, than system with the instantaneous relaying, since in the latter is not necessary equipment of the memory.

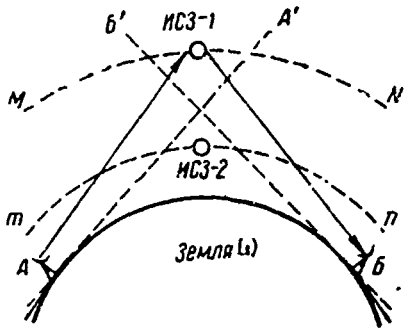

Fig. 1.1.1. Principles of realization of radio communication through ISZ.

Key: (1). Earth.

Page 7.

Let us examine case of moving satellite ISZ-2 along orbit mn (Fig. 1.1.1) with this low altitude that it cannot simultaneously be observed of points/items A and *Б* (orbit altitude lower than point of intersection line of horizon AA ' and *ББ'*). It is obvious that the signal, which enters the satellite ISZ-2, cannot be immediately transmitted to correspondent *Б*.

In this case radiolink system is fulfilled as follows: satellite, flying above correspondent A, accepts communications/reports, which after preliminary amplification are supplied to onboard equipment of memory. Above point/item *Б* on the satellite is switched on onboard transmitter and is transmitted the information, obtained by satellite from correspondent A. The start of

transmitter can be realized by supply of the special control signal, emitted by correspondent Б at the moment of the appearance of satellite ISZ-2 in the visibility range, or by onboard program unit, which calculates the satellite velocity on the basis of orbit, its height/altitude and the distance between the correspondents.

Satellite ISZ-2 with equipment of memory is aboard called satellite-messenger, and radiolink system, which uses such satellites - by communication system with memory or communication system with delayed relaying.

From apparently, more expedient for transmission of comparatively narrow-band signals - small number of telephone conversations or telegraph information, since in this case equipment of memory is more simple. In connection with this subsequently entire presentation will be led in connection with systems with the instantaneous relaying of communications/reports.

For transmission of communications/reports from point/item A to point/item Б can be proposed second method of radio communication, with which on board satellite there will not be radio equipment, but signals, sent from point/item A, will be reflected by surface of satellite ISZ-1 in direction of Earth without preliminary amplification. In this case the signals in point/item Б will be weaker than in the first version, but with a sufficient antenna gain and the sensitive receivers this method in a number of cases is

applicable. The latter/last communication system is called the communication system with the passive relaying or system with the passive satellite. It can be carried out only with the immediate relaying of signal, i.e., as system on real time.

One should mention that for purposes of simplification given examination was conducted for case, when transmission of signals was realized from one correspondent A to the next *B*. Transmission to the reverse side is realized analogously.

Page 8.

Theoretically, communication satellites can utilize any orbits; however, equatorial circular orbit, distant from surface of Earth up to distance of approximately 36000 km, is most interesting. As it will be shown below (see Chapter 2), when the direction of the motion of satellites in this orbit coincides with the direction of rotation of the Earth, to ground-based observer satellite will seem fixed (stationary satellite).

During use of three stationary satellites, located in equatorial plane at angle of 120° (Fig. 1.1.2), possible to create a world-wide communication system.

In this case from station at point I will be transmitted communications/reports to satellites ISZ-1 and ISZ-2, and from them - to the Earth. If we at point II establish the receiving-transmitting

station, which will relay program from ISZ-2 and ISZ-3, and the latter will realize transmission in the direction of the Earth, then the communication/report transmitted from point I, can be accepted in any place by terrestial globe (except the regions, which lie higher than 76.5° northern and south latitudes).

For realization of connection/communication from point II with any point on terrestial globe, installation of ground relay station at point I or III is necessary. During the transmission from point III the satellites and ground-based repeaters at point I or II are utilized. In this communication system can be used both the active and passive relaying of signals.

Important advantage of connection/communication through stationary satellites consists in the fact that reception and transmission of signals can be realized with fixed ground-based antennas. In all other cases are necessary more complicated mobile antenna systems, and also special equipment of tracking or program unit, which controls the motion of antenna. However, this complication can be justified by simpler and cheaper starting systems of satellite, by possibility of injection into orbit of satellite with the large weight with the same rocket engines. This makes it possible to increase a number of channels, relayed by satellite, or to raise the reliability of connection/communication due to the installation on board the more powerful transmitters, the sensitive receivers or the assemblies of stand-by.

Together with this of communication system through satellites, which move along inclined elliptical or circular orbits, they make it possible to ensure connection/communication with any points on surface of Earth, including with territories, which are located higher than 76.5° northern and south latitudes.

During comparison of orbits of satellites one should also note that use of stationary satellites is allowed for correspondents, who observe it simultaneously, to support long continuous bond.

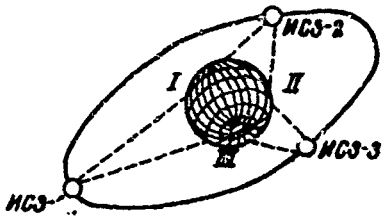

Fig. 1.1.2. Diagram of connection/communication through ISZ.

Page 9.

During the use of transient satellites, the connection/communication between correspondents A and Б (Fig. 1.1.1) will be only in the period of the motion of satellite on the section of orbit Б' A'.

Guarantee of long continuous link with comparatively low orbits is possible only with increase in number of satellites (Fig. 1.1.3). In this case in each ground-based point/item must be established two antennas a_1 and a_2, which can realize transmission and reception of signals, utilizing one of the satellites, for example, ISZ-3, that is located in the zone of the simultaneous visibility Б'-A'. When ISZ-3 leaves the visibility range, connection/communication will be realized with the aid of antennas a_2 and satellite ISZ-2. When satellite ISZ-2 leaves the visibility range, the transmission and the reception of signals must be realized by means of the satellite ISZ-1 and antennas a_1, directed toward this satellite and, etc. It is obvious that for obtaining the continuous connection/communication the distance between adjacent satellites must be less than the zone of simultaneous visibility, so forth. A number of satellites with this form of

communication is great. Depending on the distance between the correspondents, heights/altitudes, forms also of orbit inclination their number can oscillate from 10-12 to 40-50.

Communication system with several satellites can be both with active and with passive relaying of signals. Each of these methods has their advantages and disadvantages. During the passive relaying the strength of field in the sector of reception/procedure proves to be very low, so that it requires the essential complication of the ground-based equipment: the application of bulky antennas, powerful transmitters and sensitive receivers with the parametric and molecular amplifiers. However, due to the absence of onboard equipment, repeater proves to be simpler, and system with the passive relaying possesses larger reliability. Furthermore, as is noted in [1.1], width of band of the relayed signals in this case is unlimited, since the passive communications relays are linear systems and therefore can be used simultaneously in several lines connections/communications, which work at different frequencies and with different power levels, without interferences. Thus, obtaining new channels or lines of communications requires only the construction of supplementary terrestrial points/items or corresponding increase in the complex of the terrestrial receiving-transmitting equipment.

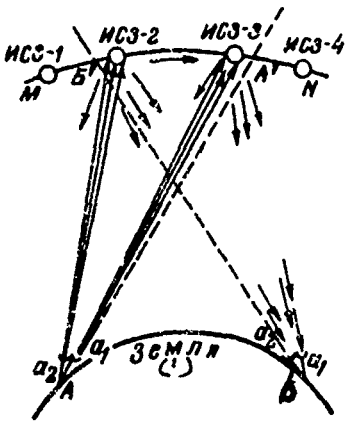

Fig. 1.1.3. System with several ISZ.

Key: (1). Earth.

Page 10.

In contrast to this active repeaters have limited frequency and dynamic range and, consequently, can be used for transmission only specific numbers of signals.

Considerably best energy indices, which make it possible to utilize simpler ground-based antennas and transmitters of smaller power, are important advantage of active relaying during transmission of signals, which occupy large frequency band.

One should emphasize that physical special features of communication systems with passive communications relays in principle make it possible to carry out transmission of television and hundreds of telephone signals. However, the level of contemporary technology yet does not make it possible to realize these possibilities during

the use of the means, compared with those, which are used during the active relaying of signals. For this reason the communication systems with the passive relaying thus far can be utilized only for the transmission of narrow-band signals, which very is in detail examined in several experimental and theoretical works [1.2, 1.3, 1.4, 1.5, 1.6].

In description and during analysis of communication systems, intended for transmission through ISZ of television or a large number of telephone signals, which occupy large frequency band on video and through group circuit, are television, subsequently we will proceed from diagrams in Fig. 1.1.1 and 1.1.3, assuming use only of active repeaters.

Let us examine now most common block diagram of connection/communication, between points/items A and Б through ISZ, shown in Fig. 1.1.4.

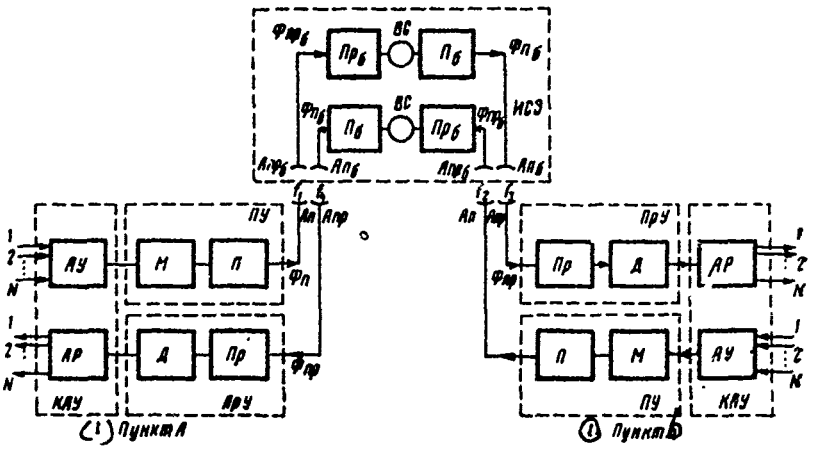

Fig. 1.1.4. Common system block diagram of connection/communication through ISZ.

Key: (1). Point.

Page 11.

Transmitted by subscribers 1, 2, ..., N signals, which can have identical frequency spectra, are fed/conducted to equipment for multiplexing channels AU. Designation/purpose of the latter - to ensure transmission along one line of communications of several signals. Systems for multiplexing channels with respect to frequency and time at present extensively are used.

From AU oscillations are fed/conducted to modulator M of transmitter PU, in which is included transmitter P. Then on the feeder of transmitter $\Phi_{п}$ the modulated oscillations/vibrations with a carrier frequency of f_1 are emitted by that transmitting antenna $A_{п}$ toward ISZ.

On board satellite of oscillation/vibration of carrier frequency f_1 will be received by receiving antenna $A_{пр_б}$ and on feeder of onboard receiver $Ф_{пр_б}$ they are conducted/supplied to input of receiver $Пр_б$. After amplification and necessary conversions in the auxiliary step/stage BC (for example, after transition/junction from the carrier frequency f_1 for a lower frequency f_3) these oscillations/vibrations will enter the input of onboard transmitter $П_б$ and through feeder $Ф_{п_б}$ to the onboard transmitting antenna, which radiates in the direction of the Earth oscillations/vibrations with frequency f_3.

Oscillations/vibrations of carrier frequency f_3 can be accepted terrestrial of receiving antenna $A_{пр}$ and then to feeder $Ф_{пр}$ they are sent to input of receiver $Пр$, which is part of complex of receiver ПрУ. After the passage of detector Д the oscillations/vibrations are fed/conducted to the equipment for the separation of channels AP, and then proceed to correspondents.

Connection/communication from one point/item Б to the next A is similar examined; however, for preventing interferences it is necessary to utilize other frequencies f_2 and f_4. Thus, for the duplex operation in systems of communication through ISZ are required four frequencies.

From examination of Fig. 1.1.4 follows that during transmission from point/item Б to point/item A is utilized equipment, similar to volume, which is used for connection/communication in direction from A

to Б. Let us note that equipment for multiplexing АУ and separation of channels АР on each point/item enters into the general/common complex of equipment for multiplexing КАУ, moreover territorially complex КАУ can be placed both in immediate proximity of the transmitting and receiver ПУ and ПрУ and at the considerable removal/distance. In the latter case КАУ must be connected with the transmitting and receiving ПУ and ПрУ, the appropriate cable or radio relay lines.

At the same time it is important to note that all remaining connecting lines (feeders $\Phi_{п}, \Phi_{пр}$) they must be as shortly as possible, since in this case quality and reliability of connection/communication is raised.

During the transmission of telephone conversations, diagram in Fig. 1.1.4 must be supplemented by differential systems. The latter provide transition/junction from the four-wire skeleton of complex of equipment for multiplexing (two wires to АУ, two wires from АР) to the two-wire circuit (Fig. 1.1.5).

Page 12.

Fig. 1.1.5 of broken arrow shows passage of signal of subscriber, located at point A. The speaking currents of this subscriber from equipment АР come through ДС only into the two-wire circuit, but also are branched/shunted to АУ, and then - along the line the connections/communications are supplied back into point/item A to the

same subscriber. As a result the subscriber at point A will hear out his speech in the form of the echo of signal in hand receiver. Echo level of signal will be determined by the quality of balancing/trimming ДС.

Analogous to that examined during poor balancing of ДС at point A echo signals will be heard by subscribers of point/item *Б*.

Is especially noticeable echo signals as interferences, which interfere with conversation, on lines of communications of large extent, where between signal of subscriber and echo by signal is obtained considerable time lag (several hundred milliseconds). Its this value can be observed in the communication systems through the satellite.

Let us emphasize that diagram in Fig. 1.1.4 allows for possibility of operation of two terrestrial stations through one communication satellite. Analogous diagrams can be used, also, for a larger number of terrestrial stations. The communication system, in which the communication satellite is utilized (it proves to be available) for operation of several terrestrial stations, is called the system of multistation communications or system with the multistation access.

Installed equipment of satellite can be constructed according to principle of single and multiplex operation. In the first case the

diagram of repeater is analogous to diagram in Fig. 1.1.4. Example of the multiplex operation of installed equipment - diagram in Fig. 1.1.6. In this case one and the same complex of installed equipment simultaneously is utilized for relaying of the communications/reports of two terrestrial stations; the reception of communications is conducted at frequencies f_1 and f_2, and transmission - at frequencies f_3 and f_4.

It is obvious that single and multiplex operation of repeater can be realized not only with two Earth stations, but also with their any number.

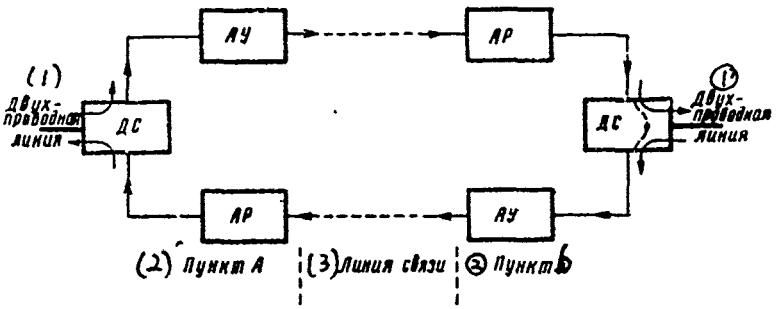

Fig. 1.1.5. Connection of differential system.

Key: (1). Two-conductor line. (2). Point/item. (3). Line of communications.

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In particular, the multiplex operation of equipment is possible in the case of the transmission of signals only from one terrestrial station. In this case the equipment component (antenna, amplifier, etc.) can repeatedly be utilized for successive reception and transmission or amplification of one and the same communication/report, if, certainly, each time will be conducted the corresponding shift/shear of the carrier frequency.

Thus, multiplex operation of equipment of repeater is in principle possible both from one terrestrial station and in multistation connection/communication.

Comparison of diagrams in Fig. 1.1.4 and 1.1.6 shows that with multiplex operation number of elements, which make up communication equipment of repeater, substantially decreases. However, the

requirements, presented to the repeatedly working elements, prove to be more complicated.

Let us note that multistation communication systems can be constructed both with frequency separation of signals of those emitted by terrestrial stations (in abbreviated form designated MCh), and with temporary/time separation (MV). In the latter case through one repeater successive operation of several Earth stations is realized.

Some questions, which relate to multistation systems and multiplex operation, are examined in Chapter 12.

1.2. Terminology.

Terminology and definitions for many questions, which relate to communication systems through ISZ, have at present been established. Therefore are given below the definitions of some terms, whose part is utilized in this book. Sign * noted the terms, which were accepted by the X or XI Plenary assemblies of MKKR.

Active communication satellite * - satellite, in which for connection/communication is utilized onboard radio station.

Deep space * - space at a distance, commensurate with distance of Moon or larger.

Terrestrial station * - station of space service, arranged/located on surface of Earth (including stations on board ship or on mobile ground-based object) or on board aircraft.

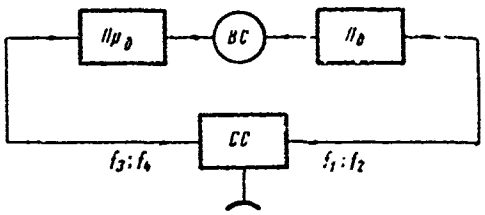

Fig. 1.1.6. Version of block diagram of relay.

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Space service * - radio communication service:

- between terrestrial to by space stations;
- or between space stations;
- or between terrestrial stations, when signals are relayed by space stations or are transmitted by means of reflection from objects, which are found in space, eliminating reflection or scattering of ionospheric or in limits earth's atmosphere.

Space station * - station of space service, arranged/located on object, which is intended for shipment beyond limits of basic part of earth's atmosphere either is located or was located beyond these limits.

Spacecraft * - any type of space means of movement, including Earth satellites, probes of deep space etc., manned or unmanned.

Let us note that in the definition of spacecraft its sizes/dimensions do not figure; therefore by spacecraft can be

understood not only very large means of transportation in space, but also objects with very small dimensions.

Multiple-synchronous satellites * - satellites, for which sidereal period of orbit of fundamental body around its own axis into whole number of times is more than average sidereal circling time of satellite around fundamental body.

Weather service, which uses Earth satellites * - space service, in which results of meteorological observations, carried out with the aid of instruments, are established/installed henna Earth satellites, they are transmitted to terrestrial stations from space stations of these satellites.

Ground station - station of nonspace service.

Nonspace services * - general/common term, applied to any or to all radio services, with exception of space service.

Orbital surface - surface, motion by which coincides with motion of Earth, formed by traces (turns) of satellite for long time.

Passive satellite * - satellite, intended for relaying of signals, scattered or reflected from it (into this concept they enter tanks/balloons of type "EKhO", orbiting dipoles, Moon, etc.).

Service or system of communications which uses Earth satellites (ISZ) *- space service or radiolink system between terrestrial stations, in which signals are relayed by many one or space stations or are transmitted by reflection or scattering from objects, which are found in orbit around Earth.

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Service of radio broadcasting, which uses Earth satellites * - space service, in which signals, transmitted either by space stations, or by reflection or scattering from objects, which are found in orbit around Earth or are intended for direct reception by population.

Synchronous satellites * - satellites, for which average sidereal period of orbit about fundamental body is equal to sidereal period of orbit of fundamental body around its own axis (latter for earth/ground it is approximately 23 hours of 56 min.).

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Chapter 2.

Motion of satellites and duration of periods of communication.

2.1. Motion of satellites.

Earth satellites can be launched on they are shaved, that are characterized by following parameters:

a) by orbit inclination, i.e., by angle between equatorial plane and orbital plane. In accordance with this there can be the equatorial orbits (orbit inclination equal to zero), the polar orbits (orbit inclination equal to 90°) and inclined orbits (during any other inclination);

b) by the shape of the orbit (circular or elliptical);

c) with orbit altitude above the surface of the Earth or the distance from the orbit to the center of the Earth.

During motion of ISZ [ИСЗ - artificial earth satellite] on it acts not only attraction of Earth, but also gravitational field other celestial body - Moon, sun, planets. Furthermore, the motion of satellite depends on the components of the gravitational force, which

appear as a result of the noncentral nature of the gravitational field of the Earth, and also due to the force, which appears as a result of the effect of magnetic field. During the motion of satellite along the orbits with the low altitude essential effect resists of the atmosphere.

Motion of satellite under effect of all enumerated forces will endure disturbances/perturbations and deviations from trajectory, determined, in the first approximation, only by attracting force of Earth. The motion of satellites around the Earth is explained by Kepler's laws. According to the first of them, one of the foci of the elliptic orbit of satellite or center of circular orbit must be located in the center of mass of the Earth. Hence follows that ISZ, in contrast to the projectiles or the aircraft, cannot fly on any trajectory above the surface of the Earth, but it can move only along such orbit, whose plane compulsorily passes through the center of mass of the Earth, in particular ISZ cannot move along any parallel of terrestial globe, with exception of equator (equatorial orbit). Page 17.

Let us determine speed of undisturbed motion of ISZ, which moves along circular orbit with radius of r, calculated off center of Earth. In this case according to the known formula, which determines the equilibrium of forces of gravity $F_т = k\frac{Mm}{r^2}$ and centrifugal force $F_ц = \frac{mv_к^2}{r}$ we will obtain:

$$v_к = \sqrt{\frac{\kappa M}{r}}, \qquad (2.1.1)$$

where

k=6.67·10⁻⁸ - constant of gravity, cm³/gs².

M - 5.974·10²⁷ - mass of Earth, g.

m - mass of ISZ, g.

$v_к$ - circular velosity of motion of ISZ, cm/s.

Expression for circular velocity can be registered on other, using value of circular velocity v_0 of hypothetical satellite, which moves along circular orbit with radius, equal to radius of Earth R=6.4·10⁸ cm.

Taking into account, given above values, let us find v_0=7.9·10⁵ cm/s; after substitution we will obtain

$$v_к = 7{,}9 \cdot 10^5 \sqrt{\frac{R}{r}}, \text{ cm/s}. \qquad (2.1.2)$$

Taking into account value R and passing from dimensionality in centimeters to kilometers, let us find

$$v_к = \frac{630}{\sqrt{r_{[км]}}}, \text{ km/s}. \qquad (2.1.3)$$

Value $v_{к}$, which is called orbital velocity, with increase in r is reduced.

Period of orbit of ISZ, which moves along circular orbit with radius of r, can be found from expression

$$T_к = \frac{2\pi r}{v_к}. \qquad (2.1.4)$$

Taking into account value corrected above for $v_{к}$, let us find

$$T_к = 10^{-2}\sqrt{r^3_{[км]}},\ с = 1{,}66 \cdot 10^{-4}\sqrt{r^3_{[км]}},\ мин. \qquad (2.1.5)$$

For hypothetical satellite, which moves along circular orbit with radius, equal to equatorial radius of Earth, period of orbit T_0=84.4 min or 5064 s.

Upon consideration of value T_0 circling time of satellite along circular orbit can be registered in the form

$$T_к = T_0\left(\frac{r}{R}\right)^{3/2}, \qquad (2.1.6)$$

where R=6400 km are radius of Earth.

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Let us note that in all resulting expressions radius of orbit of satellite can be expressed through radius R Earth and orbit altitude above surface H Earth: in this case it is obvious that previously introduced value of radius of orbit r=R+H.

Table 2.1.1 gives values, which determine period of orbit $T_к$ and speed $v_к$ of satellite, which moves along circular orbit of radius r, which corresponds to height/altitude above surface of Earth H=r-6400 km.

In accordance with data of Table 2.1.1, Fig. 2.1.1 gives dependences $v_к$ and $T_к$ on height/altitude H.

Angular satellite velocity along circular orbit can be determined according to expression

$$\omega_к = \frac{2\pi}{T_к}, \text{ rad/s.} \qquad (2.1.7)$$

Upon consideration of expressions (2.1.4) and (2.1.3)

$$\omega_к = \frac{630}{\sqrt{r^3[км]}}, \text{ rad/s.} \qquad (2.1.8)$$

Table 2.1.1.

r, км	H, км	$T_к$, мин (1)	$v_к$, $\frac{км}{сек}$ (2)
6 400	0	84,4	7,9
6 900	500	94,5	7,62
7 400	1 000	105	7,35
11 400	5 000	201,2	5,32
13 400	7 000	256,5	5,41
16 400	10 000	349	4,93
26 600	20 200	720	3,89
36 400	30 000	1117	3,3
42 270	35 870	1440 (24 часа) (3)	3,06
46 400	40 000	1664	2,92
56 400	50 000	2231	2,66
106 400	100 000	5784	1,94

Key: (1). min. (2). km/s. (3). hour.

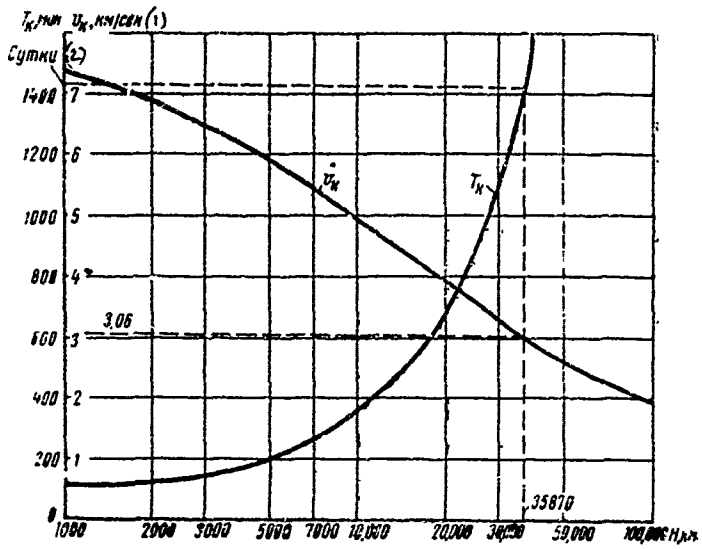

Fig. 2.1.1. Dependence $v_к$ and $T_к$ on orbit altitude.

Key: (1). min ... km/s. (2). Days.

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Let us pass to examination of fundamental principles, which determine motion of satellite according to elliptic orbit, Fig. 2.1.2.

For describing elliptic orbit following fixed points and lines usually are used:

- perigee altitude $H_П$ - distance from Earth to nearest point of orbit,

- apogee altitude H_A - distance from Earth to outermost point of orbit,

- line of apsides - line, which connects apogee points and

perigee,

- semimajor axis of orbit a - distance, equal to half of length of line of apsides:

$$a = \frac{r_A + r_П}{2}.$$

Here

r_A - distance from the center of the Earth to the apogee,

$r_П$ - distance from the center of the Earth to the perigee,

c - the focal length (see Fig. 2.1.2),

e - orbit eccentricity, defined as the ratio of focal length to the semimajor axis, i.e.

$$e = \frac{c}{a} = \frac{r_A - r_П}{2a}. \qquad (2.1.9)$$

Position ISZ in orbit is characterized by true anomaly ϑ, which is determined by angle between line, which connects center of Earth with perigee and line, drawn from center of Earth into that point of orbit, in which is located ISZ. True anomaly is counted off from the line, which connects the focus of orbit and perigee point in the direction of motion of ISZ (see Fig. 2.1.2). At the same time the position of ISZ can be characterized by eccentric anomaly E, determined by the angle between the lines, which connect the center of elliptic orbit with the perigee point and point p, which is found in conditional circular orbit of radius a. Point p is obtained as a result of the intersection of circular orbit with the straight line, perpendicular to the line of apsides АП (AP) and passing through that point in the elliptic orbit, in which at the moment of time in question is located the satellite.

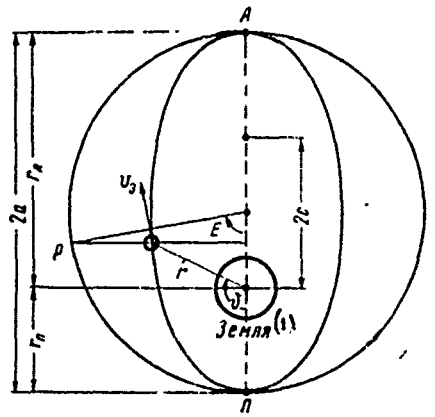

Fig. 2.1.2. Orbital parameters.

Key: (1). Earth.

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True anomaly and eccentric anomaly are connected with following equ.

$$\vartheta = 2\operatorname{arc\,tg}\left(\sqrt{\frac{1+e}{1-e}}\operatorname{tg}\frac{E}{2}\right) \qquad (2.1.10)$$

or

$$\cos\vartheta = \frac{\cos E - e}{1 - e\cos E}. \qquad (2.1.11)$$

Sometimes for describing motion of ISZ is used auxiliary parameter, called mean anomaly $\vartheta_{ср}$. Mean anomaly $\vartheta_{ср}$ is determined by the value of arc, which the satellite after the passage of perigee would describe, if it moved evenly along the conditional circular orbit of radius a, completing the complete revolution in the same time, as during the motion along the elliptic orbit.

Average/mean angular velocity of ISZ during its predicted motion along circular orbit indicated will be

$$\omega_{кв}=\frac{2\pi}{T_з}=\frac{630}{\sqrt{a^3_{[км]}}},\ км/с$$

where $T_з$ - period of orbit of ISZ along elliptic orbit. Since the motion of ISZ along the conditional circular orbit is assumed to be uniform, it is possible to register

$$\vartheta_{ср}=\omega_{кв}(t-t_0). \qquad (2.1.12)$$

t_0 here corresponds to moment of time, at which ISZ passes perigee point.

Between mean anomaly and eccentric anomaly there is following relationship/ratio:

$$E=\vartheta_{ср}+e\sin\vartheta_{ср}+\frac{e^2}{2}\sin 2\vartheta_{ср}, \qquad (2.1.13)$$

Let us now move on to determination of velocity of motion of ISZ from elliptic orbit. From the formulas in [2.1] we obtain

$$v_з=v_{кв}\sqrt{\frac{1+2e\cos\vartheta+e^2}{1-e^2}}. \qquad (2.1.14)$$

Taking into account that for apogee of orbit $\cos\vartheta_А=-1$, and for perigee $\cos\vartheta_П=1$, let us find

$$\left.\begin{aligned}v_А&=v_{кв}\sqrt{\frac{1-e}{1+e}}\\ v_П&=v_{кв}\sqrt{\frac{1+e}{1-e}}\end{aligned}\right\}. \qquad (2.1.15)$$

Thus, in apogee satellite has minimum speed, and in perigee -

greatest. Consequently, during the use of ISZ as the repeater the velocity of the motion of the terrestrial antenna, which tracks for ISZ, will be smallest during the use of a section of orbit near the apogee. This is evident from the graphs Fig. 2.1.3, which are constructed according to expression (2.1.14) for values of e=0.1; 0.5, and 0.8. Let us note that values $v_{к_a}$ can be determined according to expression (2.1.3) during replacement of r on a.

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All given above formulas determine absolute value of satellite velocity in circular and elliptic orbits without taking into account motion of Earth. The satellite velocity relative to the surface of Earth $v_{сз}$ will be equal to the vector difference

$$\bar{v}_{сз} = \bar{v}' - \bar{v}_з,$$

here of $\bar{v}'$ - projection on the surface of the Earth the velocity vector of satellite, and $\bar{v}_з$ - vector of the linear speed of daily rotation of that point in the Earth, for which is projected/designed the satellite. Value $\bar{v}'$ is determined from formulas (2.1.2) and (2.1.14), and direction is located through the location of satellite in orbit and its inclination. Vector $\bar{v}_з$ is directed along the parallel of the Earth from the west to the east, and its value $\frac{2\pi R_з \cos\zeta}{T_з} = \frac{2\pi 6400 \cos\zeta}{24 \cdot 60 \cdot 60} \approx 1.4\cos\zeta$ km/s, moreover ζ is determined the latitude of point on the surface of the Earth, for which is projected/designed the satellite.

During calculations of motion of satellites according to elliptic orbits, is frequently necessary knowledge of distance r between satellite and center of Earth (radius-vector). Radius-vector is connected with the orbital parameters with following relationship/ratio [2.1]:

$$r = \frac{r_{\text{П}}(1+e)}{1+e\cos\vartheta} = \frac{r_{\text{A}}(1-e)}{1+e\cos\vartheta}. \qquad (2.1.16)$$

Time of motion of satellite of one point of orbit into another point can be found from expression

$$t_2 - t_1 = \frac{E_2 - E_1 - e(\sin E_2 - \sin E_1)}{6{,}32\cdot 10^5}\sqrt{a^3}. \qquad (2.1.17)$$

Circling time of satellite is determined by third law of Kepler. Using it, let us find the circling time of satellite along the elliptic orbit with the semimajor axis a

$$T_a = T_0\left(\frac{a}{R}\right)^{3/2}. \qquad (2.1.18)$$

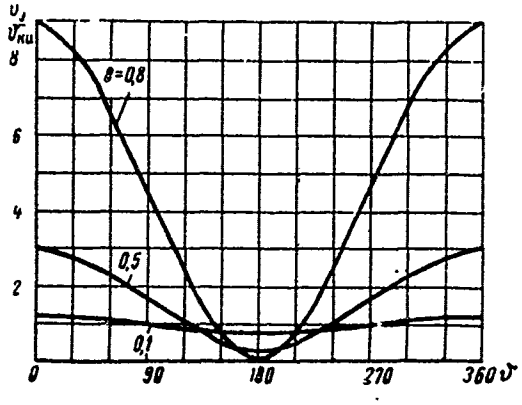

Fig. 2.1.3. Dependence of velocity with different value of eccentricity from bottom ϑ.

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Circling times of satellite $T_к$ and $T_э$ are called sidereal or star; they do not depend on motion of satellite relative to Earth and prove to be identical for satellites, which move at one height/altitude along polar and equatorial orbits. The circling time of satellite, calculated by observer, who is located on the Earth, will depend on the parameters of orbit, direction of the motion of satellite relative to the Earth and will not coincide with the sidereal period. This difference will be especially sharp for the satellites, which move along the equatorial orbits.

Circling time of satellite T Earth for ground-based observer can be found from expression

$$T = T_{сп} \frac{1}{\left|1 - \frac{T_{сп}}{24 \cdot 60} \cos i\right|}. \qquad (2.1.19)$$

Here i - dip angle, and $T_{сд}$ - circling time of satellite, expressed in minutes and determined by expression (2.1.6) or (2.1.18). Minus sign in this expression relates to the case, when the projection of the velocity vector of satellite on the equatorial plane coincides in the direction with the velocity vector of the rotation of the Earth; otherwise minus sign should be replaced by plus sign. During the motion of satellites along the circular equatorial orbits of radius r in the direction of rotation of the Earth we will obtain

$$T = T_0\left(\frac{r}{R}\right)^{3/2}\frac{1}{\left|1-T_0\left(\frac{r}{R}\right)^{3/2}\frac{1}{24\cdot 60}\right|} = 84{,}4\left(\frac{r}{R}\right)^{3/2}\frac{1}{\left|1-5{,}86\cdot 10^{-2}\left(\frac{r}{R}\right)^{3/2}\right|}. \quad (2.1.20)$$

Relationships/ratios between rounded by orbit altitude and value of sidereal and observed period of satellite, which moves along circular equatorial orbit in direction west-east (i.e. in direction of motion of Earth), designed according to expression (2.1.20), are given in Table 2.1.2.

As already mentioned, during motion of satellite along equatorial orbit with height/altitude of about 36000 km in direction of rotation of Earth, for observer in Earth satellite will seem fixed.

Table 2.1.2.

Высота орбиты км (1)	Период, часы (2)	
	сидерический (3) ($T_к$)	наблюдаемый (4) (T)
1 700	2,0	2,18
10 400	6,0	8,0
14 000	8,0	12,0
20 200	12	24
36 000	24,0	Бесконечность (5), так как спутник стационарный

Key: (1). Orbit altitude km. (2). Period, hours. (3). sidereal. (4). observed. (5). Infinity, since satellite, stationary.

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Comparison of starting/launching of satellites in elliptical and circular orbits shows that removal of satellite to elliptic orbits proves to be more simply and requires smaller expenditure of energy (fuel/propellant). Furthermore, with the identical circling times of satellite around the Earth in the case of motion along the circular and elliptic orbits the ratio of the time of the visibility of satellite from the Earth to the circling time of satellite can prove to be considerably greater during the motion of ISZ along the elliptic orbits. This is confirmed by Fig. 2.1.3. Actually, if we take r=a, then according to (2.1.6) and (2.1.18) we will obtain $T_э=T_к$. The curves of Fig. 2.1.3, show that with an increase in eccentricity e the satellite velocity in the field of view, which is chosen close to the apogee point (for example, in limits of

$120°<\vartheta<240°$), will be considerably less than the satellite velocity along the circular orbit with a radius of r=a (for circular orbit $e=0, \frac{v_э}{v_{кв}}=1$ and it does not depend on angle ϑ). Consequently, the section of orbit in the field of view will be travelled in the smaller interval of time τ in such a case, when satellite moves along the circular orbit. Since was accepted $T_э=T_к$ is obvious $\frac{\tau_к}{T_к}<\frac{\tau_э}{T_э}$. Thus, the duration of period of communication under these conditions will be greater in the case of applying the elliptic orbits.

During selection of orbit altitude and hum of orbit inclination one should take into account that on heights/altitudes approximately between 2000 km and 20000 km from surface of Earth in region of latitudes of ±60° is located radiation belt. During the motion of satellite in this belt/zone serious disturbances/breakdowns in the work of onboard radio-electronic equipment can occur, furthermore, during the motion in the region of radiation signal level on the line satellite - the Earth will sharply vary even during one revolution of satellite, which will require more complicated receiving and antenna systems on the Earth and on board ISZ.

Greater orbit altitudes have advantage with respect to scope/coverage by connection/communication of greater areas; quantity of satellites, required for realization of continuous connection/communication, in this case is decreased. However, an

increase in altitude of orbits causes the complication of ground-based and installed equipment. Consequently, the selection of orbit altitude must be produced taking into account many factors.

During subsequent analysis of communication systems through ISZ, besides previously enumerated parameters, which characterize position of orbit and motion of satellites relative to Earth, it is expedient to introduce into examination following (Fig. 2.1.4):

- ascending node of orbit (B = V) - point, at which orbit intersects equatorial plane upon transfer ISZ from southern hemisphere into north,

- descending unit of orbit (H = N) - point, at which orbit intersects equatorial plane upon transfer from northern hemisphere to south,

- nodal line (BH = VN) - line, which connects ascending and descending units,

- projection of orbit on surface of Earth - A'B'П'H' (A'B'P'N'). Page 24.

Let us note that orbit inclination i, determined by angle between equatorial plane and orbital plane, varies from zero to ±180°, positive values corresponding to ascending node, and negative - descending.

Parameters, which characterize motion of satellite, can be

changed under effect of different factors. The effect of the disturbances/perturbations, caused by the gravitational fields of the sun and Moon, first of all, is exhibited in change in altitude of perigee, which, depending on the location of orbit relative to the sun and the Moon, can be reduced or increase [2.1]. Finally, the disturbances/perturbations, caused by a difference in the form of the Earth from the form of sphere, change the orientation of orbit, leading to the continuous rotation of orbit around the rotational axis of the Earth (precession of the nodal line of orbit and the rotation of line of apsides). The latter causes the angular displacement of perigee point relative to the ascending node. The precession of the nodal line and perigee of the orbit conversely proportional to semimajor axis is proportional to eccentricity. It is interesting to note that the procession of nodal line is equal to zero for the orbit with inclination i=90°, and with i=63.5° zero rotation of line of apsides are equal to.

Transit of satellite relative to Earth is accepted also to characterize by projection of trajectory of motion of satellite on surface of Earth, i.e., by geocentric projection of trajectory. The position of geocentric projection the trajectories of the motion of satellite or geocentric projection of satellite orbit path are connected with the elements of orbit with the equality

$$\operatorname{tg}\zeta = \operatorname{tg} i \sin(\omega_3' t + \Delta\lambda), \qquad (2.1.21)$$

where ζ - the instantaneous value of the geocentric latitude of satellite, ω_s' - the total angular rate of rotation of the Earth and precession of orbit, t - time of flight of ISZ from the ascending node of orbit to the point in question, and $\Delta\lambda$ is determined by difference in longitude of the point of orbit and longitude/length of ascending node λ_B in question.

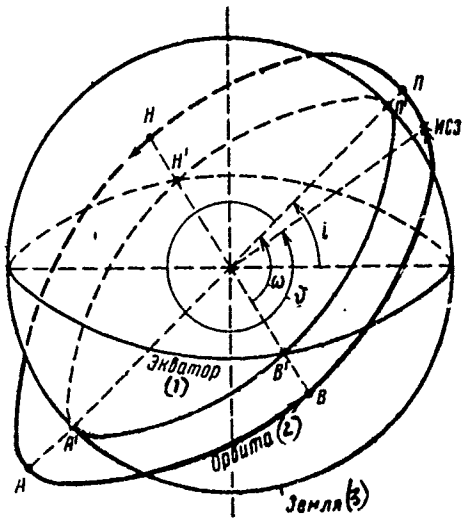

Fig. 2.1.4. Elements of elliptic orbit: A and П - apogee and perigee, A ' and П ' - projection A and П on surface of Earth; B and H - ascending and descending units of orbit; B ' and H ' - projection B and H on surface of Earth; ω - angular distance of perigee from ascending node ϑ - true anomaly, i - orbit inclination.

Key: (1). Equator. (2). Orbit. (3). Earth.

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From expression (2.1.21) it follows that geocentric projection of satellite orbit path with orbit inclination, equal to i, will contain region of latitudes from $\zeta=-i$ to $\zeta=+i$. Fig. 2.1.5 shows the geocentric projection of the satellite orbit path, which moves along the circular inclined orbit with the period of orbit $T_к=0{,}5$ h with value of i=30°. Let us note that in accordance with (2.1.21) the amount of the bias/displacement of the geocentric projection of each

subsequent satellite orbit path on the longitude/length, relative to the projection of the previous turn, comprises

$$[\omega_3' t + \Delta\lambda_n] - [\omega_3'(t-T) + \Delta\lambda_{n-1}],$$

consequently, $\Delta\lambda_n - \Delta\lambda_{n-1} = \omega_3' T$.

Here n - number of satellite orbit path, and T - orbital period. Designating $\Delta\lambda_n - \Delta\lambda_{n-1} = \delta$, we will obtain

$$\delta = \omega_3' T. \qquad (2.1.22)$$

From expression (2.1.22) it follows that with increase in circling time of satellite bias/displacement of projection of adjacent turns δ increases. In this case, taking into account the direction of rotation of the Earth, it is possible to say that the projection of each subsequent turn is displaced to the west. With $T = T_3 = 24$ h without taking into account the precession of orbit, $\delta = \omega_3 T_3 = \frac{2\pi}{T_3} T_3 = 2\pi$. During the prolonged motion of the satellite of the projection of many turns they will be a large number of the "sinusoidal" curves, similar to those shown in Fig. 2.1.5, shifted value δ. These curves with the long time of observations fill entire region of latitudes with boundaries of $\zeta = \pm i$.

Let us note that in the case of moving satellites in equatorial orbit (i=0) geocentric projections of all turns will coincide with equator of Earth.

During motion of satellites along polar orbits ($i=\pi/2$) of projection of turns they will contain entire region of latitudes from north to south pole. In this case, if the circling time of satellite to the period of the orbit of the Earth, projection of turns will multiple present closed curve. The latter is shown in Fig. 2.1.6 and 2.1.7 [2.10] for the satellites, which move along the polar circular orbits with the period of orbit 4 h and 12 h (orbit altitude with respect to 6360 km and 20200 km). In these figures, made in the form of polar coordinate grid of the Earth, the projections of satellite orbit paths are shown and their numbering is given.

Principle of construction of geocentric projections of satellite orbit paths is clarified in Appendix 1. Let us note, that during the motion of satellites along the orbits with this low altitude that the period of their orbit is considerably less than the period of the orbit of the Earth, the geocentric projections of the satellite orbit paths, which move along the circular and elliptic orbits with the identical inclination, will barely differ from each other.

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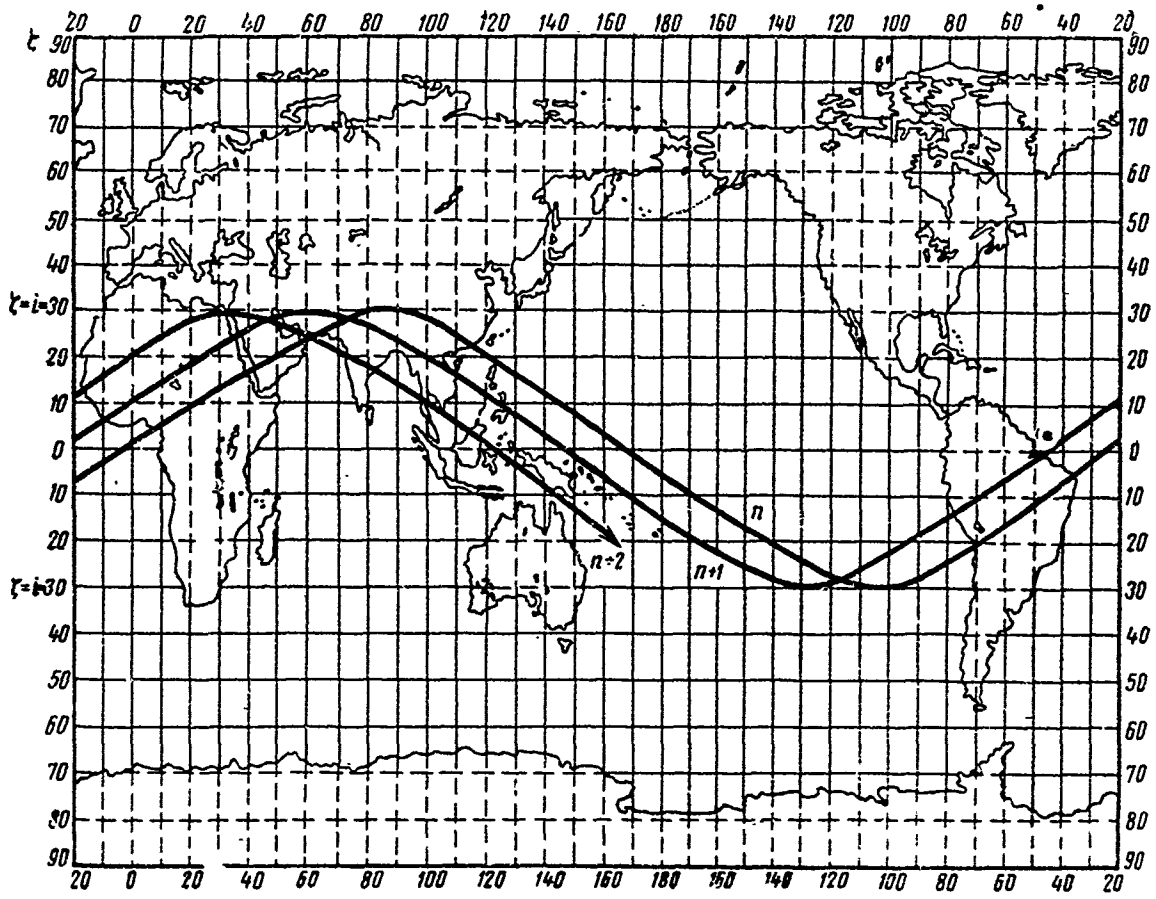

Fig. 2.1.5. Projection of passes of satellite with i=30°.

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During motion of satellite along orbits with high altitude, when circling time of satellite proves to be commensurate with period of orbit of Earth, geocentric projections of satellite orbit paths, which move along circular and elliptic orbits, will considerably differ from each other.

For simplification in further examination of special features of communication systems through ISZ it is expedient to introduce concept of orbital surface, by which we will understand surface, motion by which coincides with motion of Earth, formed by traces of turns of ISZ for long time. Thus, during the motion of satellite along the circular polar orbits with a height/altitude H orbital surface will be the sphere, whose radius is equal to a radius of orbit.

During motion of satellite along multiple-synchronous polar orbits orbital surface will be degenerated into group of lines, similar to those shown in Fig. 2.1.6 and 2.1.7. If satellite will move along the equatorial circular or elliptic orbits, orbital surface will degenerate respectively into the line, which is circle/circumference or ellipse, lying/horizontal at are equatorial the plane, and in the case of stationary satellite orbital surface is degenerated into the point, which is located in equatorial plane and coincides with the position of satellite in the space.

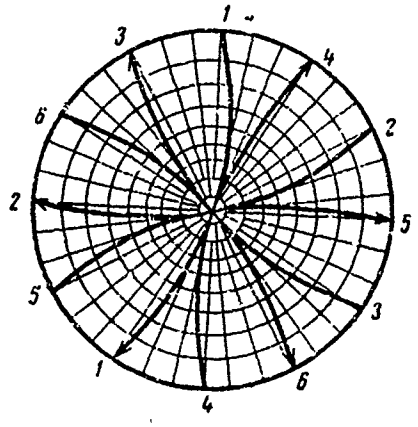

Fig. 2.1.6.

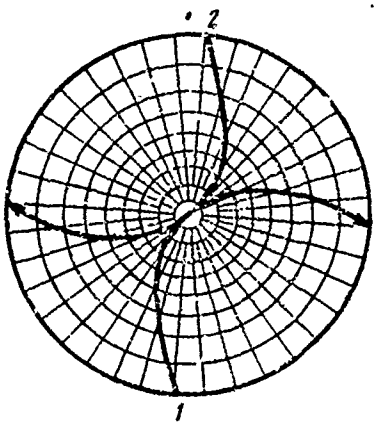

Fig. 2.1.7.

Fig. 2.1.6. Projection of satellite orbit paths with i=90° with period of 4 h.

Fig. 2.1.7. Projection of satellite orbit paths with i=90° with period of 12 h.

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During motion of satellite along circular orbits with height/altitude H and inclination, equal to i, orbital surface will be determined by part of spherical surface, limited by planes k'c" and c'k" (see Fig. 2.1.8), coordinate of latitude of which is equal to $\zeta_1=i°$ N and $\zeta_2=-i°$ S respectively.

During motion of satellite along elliptic orbit with inclination (without taking into account disturbances/perturbations) orbital surface will take form of ellipsoid, limited by planes, coordinate of latitude of which is equal to $\zeta_1=i°$ N and $\zeta_2=-i°$ S respectively.

2.2. Region of survey/coverage and the zone of coupling.

For determining of zone, in which is possible observation of satellite, and consequently also radio communication with it, let us consider Fig. 2.2.1, where three terrestrial stations А, Б and Д, which are located in plane of circular orbit of satellite with height/altitude H, are shown. We will consider that the observation of the satellite, which moves in direction $C_1C_2C_3$, from the stations А, Б and Д can be realized not from the lines of horizon, which for each point/item are shown by dotted line, but beginning from some angles of elevation (occasionally referred to as angles of elevation), equal to β.

Thus, region, which lies proto-in station in question, limited by lines of sight, drawn at angles β to line of horizon, can be named region of survey/coverage or radio field of view of this station. For A station the boundaries of the region of survey/coverage, which coincide with the plane of drawing, Fig. 2.2.1, will be lines AC_1 and AC_4.

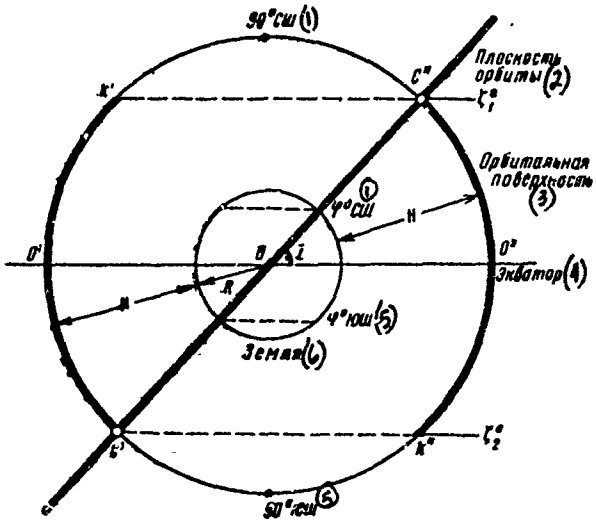

Fig. 2.1.8. Cross section of orbital surface by plane of inclined orbit.

Key: (1). NL. (2). Orbital plane. (3). Orbital surface. (4). Equator. (5). SL. (6). Earth.

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for the station Д of one of the boundaries of the region of survey/coverage the straight line ДС$_4$ is. Thus, the establishment of radio communication between points/items A and Д through the satellite possibly only then, when satellite will be located at point C$_4$, i.e., when it will prove to be simultaneously visible from point/item A and Д. Obvious also that the simultaneous observation of satellite from points/items A and Б, arranged/located on the closer distance than points/items A and Д, it is possible in such cases, when satellite will be located in section C$_2$C$_3$C$_4$ of orbit. The

boundaries of this section of orbit are determined by the mutual overlap of the regions of the survey/coverage of points/items A and Б. Therefore it is possible to say that section $C_2C_3C_4$ of orbit is the zone of the simultaneous connection/communication of satellite with points/items A and Б or the zone of the coupling of points/items A and Б through the satellite. In accordance with this the zone of the coupling of points/items A and Д will be determined by the point of contact of tangency C_4 of the regions of the survey/coverage of these points/items. Let us note that in the case of moving the satellite along the orbit with a height/altitude less H, the connection/communication between points/items A and Д to scale of single time will generally prove to be impossible. Thus, for points/items A and Д orbit altitude H is the minimum height/altitude, at which is still possible setting connection/communication.

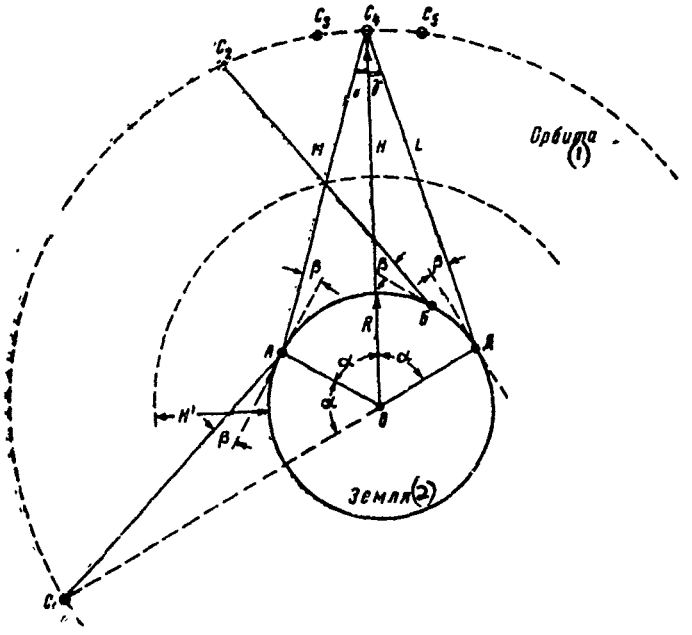

Fig. 2.2.1. To determination of visibility range of satellite.

Key: (1). Orbit. (2). Earth.

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For points/items A and Б the minimum orbit altitude of satellite H ' is determined by point M, at which intersect the boundary lines AC_4 and $БC_2$. Let us find the geometric relationships/ratios, which characterize the communication system with the minimum value of orbit altitude H. From examination ΔOAC_4, Fig. 2.2.1, it follows:

$$\gamma = \arcsin \frac{R \cos \beta}{R+H}, \text{ rad}, \qquad (2.2.1)$$

$$\alpha = \frac{\pi}{2} - (\gamma + \beta), \text{ rad}, \qquad (2.2.2)$$

$$l = \cup A\Pi = 2\alpha R, \ км, \tag{2.2.3}$$

$$L = AC_{\text{д}} = \sin\alpha \frac{R}{\sin\gamma} = \frac{R+H}{\cos\beta}\sin\alpha, \ км. \tag{2.2.4}$$

Here R≅6400 km - radius of Earth.

One should note that angle 2α determines angular dimension of region of survey/coverage of station, and value l is equal to maximum distance between stations, with which is possible setting connection/communication during motion of satellite along circular orbit with height/altitude H. The greatest distance between the station and the satellite is determined by value L, and angle 2γ characterizes the sizes/dimensions of the region of survey/coverage from the satellite during its motion at the height/altitude H.

In order to represent order of magnitudes, determined by resulting expressions, in Fig. 2.2.2 is constructed dependence L, l and 2γ on height/altitude of circular orbit H with value of angle of elevation β=0.

We use obtained results for case of stationary satellites. In this case, after taking β=5°, from expression (2.2.1) let us find that 2γ=17°. This means that in the strictly positive seat of the antenna of satellite in the space the expansion of the antenna radiation pattern is more than value of 2γ=17° inexpediently, since

it will not lead to an increase in the region of survey/coverage from the satellite. Under the same conditions from (2.2.2) it is possible to determine that $\alpha=90°-(8.5°+5°)=76.5°$. Consequently, in the communication system with the stationary satellite at the angle of elevation of $\beta=5°$ setting connection/communication with any points/items, which lie higher than 76.5° N or S proves to be impossible. For the realization of connection/communication with the points/items, which are located near the poles of the Earth, i.e., lying it is higher than the latitudes indicated, should be chosen inclined or polar orbits.

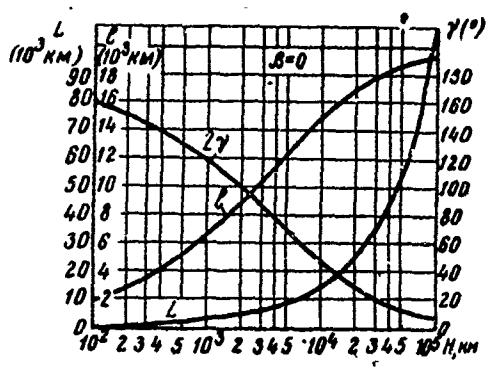

Fig. 2.2.2. Dependence of parameters of line of communications on orbit altitude.

Fig. 2.2.3 shows sizes/dimensions of regions of survey/coverage of satellites, which move along equatorial orbits with different height/altitude, for angles of elevation, equal to 0 and by 10°. Fig. 2.2.4 shows a change in the regions of the survey/coverage of stationary satellites at different angles of elevation.

Passing to more detailed examination of zone of mutual visibility, let us turn to Fig. 2.2.5, on which are depicted terrestrial stations A and Б and spherical orbital surface, which corresponds to motion of satellites along circular polar orbits.

We will consider that in antennas of stations A and Б very narrow radiation patterns, and antennas can be directed to all points of spaces, angles of elevation of which exceed certain angle β. Therefore the antenna A station can be directed to any point of zone A_1 of orbital surface, and the antenna of station Б - at any point of zone $Б_1$.

Thus, zone A_1 and $Б_1$ - region of survey/coverage for stations A and Б. Let us agree to call the zone, common to for several regions of survey/coverage, the zone of overlap; Fig. 2.2.5 shows it by dense shading. That part of the zone of the overlap of stations, through which is passed the satellite, can be the zone of coupling. As it follows from the following, the zone of overlap and the zone of coupling (or respectively their geocentric projections) sometimes

substantially are distinguished both by the area and by the layout.

Sizes/dimensions of zone of overlap are determined by geographical reference of stations A and Б, and also by size./dimensions of regions of survey/coverage, i.e., by boundaries, in limits of which is possible and is expedient displacement/movement of ground-based antennas. During the motion along the polar orbits the satellite can fall into any point of spherical surface; therefore in this case the zone of overlap is simultaneously and the zone of coupling. During the motion of satellite along the orbit with inclination i the zone of coupling can prove to be the part of the zone of overlap. Taking into account the aforesaid in the examination of Fig. 2.1.8, we will obtain that with orbit inclination, equal i, the zone of the coupling of stations A and Б will be determined by the figure Fig. 2.2.5 in the form of spherical triangle abc, whose sides ac and bc define the boundaries themselves of the regions of the survey/coverage of stations A and Б, and side ab - parallel ζ=+i.

Definition of zone of coupling examined is correct with undirected receiving and transmitting antennas of satellite or with very wide radiation patterns. In the case, when the angle Θ, which characterizes the width of the major lobe/lug of the radiation patterns of the onboard antennas, proves to be close in the value to angles $2\gamma_1, 2\gamma_2, \ldots, 2\gamma_n$ (at which terrestrial stations "are visible" from the satellite, which is located at different points on the boundaries of the zone of overlap, see Fig. 2.2.1), for the definition of the

zone of coupling it is necessary to consider, besides θ, and the angle, which characterizes error in the orientation of onboard antennas relative to terrestrial stations.

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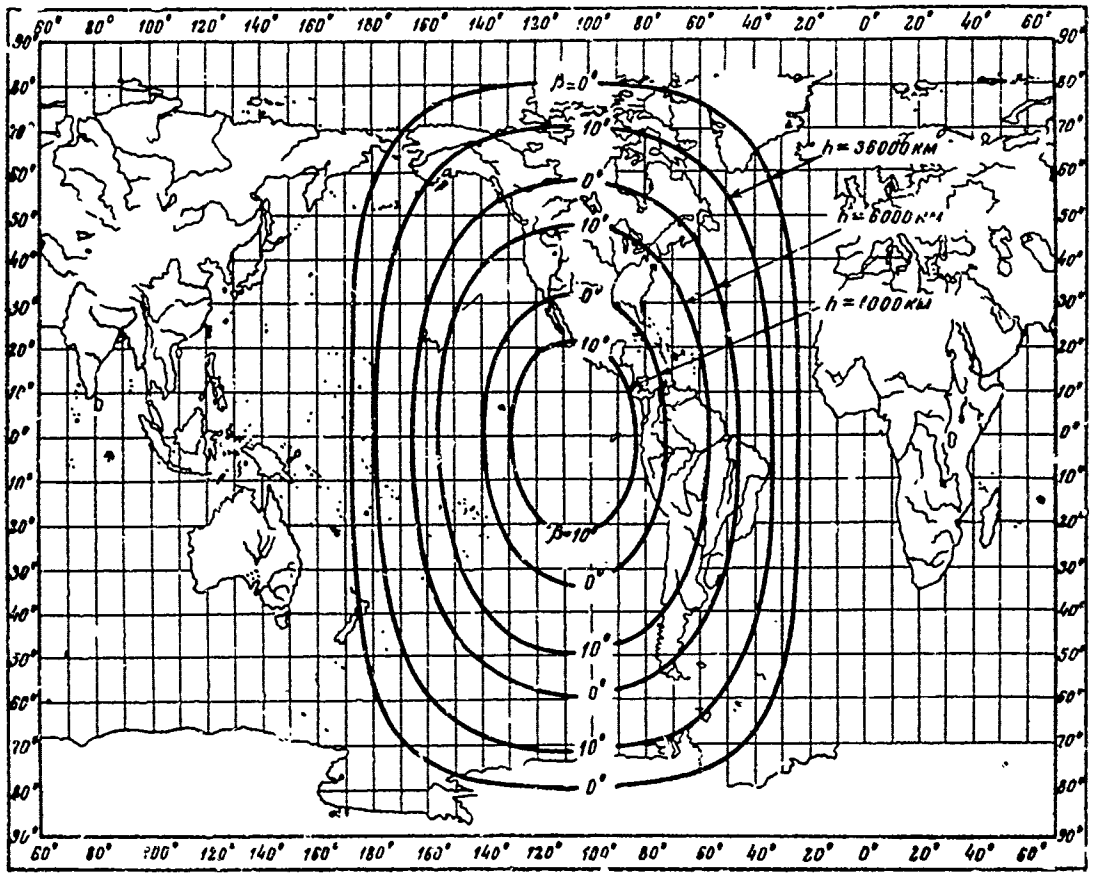

Fig. 2.2.3. Change of region of survey/coverage in dependence on height/altitude of satellite [3.3].

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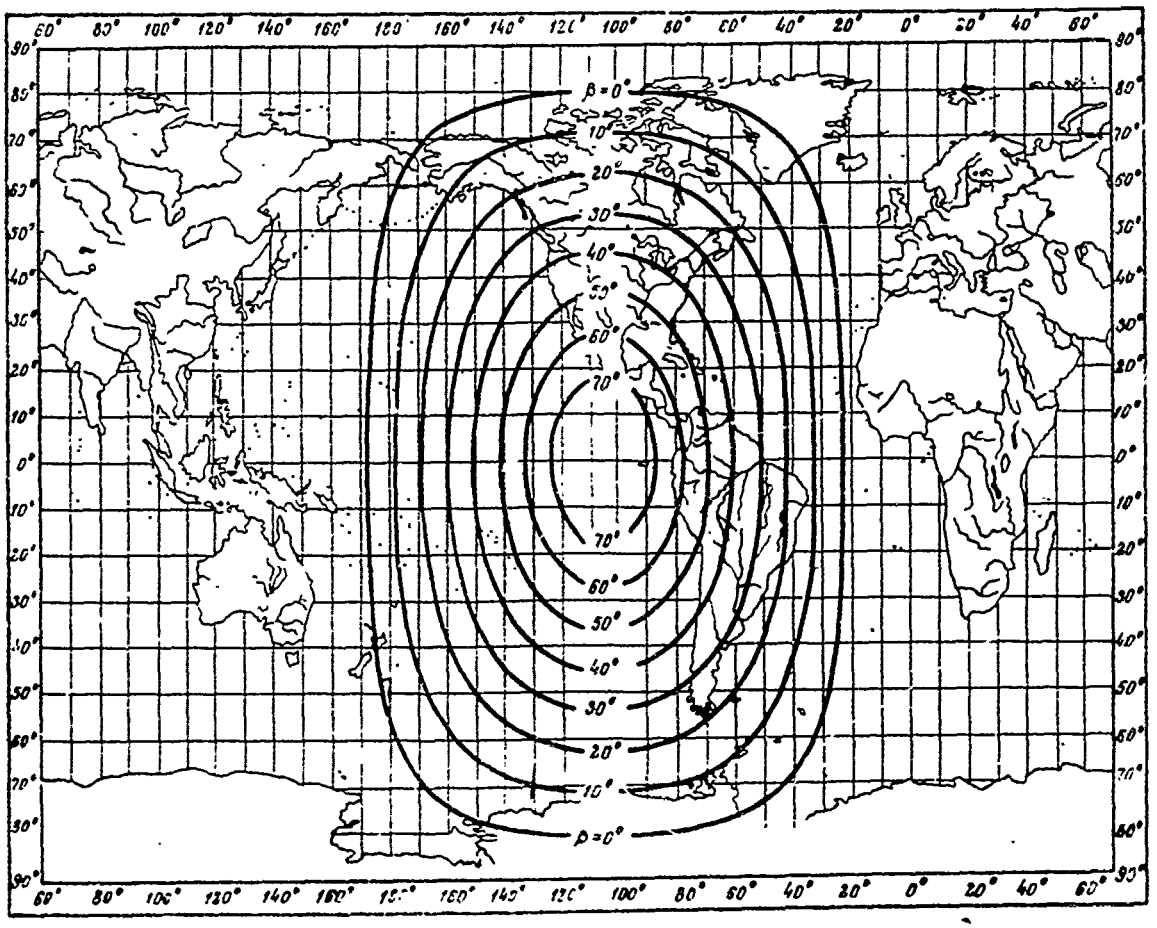

Fig. 2.2.4. Map/chart of zone, operated by stationary satellite, at different angles of elevation [3.3].

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It is obvious that in these cases the zone of coupling decreases in the sizes/dimensions and it will differ significantly from the zone of overlap.

With following presentation we will count width of radiation pattern of on-board antennas to be so large that taking into account errors in orientation it exceeds values of angles $2\gamma_1$, $2\gamma_2$, ..., $2\gamma_n$ on

boundaries of the region of survey/coverage. This means that the antenna of satellite contains terrestrial stations of all points of that part of the zone of overlap, in which can be located the satellite.

Determination of analytical expression, which determines coordinates of zone of coupling or by its geocentric of projections, sufficiently labor-consuming problem. Expressions obtained in this case are bulky and inconvenient for the practical use.

For determining geocentric projection of zone of coupling it is most expedient to use the graphic method. Let us consider the order of construction. First on the globe or the map/chart around each station the geocentric projections of the regions of survey/coverage will be plotted. The latter are circles with radius r_0, which can be found in accordance with Fig. 2.2.1 from the expression

$$r_0 = \alpha R.$$

Substituting expressions (2.2.2) and (2.2.1) we will obtain

$$r_0 = R\left[\frac{\pi}{2} - \left(\arcsin\frac{R\cos\beta}{R+H} + \beta\right)\right]. \qquad (2.2.5)$$

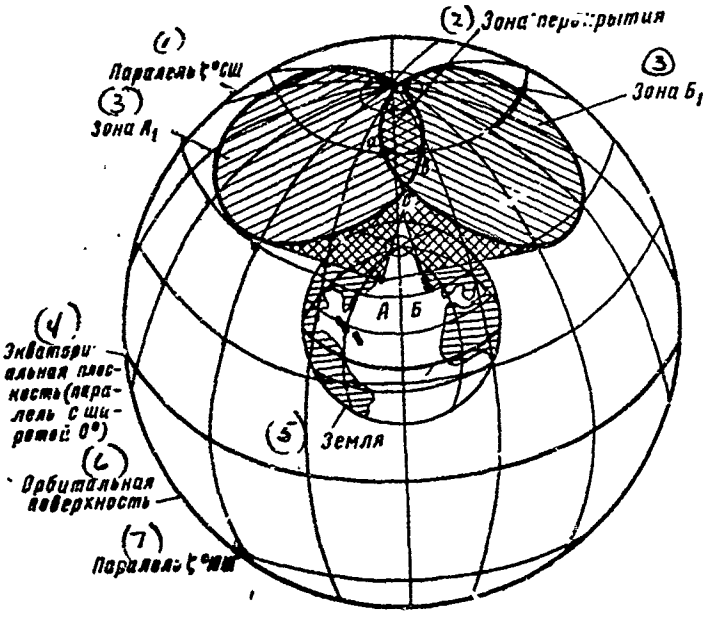

Fig. 2.2.5. Region of mutual visibility.

Key: (1). Parallel of ζ° NL. (2). Zone of overlap. (3). Zone. (4). Equatorial plane (parallel with latitude of 0°). (5). Earth. (6). Orbital surface. (7). Parallel with ζ° SL.

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Let us note that during plotting of projections of regions of survey/coverage on geographical maps/charts, it is necessary to consider scale change in different sections of map/chart, caused by special features of projection, accepted during its compilation. After the projections of the regions of survey/coverage are replaced, on the figure, formed by the mutual overlap of several zones of survey/coverage, are noted two parallels, whose latitude is equal to the dip angle of orbit. The region, which lies within the projection of the region of the overlap and the parallels indicated, will be the geocentric projection of the zone of coupling.

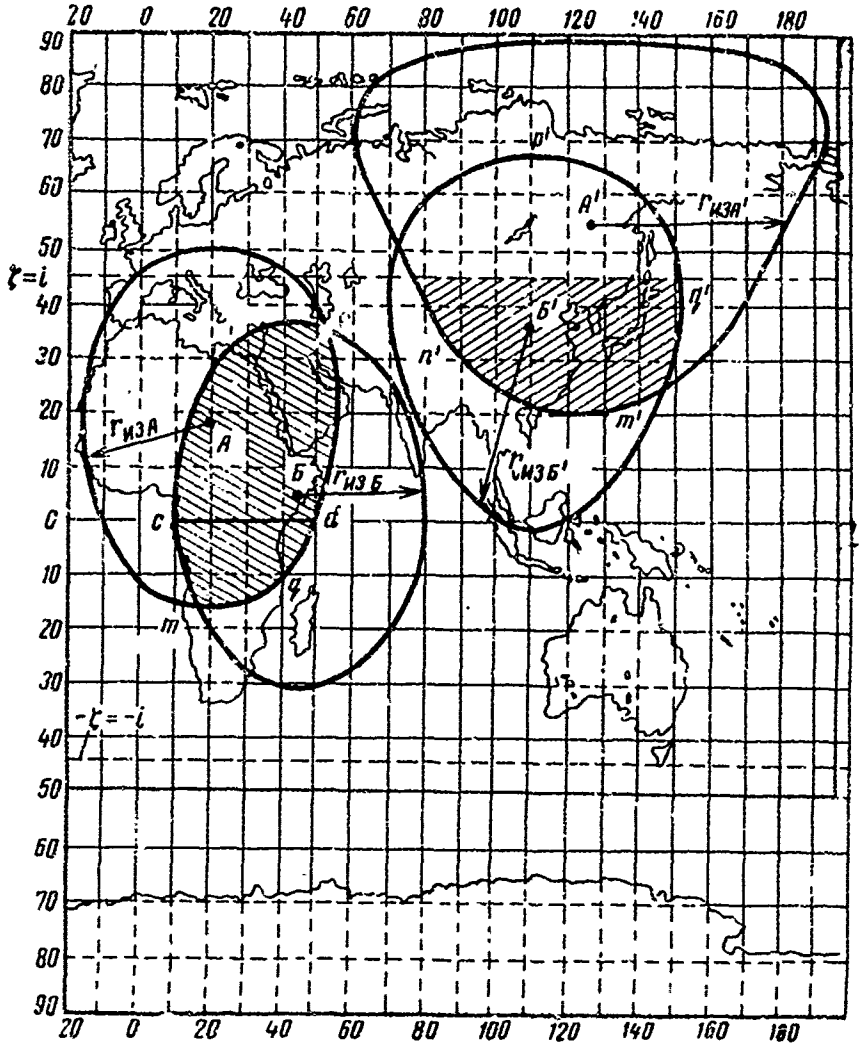

Fig. 2.2.6. Geocentric projections of zones of coupling.

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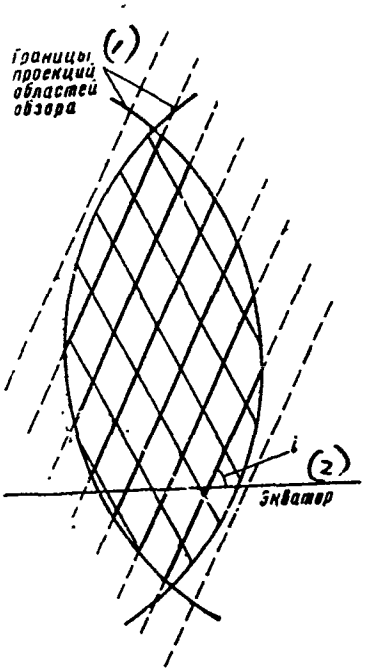
As example Fig. 2.2.6 shows geocentric projections of regions of survey/coverage for stations A and Б and A ' and Б'. For the satellite, which moves along the polar orbit by the geocentric projections of the zone of coupling, there will be regions mcnpdq and m'n'p'q'; for the satellite, which moves along the orbit with

inclination i, shaded regions.

During motion of satellite in equatorial orbit geocentric projection of any satellite orbit path will coincide with equator. Hence follows that if the geocentric projection of the zone of overlap does not contain equator, the connection/communication of the stations through a satellite which moves along equatorial orbits, it is impossible at any orbit altitude. To this case corresponds an example A' stations and Б', Fig. 2.2.6.

On geocentric projections of zones of mutual visibility they can be replaced to projection of turns, along which moves satellite. Taking into account expression (2.1.22) and Fig. 2.1.5, it is possible to say that the projections of turns in the geocentric projection of the zone of coupling take the form Fig. 2.2.7 and 2.2.8. Let us note that if the circling time of satellite is not multiple to the period of the orbit of the Earth or if satellite moves not along the equatorial orbits, then with the sufficiently large duration of the time of the observation of the projection of the ascending sections of turns (or the analogously descending turns) they will completely fill the zone of coupling.



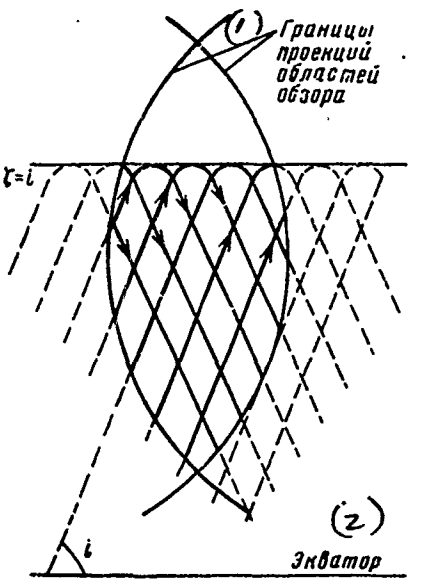


Fig. 2.2.7. Fig. 2.2.8.

Fig. 2.2.7. Geocentric projections of satellite orbit paths with orbit with inclination.

Key: (1). Boundaries of the projections of the regions of survey/coverage. (2). Equator.

Fig. 2.2.8. Geocentric projections of satellite orbit paths with orbit with inclination.

Key: (1). Boundaries of the projections of the regions of survey/coverage. (2). Equator.

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2.3. Duration and the probability of periods of communication.

Duration of period of communication $\Delta t_{св}$ between two terrestrial stations is determined time interval, during which satellite passes zone of coupling. During the motion of satellite along the circular orbit the speed of its motion is constant, consequently, the duration of period of communication during one circling time of satellite will be determined by the extent of that part of the turn, which is located in the zone of coupling.

Taking into account Fig. 2.2.7 and 2.2.8, it is possible to say that for each circling time of satellite along inclined or polar orbit extent of that part of turn, which falls into zone of mutual visibility, will different and vary from zero (when turn does not pass through zone of coupling) to certain maximum value $l_{с\ макс}$. In accordance with this the duration of period of communication will also vary from zero to the certain maximum value

$$\Delta t_{св\ макс} = \frac{l_{с\ макс}}{v}, \qquad (2.3.1)$$

here $v_{ср}$ - the satellite velocity. In the case of the circular orbit

$$\Delta t_{св\ макс} = \frac{l_{с\ макс}}{v_к} = \frac{l'_{с\ макс}}{v_к}\frac{R+H}{R}. \qquad (2.3.2)$$

Here $l'_{с\ макс}$ - extent of geocentric projection of turn. During the motion of satellite along the equatorial orbit, and also along a repeatedly synchronous inclined or polar orbit, the trajectory of the motion of satellite in the space they will coincide. Therefore value $\Delta t_{св}$ will prove to be identical for each period of orbit (for each turn).

Duration of period of communication $\Delta t_{св}$ on any turn can be determined according to expression

$$\Delta t_{св} = \frac{l_c}{v} = \frac{l'_c}{v}\frac{R+H}{R}, \qquad (2.3.3)$$

where l_c or l'_c they find graphically through extent of part of turn in zone of coupling or its geocentric projection in projection of this zone.

During motion of satellite along circular orbit $v=v_к$ it is constant. During the motion of satellite along the elliptic orbit its speed v is changed; therefore its value in the zone of coupling will depend on the position of this zone relative to perigee.

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Together with parameter $\Delta t_{св\,макс}$ important characteristics of system of connection/communication through ISZ - average/mean value of probability of connection/communication at use of one satellite in prolonged period of observation $T_н$. The value indicated can be found from the expression

$$p(1) = \frac{t_{св\,\Sigma}}{T_н}. \qquad (2.3.4)$$

Here $t_{св\,\Sigma}$ - indicates total time of duration of all periods of communication for time $T_н$, which considerably exceeds circling time of satellite.

For determination of $t_{св}$ we will use Fig. 2.3.1. Are here shown turns, on which the satellite moves in the zone of coupling. With the long time of observation the turns will evenly cover/coat the zone of coupling, moreover the distance between the nearest turns, measured along the parallel, is equal to δ. Let us isolate in the zone of coupling along one of the parallels with latitude ζ_a the elementary strip with an extent of $\Delta\varphi(\zeta_a)$ and angular dimension $d\alpha$. Then in the limits of this strip the extent of the part of turn $l=(R+H)d\alpha$, and number of turns, which evenly cover/coat this strip $n=\frac{\Delta\varphi(\zeta_a)}{\delta}$. Consequently, the time, during which the satellite is within the limits of the chosen strip,

$$\Delta t_{св}=\frac{ln}{v}=n\frac{R+H}{v}d\alpha=\frac{\Delta\varphi(\zeta_a)}{\delta}\cdot\frac{(R+H)d\alpha}{v}. \qquad (2.3.5)$$

In limit, with very long time of observation, turns will fill entire zone of coupling.

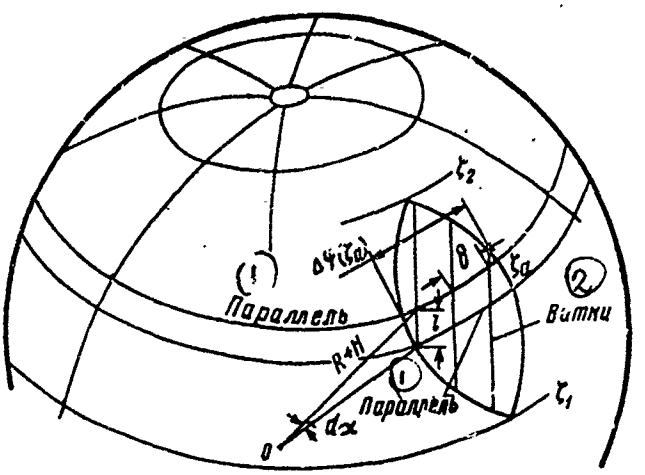

Fig. 2.3.1. Determination of probability of coupli..g.

Key: (1). Parallel. (2). Turns.

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Consequently, the total time, during which the satellite will pass this zone, being located only on those ascending or only in the descending sections of turns, will be determined by the area of zone, i.e.

$$\Delta t_{\text{сов}} = \int_{\alpha(\zeta_1)}^{\alpha(\zeta_2)} \frac{\Delta\varphi(\zeta)}{\delta} \frac{R+H}{v} d\alpha. \qquad (2.3.6)$$

Here $\Delta\varphi(\zeta)$ - function, which determines change in extent of zone of mutual visibility along parallels, which are located on latitudes with coordinates ζ. Let us note that values α and ζ are connected with a certain function, examined in Appendix 2. Integration limits must be undertaken on the angle α, which corresponds to a change in the latitude from ζ_1 to ζ_2.

If we consider only ascending (or only descending) sections of projections of turns, it is possible to draw conclusion that they will cross half of equator. If angular distance between adjacent turns is equal δ_1, then number of intersections of equator with the projections of the ascending (or descending) sections of turns will comprise $n_1=\pi/\delta_1$. We will consider that the duration of one period of revolution of satellite is equal to T, then the time of the observation

$$T_н=n_1T=\frac{\pi}{\delta_1}T. \qquad (2.3.7)$$

Substituting latter/last two expressions in (2.3.4), let us find

$$p(1)=\frac{\int_{\alpha(\zeta_1)}^{\alpha(\zeta_2)}\frac{R+H}{v}\Delta\varphi(\zeta)\,d\alpha}{\pi T}. \qquad (2.3.8)$$

Expression (2.3.8) is correct for orbits of different form. In the case of the elliptic orbits of value v and H it follows to consider it different in the different sections of the zone of coupling.

For circular orbits periods of orbit $T=T_к$ and values H and $v=v_к$ are constant and therefore they can be carried out as integral sign. Taking into account expression (2.1.4), and also that r=R+H, for the circular orbit we will obtain

$$p_к(1)=\frac{1}{2\pi^2}\int_{\alpha(\zeta_1)}^{\alpha(\zeta_2)}\Delta\varphi(\zeta)\,d\alpha. \qquad (2.3.9)$$

After using results of appendix 2, let us replace angular dimension of dα with angular dimension, calculated along meridian dζ. After the substitution of expression (П.2.7) instead of (2.3.9) we will obtain

$$p_{\kappa}(1)=\frac{1}{2\pi^2}\int_{\zeta_1}^{\zeta_2}\frac{\Delta\varphi(\zeta)\cos(\zeta)\,d\zeta}{\sqrt{\sin^2 i-\sin^2\zeta}}. \qquad (2.3.10)$$

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Here ζ_1 and ζ_2 - in accordance with Fig. 2.3.1 - limiting values of latitudes on edges of zone of coupling or its geocentric projection, and i - orbit inclination of satellite. If satellite moves along the equatorial orbit i=0, value $\Delta\varphi(\zeta)=\Delta\varphi_0$ is determined by the segment of equator in the zone of coupling. In this case we will obtain

$$p_{\kappa}(1)=\frac{\Delta\varphi_0}{2\pi}. \qquad (2.3.11)$$

Let us note that with $\zeta_1=-i$ and $\zeta_2=i$, expression (2.3.10) leads to uncertainty/indeterminacy.

Calculation of integral (2.3.10) in first case can be carried out in accordance with application/Appendix 2 on formula

$$p_{\kappa}(1)=\frac{1}{2\pi^2}\int_{\zeta_1}^{+i-2{,}5^{\circ}}\frac{\Delta\varphi(\zeta)\cos\zeta\,d\zeta}{\sqrt{\sin^2 i-\sin^2\zeta}}+A\Delta\varphi(i)-B\,\Delta\varphi(i-2{,}5^{\circ}), \qquad (2.3.12)$$

where A and B coefficients, which depend on dip angle i and determined according to graphs Fig. 2.3.2.

One should note that value p(1) can be defined not only as average/mean value of probability of connection/communication during wide interval of time of observations, but also as ratio of average/mean value of time of connection/communication to circling time of satellite around Earth. The latter follows from the substitution of expression (2.3.7) in (2.3.4); actually, with this substitution

$$p(1) = \frac{t_{\text{св }\Sigma}}{n_1 T} = \frac{t_{\text{св ср}}}{T}. \qquad (2.3.13)$$

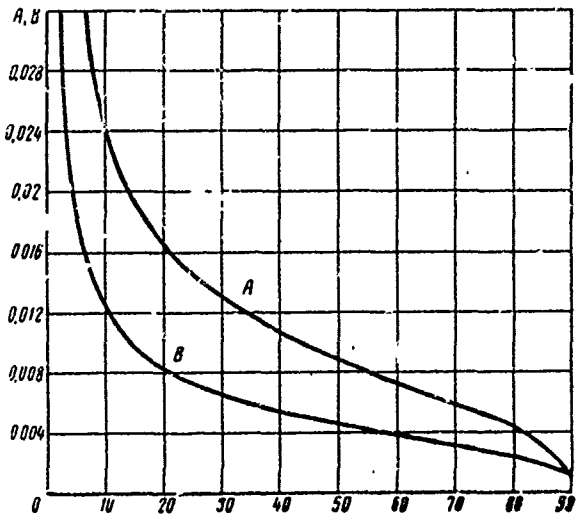

Fig. 2.3.2. Values of coefficients of A and B from orbit inclination.

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Furthermore, taking into account that in circular orbits the satellite velocity is constant, after the multiplication of numerator and denominator of latter/last expression on $v_к$ we obtain

$$p(1) = \frac{l_{св\,ср} v_к}{T_к v_к} = \frac{l_{св\,ср}}{l}, \qquad (2.3.14)$$

where $l_{св\,ср}$ - the mean pathlength of satellite in the zone of coupling, and $l = T_к v_к = 2\pi(R+H)$ - extent of the circular orbit of satellite. Thus, on the basis (2.3.14) value p(1) characterizes the ratio of the average/mean path, passed by satellite in the zone of coupling, to the extent of orbit.

Consider[illegible] number of works is devoted to determination of value $p_к(1)$. Thus, for example, in article [2.6] is given the following expression for determination $p_к(1)$ during the motion of satellites from

the polar orbits of radius r=R+H, when $\beta=0$:

$$p_{\kappa}(1) = \frac{1}{\pi}\left[\frac{\pi}{2} - \arcsin\alpha\right]; \qquad (2.3.15)$$

$$\alpha = \frac{R\sec\theta\sec^2\varphi + \sqrt{(r\operatorname{tg}\theta)^2(\operatorname{tg}^2\theta+\sec^2\varphi) - (R\sec\theta\operatorname{tg}\theta\sec\varphi)^2}}{(\sec^2\varphi+\operatorname{tg}^2\theta)r}.$$

All values entering this expression, are shown in Fig. 2.3.3, moreover 2θ is determined angle, at which intersect planes, tangents to surface of Earth in locations of stations A and Б.

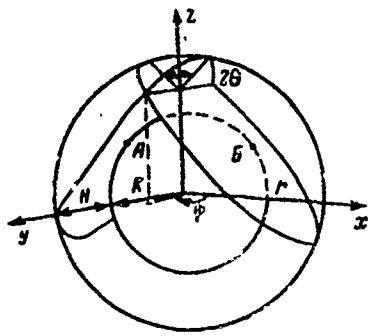

Fig. 2.3.3.

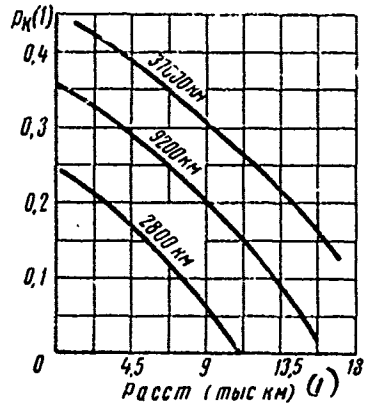

Fig. 2.3.4.

Fig. 2.3.3. To determination of region of mutual visibility.

Fig. 2.3.4. Dependence $p_к(l)$ on distance between terrestrial points. Key: (1) Distance (thousands of km).

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Solution of this expression for orbits height/altitude of 2800 km, 9200 km and 37000 km is given in Fig. 2.3.4 as dependence $p_к(l)$ on distance between points/items A and Б.

Character of change in value $p_к(l)$ at different angles of elevation is visible based on example, line of communications Maine-London (Fig. 2.3.5 [2.7]) with extent about 6000 km, passing almost along 50° N. This calculation is carried out for the satellite, which moves along the polar circular orbit with a height/altitude of 11000 km.

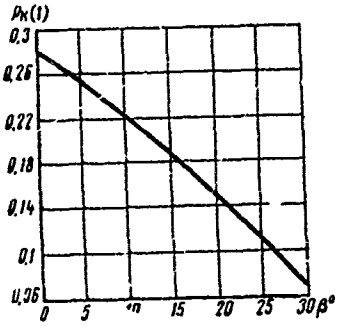

Fig. 2.3.5.

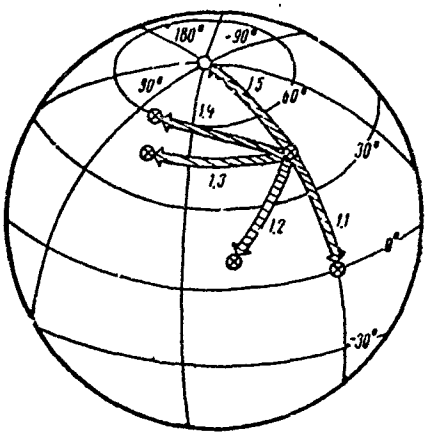

Fig. 2.3.6.

Fig. 2.3.5. Dependence $P_K(1)$ on angle of elevation: orbit polar H=11000 km, route - almost parallel to 50° N, extent of approximately 6000 km.

Fig. 2.3.6. To determination of optimum inclination.

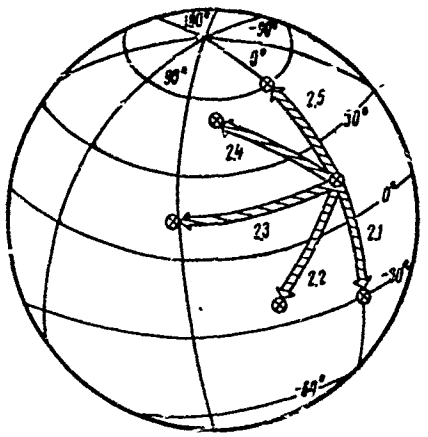

Fig. 2.3.7.

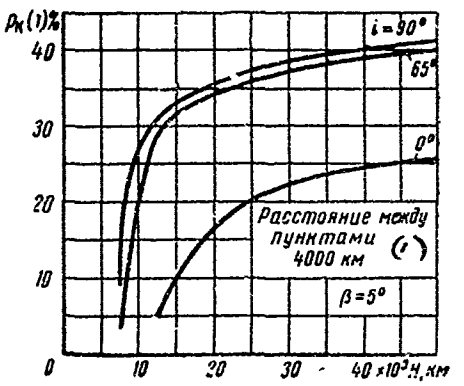

Fig. 2.3.8.

Fig. 2.3.7. Determination of optimum inclination.

Fig. 2.3.8. Value $p_к$ (l) with distance between points/items 4000 km.
Key: (1). Distance between the points/items 4000 km.

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Very careful calculations $p_к(l)$ for circular orbits with height/altitude of 1800-9000 km in the case of routes of different extent and differently arranged/located relative to orbital plane are carried out in works [2.8, 2.3]. The results of work [2.8] are given in Table 2.3.1 for different location of route and orbit inclination. During calculations of Table 2.3.1 orbit altitude is considered equal to 9000 km, the distance between the message centers over the surface of the Earth - 5000 km, the angle of elevation of $\beta=5°$. Line the connections/communications of different direction, given in Table 2.3.1 under numbers 1.1-1.5, are clarified by Fig. 2.3.6, and the lines of communications 2.1-2.5 - Fig. 2.3.7.

Fig. 2.3.8 and 2.3.9 give results of calculating dependence $p_{\text{к}}(I)$ on height/altitude of circular orbits with inclination 0, 65 and 90° for case, when both points/items of route are arranged/located on 56° N and at a distance of 4000 km (Fig. 2.3.8) and 8000 km (Fig. 2.3.9) from each other. On the basis of Table 2.3.1 and Fig. 2.3.8 it is possible to draw the general/common conclusion: the optimum value of orbit inclination is determined by the location of route relative to the zone of overlap. For those stations, the zones of overlap of which are located predominantly in equatorial regions, optimum orbit will have an inclination, equal to zero, when the zone of coupling is distant from the equator, optimum orbit it will be polar.

Table 2.3.1.

(1) № линии	(2) Угол между линией связи и экватором	i	$p_к$ (1)
1.1	—90	0	0,24
		45	0,16
		90	0,10
1.2	—60	0	0,20
		45	0,15
		90	0,10
1.3	0	0	0,06
		45	0,11
		90	0,20
1.4	60	0	0
		45	0,10
		90	0,21
1.5	90	0	0
		45	0,11
		90	0,22
2.1	—90	0	0,3
		45	0,14
		90	0,10
2.2	—60	0	0,25
		45	0,15
		90	0,09
2.3	0	0	0,19
		45	0,15
		90	0,10
2.4	60	0	0,16
		45	0,14
		90	0,11
2.5	90	0	0
		45	0,15
		90	0,11

Key: (1). Line. (2). Angle between line of communications and equator.

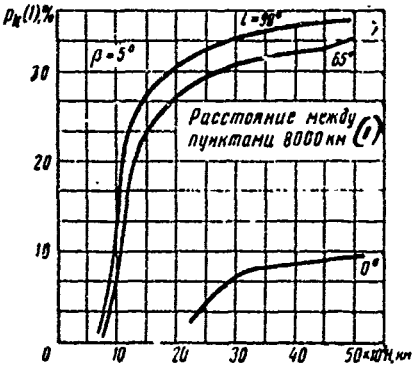

Fig. 2.3.9. Value $p_н(l)$ with distance between points/items 8000 km.
Key: (1). Distance between the points/items 8000 km.

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Now let us determine number of satellites n, necessary for providing connection/communication during assigned part of m of total time of observations $T_н$. In this case we will consider that the total time is sufficiently great ($T_н \gg T$, where T - circling time of satellite), and satellites are distributed in the space according to the random law. It is obvious that in the zone of overlap at some moments of time simultaneously can be observed one or several satellites, and at other moments of time - one.

Since p(1) is determined average/mean value of probability of determination of one satellite in zone of overlap for time $T_н$, it is possible to say that average/mean value of probability of fact that in zone of overlap for the same time it will be not one satellite, is equal

$$q(1) = 1 - p(1). \qquad (2.3.16)$$

Taking into account random distribution in space of n satellites, and also independence of their appearance in region of overlap, let us find probability of fact that none of these satellites is located in region of overlap. This probability

$$q(n) = [1 - p(1)]^n . \qquad (2.3.17)$$

It is obvious that probability q(n) will determine that part of time $T_н$ during which connection/communication between points/items will be absent, i.e.

$$1 - m = q(n) = [1 - p(1)]^n . \qquad (2.3.18)$$

Hence let us find necessary number of satellites, with which connection/communication will be provided during assigned part of m of total time

$$n = \frac{\lg(1-m)}{\lg[1-p(1)]} . \qquad (2.3.19)$$

Fig. 2.3.10 gives graphic dependences, calculated according to expression (2.3.19).

From expression (2.3.18) it follows also that part of time, during which connection/communication is ensured, at least, through one satellite,

$$m_1 = 1 - [1 - p(1)]^n . \qquad (2.3.20)$$

Determination of dependence of value m with increase in number of

satellites is of interest. This dependence, calculated according to expression (2.3.20) when for a certain value of n value $m_n=0{,}9$, is given in the form of Table 2.3.2.

Let us determine now probability that part of time, during which in region of overlap is more than k will be simultaneously found from n satellites, if probability of determination in this region of one satellite (any of number n) is known and equal to $p(1)=p_1$.

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For solving this question should be used theorem about repetition of experiments [2.9]. From it it follows that the probability of appearance in the region of overlap d from n satellites

$$p_d = C_n^d p_1^d (1-p_1)^{n-d}. \qquad (2.3.21)$$

Here $C_n^d = \frac{n!}{d!\,(n-d)!}$ is determined number of combinations of n elements on d, moreover $C_n^1 = n$, and $C_n^n = C_n^0 = 1$.

Consequently, probability of appearance in region of overlapping number of satellites is less than value k

$$p(d<k) = p_{k-1} + p_{k-2} + \ldots + p_1 = \sum_{l=1}^{l=k-1} C_n^l p_1^l (1-p_1)^{n-l}. \qquad (2.3.22)$$

If we take into account also that probability of absence in region of overlapping at least one of n satellites is determined by expression (2.3.17), it is possible to find probability of the presence in region of overlap of k or more satellites with their total

number being n. It is obvious that this probability will also determine probability m_k of the realization of the connection/communication through k or larger number of satellites. Thus, upon consideration of expressions (2.3.17 and (2.3.22)

$$m_k = 1 - [p(d < k) + q(n)] = 1 - \sum_{l=0}^{k-1} C_n^l p_1^l (1-p_1)^{n-l}. \quad (2.3.23)$$

Taking into account relationship/ratio (2.3.20), for different values of p_1 it is possible to construct dependence m_k on probability of realization of connection/communication at least through one satellite m_1.

Table 2.3.2.

(1) Число спутников	n	$2n$	$3n$	$4n$
(2) Часть времени m, в течение которой обеспечивается связь	0,9	0,99	0,999	0,9999

Key: (1). Number of satellites. (2). Part of time m, during which is ensured connection/communication.

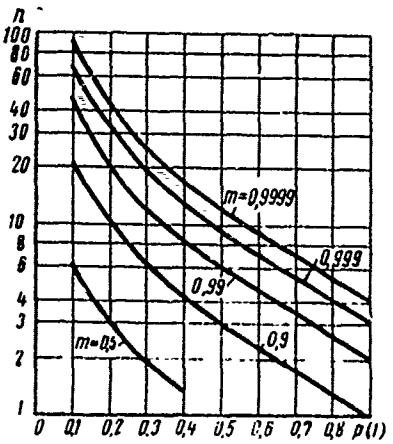

Fig. 2.3.10. Number of satellites, which ensure assigned reliability of connection/communication m.

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Thus, value m_k characterizes that part of the time, during which in the system of the n satellites is possible the redundancy of the connection/communication through k or larger number of satellites, which are located in the region of coupling.

Calculations m_k as functions m_1 with $p_1=p(1)=0.18$ are given in Fig. 2.3.11. As it is noted in [2.7], with change p(1) within some limits value m_k is changed insignificantly; therefore the dependences Fig. 2.3.11 it is possible to use when $0.05 \leq p(1) \leq 0.3$.

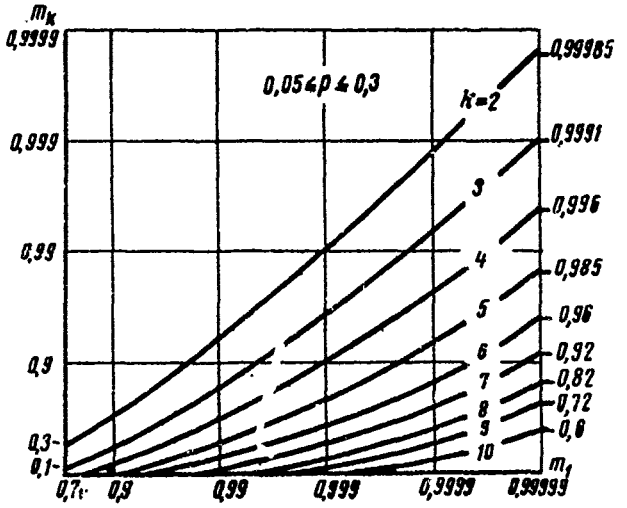

Fig. 2.2.11. Dependence of m_k on m_1 with number of satellites k in zone of mutual visibility.

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Chapter 3.

Signal level at the input of receiver.

3.1. Initial positions.

Indices of communication systems, which use ISZ, are to a considerable extent determined by ratio of signal level to noise at the input of receiver. Correct selection of this relation upon consideration of its possible changes - one of the fundamental factors, which determine the technical-economic effectiveness of the communication system, and also its noise immunity and reliability. Let us examine the schematic of the channel of communication (with extent of L) of Fig. 3.1.1, in which are shown the transmitter П, receiver Пр, transmitting and receiving antennas $A_п$ and $A_{пр}$, characterized by effective surfaces $S_{эп}$ and $S_{эпр}$ and maximum values of amplification factor with respect to isotropic emitter $G_п$ and $G_{пр}$ (taking into account the effect of fairing).

Antennas are connected with receiver and transmitter by feeders $Ф_п$ and $Ф_{пр}$ through filters Ф (band or separating), efficiency of filters and attenuation of filters are designated through $\eta_{Ф_п}$, $\eta_{Ф_{пр}}$, $b_{f_п}$, $b_{f_{пр}}$ respectively. Attenuation of the signal in the communications section considering the antenna isotropic (not directed) is designated by $b_{L_п}$. In this figure $P_п$ defines the power output of the transmitter and $P_{пр}$ - the power of the signal at the receiver input. Besides the signal, interference and noise act at the input of the receiver, the powers of these are $P_х$ and $P_ш$, respectively.

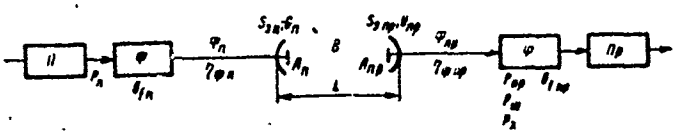

Fig. 3.1.1. Schematic of communication channel.

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Let us designate by symbols without index of "0" absolute power coefficients (W) and attenuation length either amplification (number of times), and by symbols with index "0" - power levels relative to one W (dBW) and attenuation length or amplifications, expressed in dB. Then for the channel of communication, Fig. 3.1.1, the power of signal at the input of receiver can be found from the relationships/ratios:

$$P_{\text{пр}} = \frac{P_{\text{п}}}{B_{\text{з}}}, \tag{3.1.1}$$

$$B_{\text{з}} = \frac{B b_{f_{\text{п}}} b_{f_{\text{пр}}} \varepsilon_{\Phi_{\text{п}}} \varepsilon_{\Phi_{\text{пр}}}}{G_{\text{п}}(\Theta) G_{\text{пр}}(\Theta) \kappa_{\text{пол}}} \tag{3.1.2}$$

or

$$P_{0\,\text{пр}} = P_{0\text{п}} - B_{0\text{з}}, \tag{3.1.3}$$

$$B_{0\text{з}} = B_0 - [G_{0\text{п}}(\Theta) + G_{0\,\text{пр}}(\Theta)] + b_{0f_{\text{п}}} + b_{0f_{\text{пр}}} + \varepsilon_{0\Phi_{\text{п}}} + \varepsilon_{0\Phi_{\text{пр}}} - \kappa_{\text{пол}}, \tag{3.1.4}$$

$$\varepsilon_{0\Phi_{\text{п}}} = 10\lg \varepsilon_{\Phi_{\text{п}}}; \quad \varepsilon_{\Phi_{\text{п}}} = \frac{1}{\eta_{\Phi_{\text{п}}}}, \tag{3.1.5}$$

$$\varepsilon_{0\Phi_{\text{пр}}} = 10\lg \varepsilon_{\Phi_{\text{пр}}}; \quad \varepsilon_{\Phi_{\text{пр}}} = \frac{1}{\eta_{\Phi_{\text{пр}}}}. \tag{3.1.6}$$

Here $B_{\text{з}}$ - total weakening of signal between output of transmitter and input of receiver;

Θ - angle between direction of maximum gain and direction to correspondent (Fig. 3.1.2);

$K_{пол}$ -- value of polarizational losses, caused by nonidentity of polarizational characteristics of antennas and by change in plane of polarization, caused by Faraday effect.

Below, in Section 3.2, we examine the reasons for weakening of signal in free space; in Section 3.3 fundamental parameters of antennas are given, and in Section 3.4 polarizational losses taking into account Faraday effect are determined. The attenuation of filter b_{0f} depends not only on its construction/design and conductivity of material, from which it is prepared, but also from the tuning precision of filter to the band of signal frequencies; in Section 3.5 the determination of drift, caused by the Doppler effect, is given. Let us note that during determination of $b_{f_п}$ and $b_{f_{пр}}$ it is necessary to also consider frequency stability of transmitter.

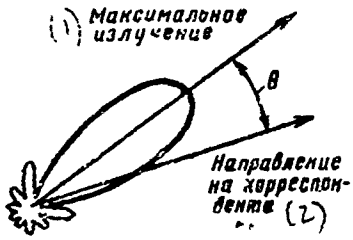

Fig. 3.1.2. Determination of angle Θ.

Key: (1). Maximum emission. (2). Direction to correspondent.

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3.2. Weakening signal in the section between the antennas.

Total weakening B of signal in section of connection/communication with passage from transmitting to receiving antenna is determined by expression

$$B = B_{св} B_{а} \text{ or } B_{0} = B_{0\,св} + B_{0а}. \qquad (3.2.1)$$

Here $B_{св}$ - weakening signal with the passage of free space, when as the transmitting and receiving antenna hypothetical isotropic emitters are examined, and $B_{а}$ - fading signal in transit through the Earth's atmosphere, moreover $B_{св} \gg B_{а}$. One should note that expression (3.2.1) does not consider signal fading during the propagation, since usually in the communication systems through ISZ with the work with comparatively large angles of elevation of fading virtually they are absent. The latter is confirmed by the comparison of the theoretical and experimental values of signal levels.

Value $B_{cв}$ can be determined on formula

$$B_{cв} = \left(\frac{4\pi L}{\lambda}\right)^2. \qquad (3.2.2)$$

Here L - distance between antennas, m,

λ - wavelength, m.

Let us switch over to determination of value B_a. With the passage of the radio waves through the atmosphere it is necessary to consider the absorption of their energy by the atmosphere in the absence of precipitation B_a' and absorption, caused by precipitation, B_a''. Thus, $B_a = B_a' B_a''$ or passing to decibels $B_{0a} = B_{0a}' + B_{0a}''$. As is known, the atmosphere possesses the frequency-dependent properties: the oscillations/vibrations of some frequencies noticeably are weakened/attenuated. The theoretical dependence of attenuation B_{0a}' in the atmosphere on the frequency in the absence of precipitation has two sharply pronounced maximums [3.1]. The first, at frequencies of around 22 GHz, is explained by resonance absorption in the water vapors, and the second, at the frequency of 60 GHz, by resonance absorption in oxygen [3.2].

Attenuation length depends on continuously changing state of atmosphere and on path length of radio waves in the atmosphere. Consequently, absorption will be minimum at the angle of elevation of $\beta=90°$, and maximum with the values β, close to zero. During the determination of attenuation a certain role plays the height/altitude

of station above sea level, since it, just as value β, characterizes the path length of radio waves in the atmosphere.

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For the determination of attenuation in uncontaminated atmosphere of area with moderate humidity in the absence of precipitation in report of MKKR [3.3] are given theoretical curves of Fig. 3.2.1, at $\beta=90°$. Attenuation for other angles of elevation can be determined according to the expression

$$B'_{0a} = B'_{0a}(90°)\alpha, \text{ dB}. \qquad (3.2.3)$$

Here B'_{0a} (90°) - attenuation in the atmosphere for vertical route ($\beta=90°$), which is determined according to curves of Fig. 3.2.1, and α - coefficient, not depending on frequency. The value of this coefficient depending on angle of elevation β is found from Table 3.2.1 [3.3].

One should note that, according to [3.3], curves 3.2.1 and data of Table 3.2.1 carry preliminary character.

It is necessary to say that work [3.1] gives theoretical attenuation lengths B'_{0a} (Fig. 3.2.2), differing somewhat from those, which are obtained at a rate of formula 3.2.3, recommended by MKKR. The differences indicated, apparently, are explained by the procedure of the calculation of these losses.

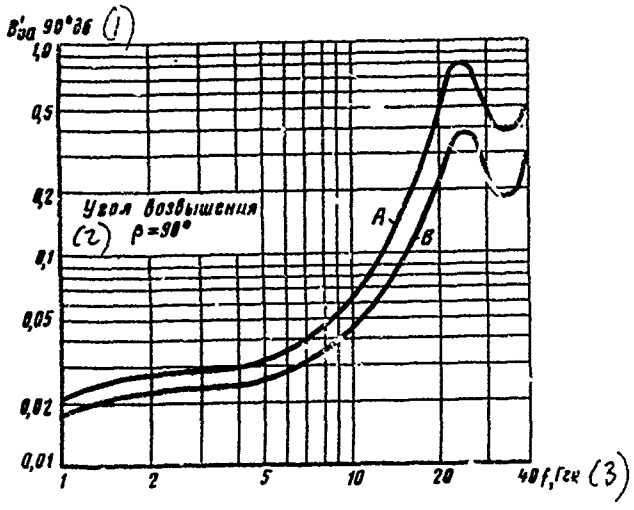

Fig. 3.2.1. Attenuation in the atmosphere without precipitation: A - station at the level of sea, B - station on 2 km is higher than sea level.

Key: (1). dB. (2). Angle of elevation. (3). GHz.

Table 3.2.1.

β	0°	1°	2°	3°	4°	5°	7,5°	10°	30°	90°
α	80	40	25	18	14	11	7,5	5,5	2	1

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Fading signal in the atmosphere, caused by rain, by clouds and by fog in section with length of l km and in transit through radome, can be found from expression

$$B''_{0a} = \delta_0 l + B_{0\kappa}. \qquad (3.2.4)$$

Here value δ_0 determines attenuation in precipitation on route

with extent of 1 km and it can be determined according to Fig. 3.2.3 [3.3, 3.4], and value B_{0_2} characterizes losses, determined by passage of electromagnetic energy through shielding radio-transparent radome, covered with layer of water.

Value l, in expression (3.2.4), in the case of $\beta=90°$ can be undertaken equal from 3 to 10 km. At $\beta\neq90°$ value l is determined by the extent of region, in which falls precipitation. With the drizzling and moderate rains such regions can have sizes/dimensions to 1000 km into the length and from 30 to 300-400 km into the width. In the case of showers the sizes/dimensions of regions compose tens of km into the length and width.

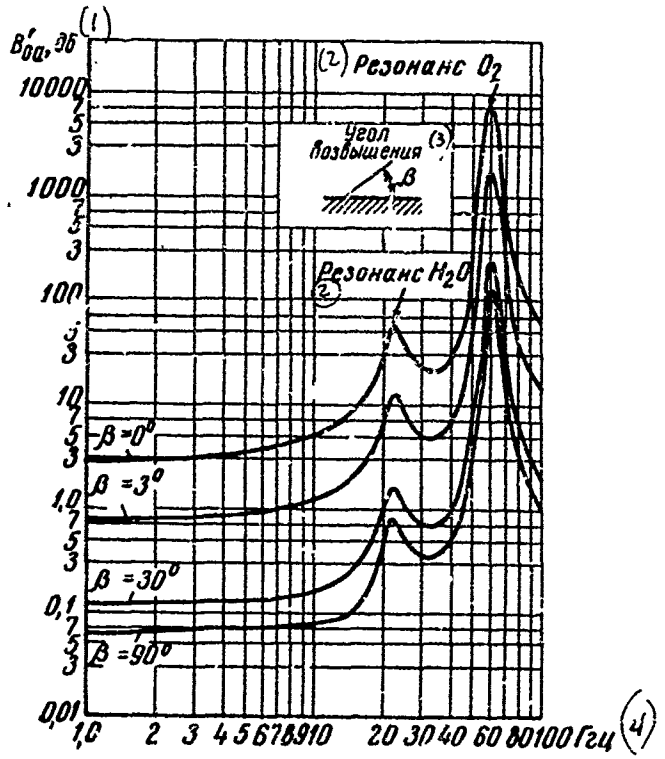

Fig. 3.2.2. Attenuation in the atmosphere without precipitation. Key: (1). dB. (2). Resonance. (3). Angle of elevation. (4). GHz.

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One should note that precision determination of attenuation in precipitation becomes complicated by fact that at identical rainfall intensity distribution of drops in value can be differently. The dependences Fig. 3.2.3 correspond to the typical distribution of drops in the value in the continuous rain.

In order to determine percentage of time, during which occurs this fading of signal in the atmosphere, caused by precipitation, are

necessary statistical data about horizontal and vertical extent of rains, thickness of clouds and distribution of cloud cover. The evaluation, based on the statistic study of clouds [3.3] and rains of South England, at $\beta=5°$ at the frequency of 4 GHz, shows that the attenuation, exceeded during 1% of the time of observations, is approximately 0.25 dB with the rain and about 0.7 dB with the clouds.

Value B_{0_x} is located according to graphs/curves of Fig. 3.2.4 [3.11]. Dependences are designed on the basis of works [3.12 and 3.13]; in this case solid lines in the figure showed losses in a layer of the water, which covers fairing during the perpendicular incidence/drop in the energy, and broken - joint losses, caused not only by a layer of water by thickness κ, but also by the fairing, made from the material with a thickness of about 1.5 mm with the dielectric constant $\epsilon=3-i\ 0.045$.

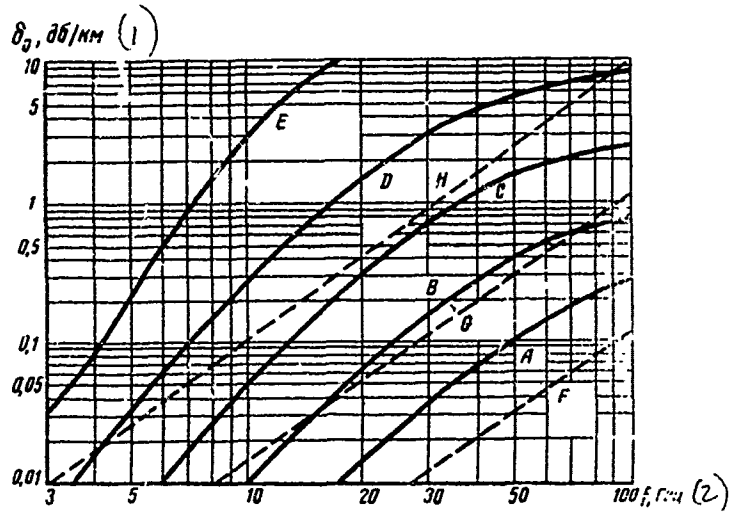

Fig. 3.2.3. Attenuation, caused by precipitation on route 1 km: A - attenuation with rain 0.25 mm/h (drizzling rain), B - attenuation with rain 1 mm/h (weak rain), C - attenuation with rain 4 mm/h (moderate rain), D - attenuation with rain 16 mm/h (strong rain), E - attenuation with rain 100 mm/h (shower), F - attenuation with fog or cloudiness 0.032 g/m³ (visibility >600 m), G - attenuation with fog or cloudiness 0.32 g/m³ (visibility ~120 m), H - attenuation with fog or cloudiness 2.3 g/m³ (visibility ~30 m).

Key: (1). dB/km. (2). f, GHz.

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Fig. 3.2.5 gives experimental dependence of losses due to wet cap/hood for rain of different intensity, taken for antenna in Andover [3.14] at frequency of 4000 MHz. One should note that depending on the climatic conditions, and also on form and material of fairing, by which is determined the thickness of the layer of water on it, magnitude of losses, caused by fairing, can differ somewhat from the

dependence Fig. 3.2.5.

Taking into account everything said, with substitution into expression (3.2.1) of values, determined (3.2.2), (3.2.3) and (3.2.4), we will obtain

$$B_\theta = 20\lg\left(\frac{4\pi L}{\lambda}\right) + B'_{0\alpha}(90^\circ)\alpha + \delta_0 l + B_{0\alpha}. \qquad (3.2.5)$$

During determination of direction of maximum antenna radiation it is necessary to consider atmospheric refraction, i.e., deviation of direction of radio beam from straight line. This phenomenon more strongly is developed at small angles of increase and does not depend on frequency in the range 0.1-10 GHz [3.3].

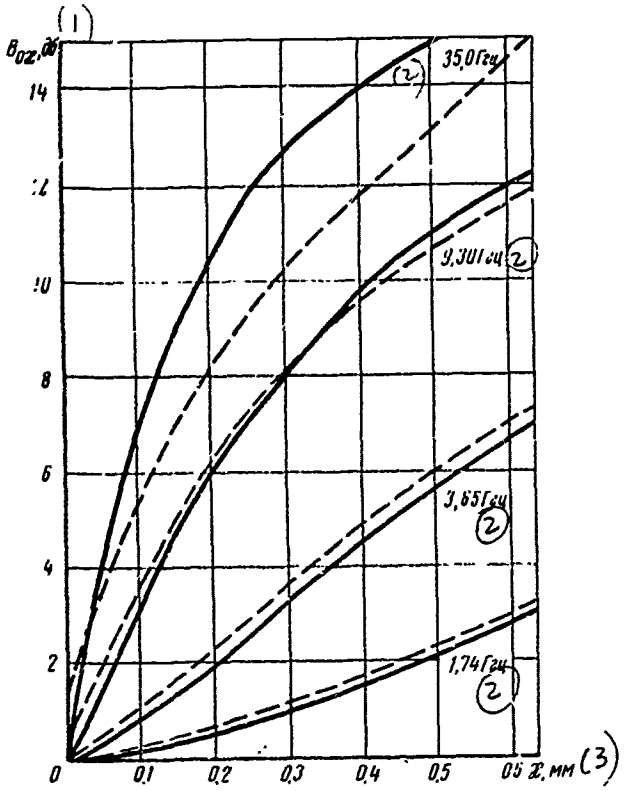

Fig. 3.2.4. Losses, determined by wet fairing.

Key: (1). dB. (2). GHz. (3). κ, mm.

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Depending on angle of elevation β the average/mean value of a change in the angle of incidence of radio beam is determined according to graph/curve, Fig. 3.2.6 [3.3]. From the curves it follows that the error in the angle of arrival of radio beam when β of more than 5° proves to be low (less than 0.2°). Nevertheless this value for the antennas with the high amplification factor must be considered during the determination of antenna bearing to the side of satellite. Otherwise is possible the decrease of signal level or even "loss" of

satellite. If the width of major antenna lobe exceeds 0.7±1°, at frequencies of higher than 1000 MHz the refraction can be disregarded/neglected.

3.3. Fundamental parameters of antennas.

Requirements, presented to antennas in communication systems through ISZ, are reduced in essence to obtaining of necessary directive gain (KND), factor of amplification and minimum side-lobe level radiation pattern.

For any antenna KND it is identical both during transmission and reception.

Let us examine determination of KND based on example of transmitting antenna [3.5].

Let us place antenna into point O of space and will describe sphere, whose center coincides with point O (Fig. 3.3.1). We will consider radius of sphere r to be so large that the form of solid coverage of antenna in the region of sphere will be virtually the same as in the infinite region.

Strength of field at any point of sphere in general form can be represented as

$$E = AF(\theta, \varphi). \tag{3.3.1}$$

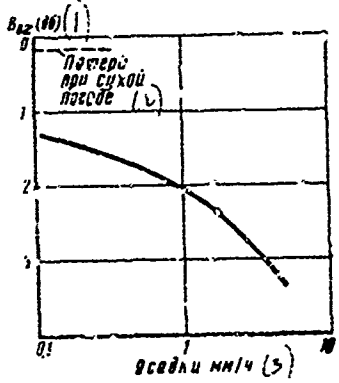

Fig. 3.2.5.

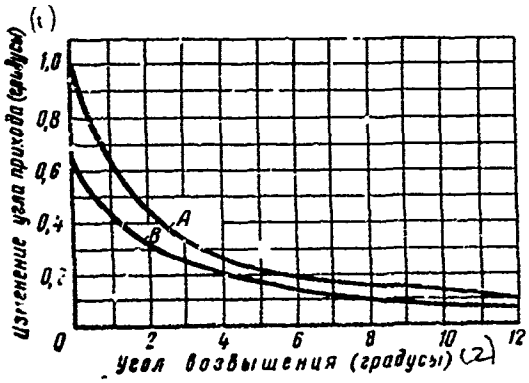

Fig. 3.2.6.

Fig. 3.2.5. Losses, determined by wet cap/hood.

Key: (1). dB. (2). Precipitation, mm/h.

Fig. 3.2.6. Average/mean value of change in angle: A - tropical maritime air, B - polar continental air.

Key: (1). Change in the angle of arrival (degrees). (2). Angle of elevation (degrees).

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Here A - proportionality factor, not depending on Θ and φ;

$F(\Theta, \varphi)$ - form of antenna radiation pattern for that point of sphere, which is determined by coordinates Θ and φ radius of vector.

KND of antenna is called relation of square of strength of field in direction in question to average/mean (in all directions) value of square of strength of field:

$$D_{\text{н}} = \frac{E^2(\Theta_0, \varphi_0)}{E^2_{\text{ср}}}. \qquad (3.3.2)$$

Here $E(\Theta_0, \varphi_0)$ - strength of field on element of surface of

sphere dF with coordinates Θ_0 and φ_0 (Fig. 3.3.1),

$$E_{cp} = \frac{\int_F E^2 dF}{F},$$

where F - surface of sphere.

Using expression for E_{cp}, we will obtain

$$D_H = \frac{F^2(\theta_0, \varphi_0) F}{\int_F F^2(\theta, \varphi) dF}. \qquad (3.3.3)$$

Taking into account that surface of sphere $F=4\pi r^2$, and $dF=r^2 \cos \Theta d\Theta d\varphi$, can be obtained

$$D_H = \frac{4\pi F^2(\theta_0, \varphi_0)}{\int_0^{2\pi} d\varphi \int_{-\pi/2}^{\pi/2} F^2(\theta, \varphi) \cos\Theta \, d\theta}. \qquad (3.3.4)$$

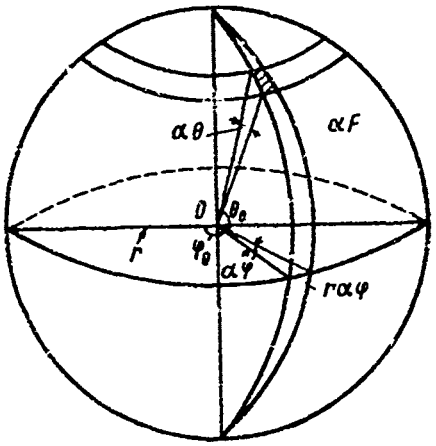

Fig. 3.3.1. Determination of parameters of antennas.

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During calculation of KND to more conveniently use value, calibrated relative to $F(\theta_0, \varphi_0)$. Therefore, if we introduce the designation

$$\frac{F(\theta,\varphi)}{F(\theta_0,\varphi_0)} = F_1(\theta,\varphi),$$

we will obtain

$$D_н = \frac{4\pi}{\int_0^{2\pi} d\varphi \int_{-\pi/2}^{\pi/2} F_1(\theta,\varphi)\cos\theta\, d\theta}. \qquad (3.3.5)$$

Relation

$$G = \frac{\frac{E^2}{P_0}}{\frac{E_н^2}{P_н}} \qquad (3.3.6)$$

is called antenna gain G.

Here E and $E_н$ - strengths of fields, created at point of space in question with respect to this antenna and a nondirectional antenna (isotropic), creating identical strength of field in all directions.

P_0 and $P_н$ - power, applied to this antenna and omnidirectional antenna.

Value G is determined when $P_0=P_н$, and efficiency of isotropic antenna is equal to one.

Let us note that for any antenna by efficiency η_A is understood relation

$$\eta_A = \frac{P_\Sigma}{P_0} = \frac{P_\Sigma}{P_\Sigma + P_{пот}}, \qquad (3.3.7)$$

where P_Σ - power, emitted by antenna, $P_{пот}$ - power, lost in antenna.

Substituting (3.3.2) into expression (3.3.6) and taking into account aforesaid, we will obtain

$$G = \frac{D_н E_{ср}^2}{E_н^2}.$$

Keeping in mind (3.3.7), let us find

$$G = D_н \eta_A. \qquad (3.3.8)$$

On the basis of reciprocity principle it is possible to demonstrate that parameter G of any antenna is numerically identical at transmission and reception.

Usually value η_A is very close to one; therefore with accuracy sufficient for practical purposes it is possible to count $G = D_н$, i.e. that factor of amplification and directive gain of antenna coincide.

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Receiving antennas are conveniently characterized by the effective surface, which according to [3.5] is determined by expression

$$S_{эфф} = \frac{\lambda^2}{4\pi} D_н = \frac{\lambda^2 G}{4\pi\eta_A}. \qquad (3.3.9)$$

From (3.3.9) it is possible to obtain another expression, frequently utilized during calculations

$$G = \frac{4\pi}{\lambda^2} S_{эфф}\, \eta_A \cong \frac{4\pi}{\lambda^2} S_{эфф}. \qquad (3.3.10)$$

There is several types antenna of SHF; however, widest use in radiolink systems through ISZ received parabolic, horn-parabolic antennas and some of their modifications - "convoluted" horn-parabolic antenna [3.6] and two-mirror antenna of Cassegrain [3.5] and [3.7].

Directive gain of parabolic antenna with radius of aperture of R_0 and focal length of f (see Fig. 3.3.2) is determined by following expression:

$$D_н = \frac{4\pi S}{\lambda^2} \kappa_1. \qquad (3.3.11)$$

Here $S=\pi R_0^2$ - surface of aperture of paraboloid, $\kappa_1 = f\left(\frac{R_0}{f}\right) D_{об.п} \frac{S_{эфф}}{S}$ - coefficient of use of surface of aperture of paraboloid, which considers nonuniformity of phase distribution and

amplitudes of field in aperture,

$D_{\text{ф.и}}$ - KND of antenna feeder.

Let us note that depending on type of antenna and utilized irradiator value k_1 can vary within the range of 0.55 to 0.96 [3.7]; however, in practice usually there is order 0.5=0.6 for parabolic antennas, 0.65-0.78 for horn-parabolic and 0.6-0.65 for two-mirror.

Substituting in expression (3.3.11) enumerated values with value of k_1=0.5, we will obtain approximation formula for determining KND

$$D_{\text{н}} \cong 20\left(\frac{R_0}{\lambda}\right)^2 = 5\left(\frac{D_0}{\lambda}\right)^2. \qquad (3.3.12)$$

Here R_0 and D_0 respectively radius and diameter of aperture of reflector (see Fig. 3.3.2).

Radiation pattern of parabolic antenna can be calculated from expression (3.3.5).

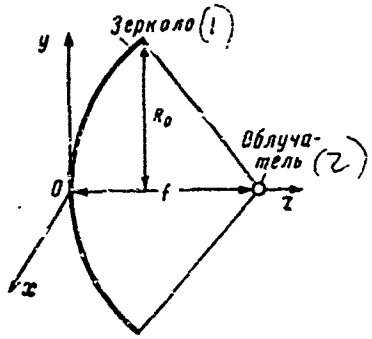

Fig. 3.3.2. Parameters of parabolic antenna.

Key: (1). Mirror. (2). Irradiator.

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Fig. 3.3.3 [3.5] shows the radiation patterns of this antenna in planes xz and yz with the different values of ratio R_0/f (see Fig. 3.3.2). By the selection it it is possible substantially decrease side-lobe level and the width of the major lobe/lug of the antenna radiation pattern.

Value of minor lobes of receiving antennas determines noise level and interferences/jammings, taken out of main trend, and in the case of transmitting antennas - interference level, created to other radio aids. For the parabolic antenna optimum value $\frac{R_0}{f}=1{,}3$, in this case is obtained maximum value $k_1=0.83$, and consequently, and the maximum value of directive gain $D_{\text{н}}$.

For this case width of radiation pattern of parabolic antenna on half power it is determined by following expressions [3.5]:

in plane yz

$$\theta_{0,5yz} \cong 1,2 \frac{\lambda}{2R_0}, \text{ rad, (3.3.13).}$$

in plane xz

$$\theta_{0,5xz} \cong 1,3 \frac{\lambda}{2R_0}, \text{ rad, (3.3.14).}$$

or in the degrees

$$\left.\begin{aligned} \theta^0_{0,5yz} &\cong 70 \frac{\lambda}{2R_0} \\ \theta^0_{0,5xz} &\cong 75 \frac{\lambda}{2R_0} \end{aligned}\right\}. \qquad (3.3.15)$$

Keeping in mind (3.3.13) and (3.3.14), expression (3.3.11) can be recorded in the form

$$D_{\max} = 4\pi \frac{\pi R_0^2}{\lambda^2} \kappa_1 = \pi^2 1,2 \cdot 1,3 \frac{1}{\theta_{0,5xz}\, \theta_{0,5yz}} \kappa_1.$$

Accepting $k_1=0.5$ and expressing angles in degrees, we will obtain approximation formula, convenient for calculations of KND of antenna with given (or known) values of width of major lobe/lug

$$D_{\max} = \frac{25000}{\theta^0_{0,5xz}\, \theta^0_{0,5yz}}. \qquad (3.3.16)$$

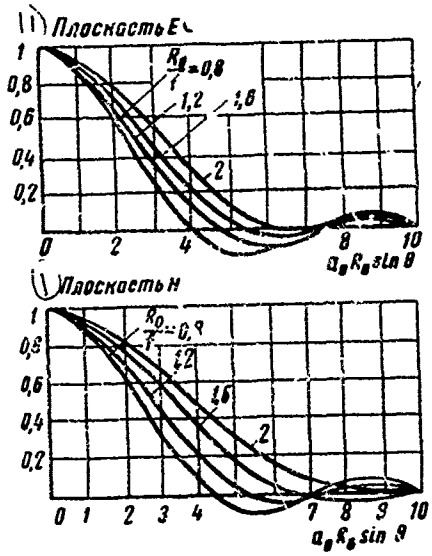

Fig. 3.3.3. Radiation patterns of parabolic antenna.

Key: (1). Plane.

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Expression (3.3.16) correctly, when angles $\theta_{0,5\,xz}$ and $\theta_{0,5\,yz}$ do not exceed several degrees, i.e., when antennas have sufficiently high amplification factor. Let us note that expressions (3.3.12)-(3.3.16) are approximate and can be utilized for the evaluation of the parameters of parabolic as well as horn-parabolic antennas and their modifications.

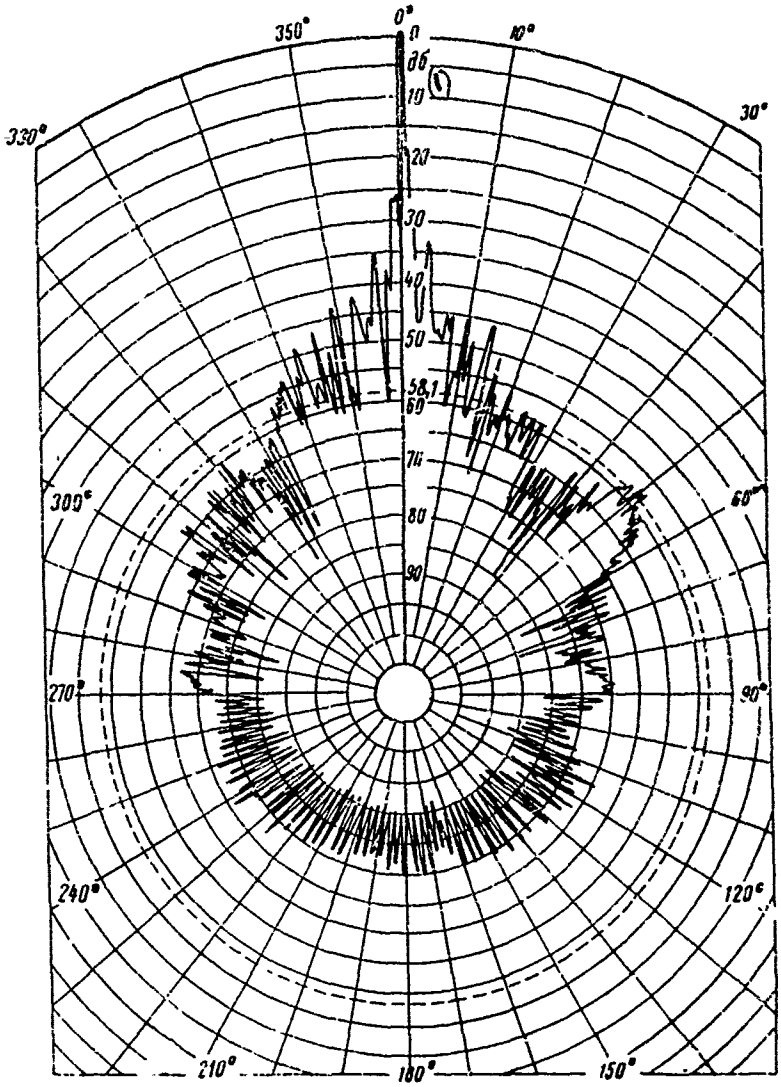

Fig. 3.3.4. Antenna radiation pattern: - - - - - level of corresponding diagram of isotropic emitter.
Key: (1). dB.

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Radiation patterns of real antennas usually differ from those designed due to imperfect irradiation of surface of reflector, which leads to nonuniformity of phase distribution and amplitudes of field,

inaccuracy in manufacture of irradiator and reflector, presence in field of irradiator of different structural parts of antenna (brackets, attachment, etc.). All this leads to the brokenness of radiation pattern - to the appearance of a large number of minor lobes (Fig. 3.3.4), whose calculation presents considerable difficulties. For this reason for the calculations are utilized the idealized antenna radiation patterns, in which are considered only the averaged values of gain in different directions due to the minor lobes.

Fig. 3.3.5 and 3.3.6 give "those averaged" on minor lobes of radiation pattern of terrestrial antennas of different types (not only parabolic) [3.8], these antennas having aperture (aperture) from 9 to 27 m, and amplification factors - from 45 to 60 dB. In the same figures the straight line, which approximates the antenna gain $G(\Theta)$ at $\Theta \leq 50°$, constructed according to the expression

$$G(\Theta)\ 32-25\lg(\Theta),\ \text{dB} \qquad (3.3.17)$$

is plotted/applied.

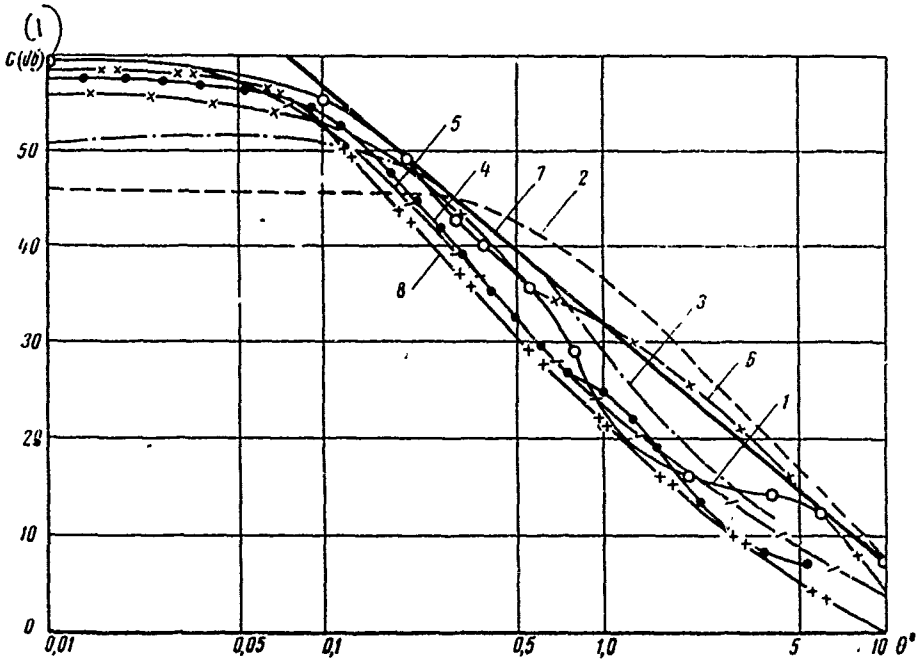

Fig. 3.3.5. Gain and approximation at $0.01°>\Theta>10°$: 1 - Mill Village, 2 - Goldstone, 3 - Wallops Island, 4 - Andover, 5 - West Ford, 6 - Goonhilly, 7 - $G(\Theta)=32-25 \lg \Theta$, 8 - Goonhilly after change. Key: (1). dB.

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Here Θ - angle (in degrees) between axis of major lobe/lug of antenna radiation pattern and selected direction. From the comparison of the antenna radiation pattern and idealized diagram their satisfactory coincidence follows. Therefore in work [3.8] is done the conclusion that in the limits of value of $\Theta\leq50°$ amplification of the minor lobes of different antennas, found relative to isotropic emitter, can be approximated by expression (3.3.17). At $\Theta>50°$ it is considered that the amplification of minor lobes is -10 dB relative to isotropic emitter. Within the limits of the main lobe of radiation of

antenna its amplification is accepted as constant/invariable and that depending on the ratio of the diameter of antenna to the wavelength. Value and the width of major antenna lobe can be designed from approximate formulas (3.3.12)-(3.3.16).

Fig. 3.3.7 gives idealized antenna radiation pattern, constructed in accordance with expression (3.3.17) and expressed considerations. In this case the gain in major lobe can be determined according to expression (3.3.12).

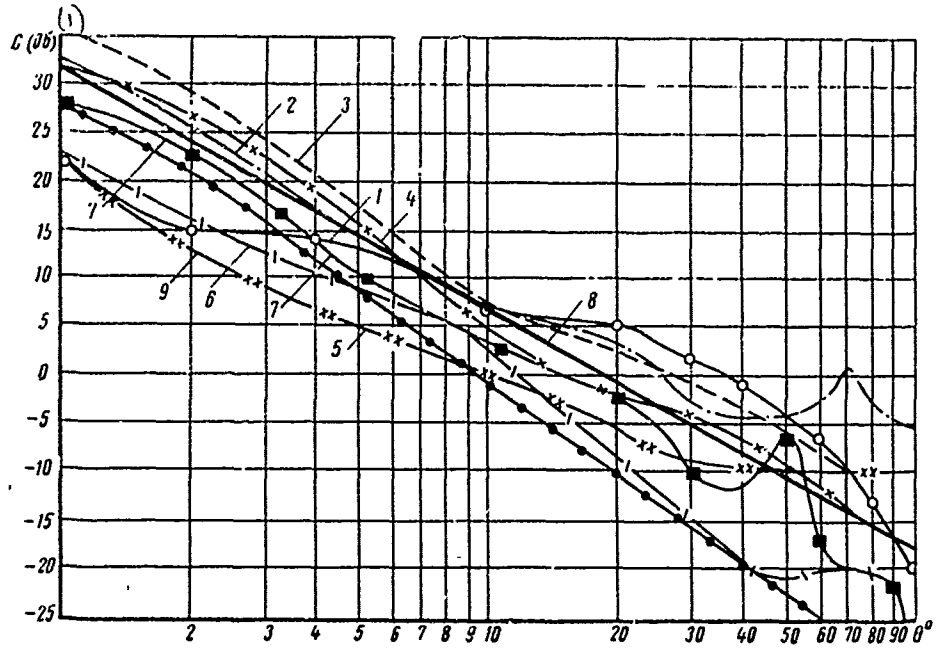

Fig. 3.3.6. Gain and approximation at $1°<\Theta<100°$: 1 - Mill Village, 2 - Holmdel, 3 - Goldstone, 4 - Goonhilly, 5 - Andover, 6 - West Ford, 7 - Raysting, 8 - $G(\Theta)=32-25 \lg \Theta$, 9 - Goonhilly after change.

Key: (1). dB.

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In [3.8] it is noted that accuracy of approximation in Fig. 3.3.7 is approximately ±3 dB for horizontal sections and about 4 dB in inclined section.

As it follows from given formulas at selected frequency (or wavelength), increase in antenna gain can be achieved/reached by increase in its sizes/dimensions. However, with an increase in the sizes/dimensions are unavoidable the large deviations of the form of reflector from the given one, and also the appearance of considerable

nonuniformities of the surface of antenna dish. This will destroy the coplanarity of field in antenna aperture. Tolerance level for the deviation from the theoretical form of mirror must not exceed $\lambda/16$. With the large sizes/dimensions of antennas and at the high frequencies this requirement becomes impracticable. Fig. 3.3.8 shows the dependence of amplification on the frequency for several very precisely executed antennas, whose ratio of maximum standard deviation σ to the diameter of antenna is equal to 10^{-4} [3.9]. From Fig. 3.3.8 it follows that the maximum gain in this case is approximately 57 dB. Fig. 3.3.9 gives the dependence of the diameter of antenna with the value $(\sigma/D)=10^{-4}$ from the frequency, at which is obtained maximum gain.

Comparison of parabolic antennas with horn-parabolic shows that latter possess considerably smaller minor lobes. The considerably large sizes/dimensions with the same amplification factors are a definite deficiency in the horn-parabolic antennas in comparison with the parabolic antennas. However, structural/design changes in horn-parabolic antennas [3.6] make it possible to considerably shorten their sizes/dimensions almost without a change in the electrical characteristics.

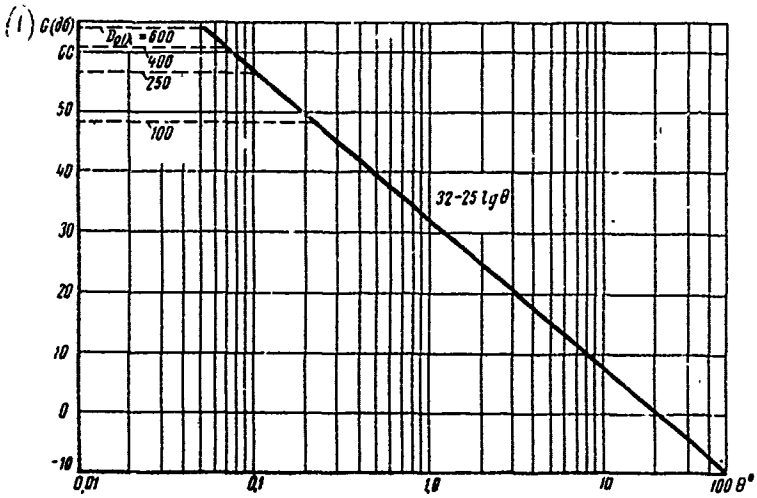

Fig. 3.3.7. Approximation of gain in direction Θ: D_0 - diameter of aperture, m, λ - wavelength, m.

Key: (1). dB.

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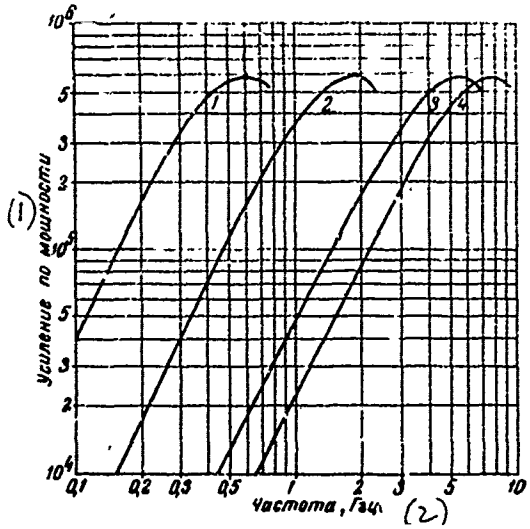

Fig. 3.3.8. Dependence of amplification on frequency for several values of the diameter of antenna with $\sigma/D = 10^{-4}$: 1 - D=230 m, 2 - D=76 m, 3 - D=26 m, 4 - D=18 m.

Key: (1). Power gain. (2). Frequency, GHz.

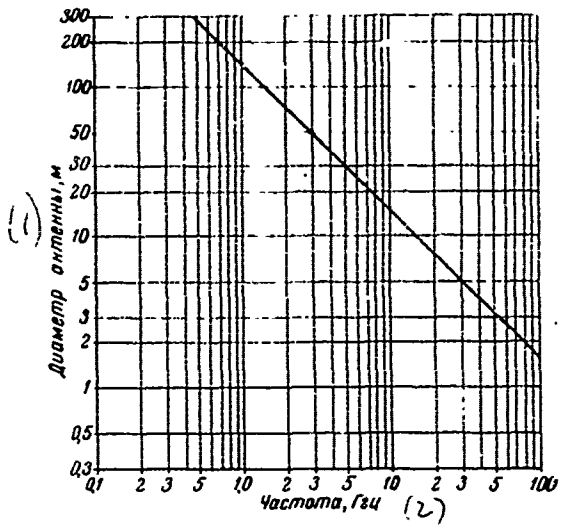

Fig. 3.3.9. Dependence of frequency, which corresponds to maximum amplification on diameter of antenna with $\sigma/D=10^{-4}$.

Key: (1). Diameter of antenna. (2). Frequency, GHz.

One should note that ground antennas of communication systems through ISZ sometimes are placed within radio-transparent fairings. The latter, shielding antennas from the wind load and precipitation, allow together with simplification in the construction/design of antenna system (reduction in the weight, an improvement in the evenness and the accuracy of the displacement/movement of antenna, decrease of the power of drive assemblies) to simplify both operation and servicing of antennas in any weather.

Radomes are of two types - inflatable and rigid. Inflatable fairings do not have a framework/body and are fulfilled from the dielectric film in the form of the sphere, inside which is pumped air. At present more frequently are used the rigid fairings, which consist of the framework/body, to which are fastened the parts and the elements of construction/design.

It is obvious that fairing, being located in immediate proximity to antenna, will somewhat change its electrical characteristics. Thus, according to [3.10] contemporary fairings in the range of SHF in the absence of precipitation create, approximately, the following changes in the parameters of the antenna:

1. Decrease of amplification factor on 0.5 dB.

2. Change in side- lobe levels

on 1 dB on level - 20 dB,

on 2 dB on level - 25 dB,

on 4 dB on level - 30 dB.

3. Error in guidance of beam of 0.005-0.25°.

Effect of wet radomes on signal level is examined in Section 3.2, and on value of noise temperature of antenna - in Section 4.4.

3.4. Polarizational losses.

Let us examine losses, caused by disagreement of polarization of oscillations/vibrations of transmitting and receiving antennas. We will consider that the terminus of the vector of electric field E (or magnetic field H), created at the point of the space of certain antenna in question, describes during the period of high frequency the ellipse. This ellipse lies/rests at plane x-y, perpendicular to the direction of propagation of energy z, and is called polarizational (Fig. 3.4.1).

Let us represent vector E in the form of sum of two components: vector E_x and E_y. It is obvious that in the general case the instantaneous values of the coordinates, which determine the value of components E_x and E_y, can be recorded in the form:

$$x = E_1 \sin \omega t, \tag{3.4.1}$$

$$y = E_2 \sin (\omega t - \alpha). \tag{3.4.2}$$

Here E_1 and E_2 - maximum values of components E_x and E_y, and value α characterizes phase displacement between E_x and E_y.

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Keeping in mind, that according to (3.4.1)

$$\sin \omega t = \frac{x}{E_1}$$

after substitution in (3.4.2) let us find

$$y = E_2[\sin \omega t \cos \alpha - \cos \omega t \sin \alpha] = E_2\left[\frac{x}{E_1}\cos \alpha - \sqrt{1 - \frac{x^2}{E_1^2}}\sin \alpha\right].$$

Hence

$$\frac{x^2}{E_1^2} + \frac{y^2}{E_2^2} - \frac{2xy\cos \alpha}{E_1 E_2} = \sin^2 \alpha. \qquad (3.4.3)$$

Expression (3.4.3) is equation of ellipse in system of coordinates x, y. With $\alpha = n\pi$, where n=0, 1, 2, ..., expression (3.4.3) will take form $y=(E_2/E_1)x$, which corresponds to field with the linear polarization.

In the case of circular polarization $E_1=E_2$, expression (3.4.1) and (3.4.2) can be recorded in the form

$$\left.\begin{aligned} x &= E\sin \omega t \\ y &= E\cos \omega t \end{aligned}\right\}. \qquad (3.4.4)$$

Hence follows that oscillation/vibration with circular polarization it is possible to represent in the form of sum of two oscillations/vibrations with identical amplitude and linear polarization, shifted angle $\alpha = (\pi/2)$, excited in orthogonal planes. Fig. 3.4.2 shows the diagram, which illustrates the principle of

obtaining oscillations/vibrations with the circular polarization: to the emitters, which create the linearly polarized vibrations in the orthogonal planes x and y, vibrations E cos ωt are fed/conducted. In this case for obtaining the circular polarization before one of the emitters the device/equipment, which turns the phase of vibrations on 90°, is established/installed.

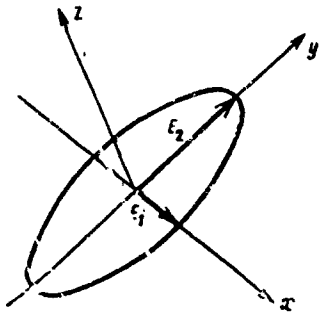

Fig. 3.4.1.

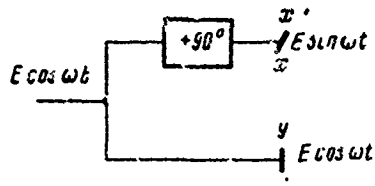

Fig. 3.4.2.

Fig. 3.4.1. Ellipse of polarization.

Fig. 3.4.2. Principle of obtaining vibrations with circular polarization.

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Let us consider case, when $\alpha=\pi/2\ (2n+1)$, and $E_1 \neq E_2$. In this case expression (3.4.3) of signs the form

$$\frac{x^2}{E_1^2}+\frac{y^2}{E_2^2}=1,$$

which corresponds to the ellipse, whose axes coincide with the coordinate axes.

In general case, when axes of ellipse of polarization form with coordinate axes angle γ (Fig. 3.4.3), for projections of vectors E_1 and E_2 on coordinate axis it is possible to record

$$\left.\begin{aligned} E_x &= E_2\cos\omega t\sin\gamma+E_1\sin\omega t\cos\gamma \\ E_y &= E_2\cos\omega t\cos\gamma-E_1\sin\omega t\sin\gamma \end{aligned}\right\}. \qquad (3.4.5)$$

Let us show that oscillation/vibration with elliptical polarization, characterized by components E_1 and E_2 can be represented in the form of two orthogonal oscillations/vibrations E* and E** with circular polarization in opposite directions and characterized respectively by amplitudes E ' and E" and by phases φ', φ'' (Fig. 3.4.4.). If we through j and $\bar{j}$ designate the unit vectors, directed along x and y axes, then

$$\left.\begin{aligned} E^* &= jE'\cos(\omega t+\varphi')+\bar{j}E'\sin(\omega t+\varphi') \\ E^{**} &= jE''\cos(\omega t+\varphi'')-\bar{j}E''\sin(\omega t+\varphi'') \end{aligned}\right\}. \qquad (3.4.6)$$

In accordance with Fig. 3.4.4.

$$\left.\begin{aligned} E_x &= E'\sin(\omega t+\varphi')+E''\sin(\omega t+\varphi'') \\ E_y &= E'\cos(\omega t+\varphi')-E''\cos(\omega t+\varphi'') \end{aligned}\right\}. \qquad (3.4.7)$$

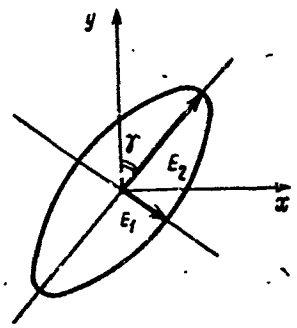

Fig. 3.4.3.

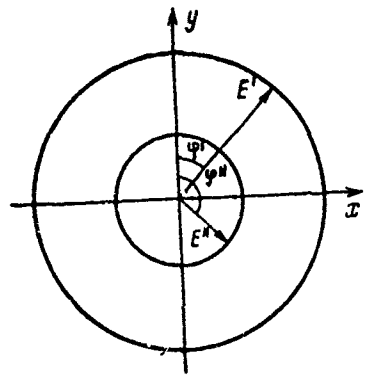

Fig. 3.4.4.

Fig. 3.4.3. Location of ellipse in coordinate system.

Fig. 3.4.4. Orthogonal components with circular polarization.

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Equalizing (3.4.5) and (3.4.7), after introduction of elliptic coefficient $e_1=E_2/E_1$ let us find

$$\left.\begin{aligned} \varphi' &= \gamma; \quad \varphi'' = -\gamma \\ E' &= \frac{E_1+E_2}{2} = \frac{E_1}{2}(1+e_1) \\ E'' &\frac{E_1+E_2}{2} = \frac{E_1}{2}(1-e_1) \end{aligned}\right\}. \qquad (3.4.8)$$

After substitution (3.4.8) in expressions (3.4.6) let us find

$$\left.\begin{aligned} E^* &= j\frac{E_1}{2}(1+e_1)\cos(\omega t+\gamma)+i\frac{E_1}{2}(1+e_1)\sin(\omega t+\gamma) \\ E^{**} &= j\frac{E_1}{2}(1-e_1)\cos(\omega t-\gamma)-i\frac{E_1}{2}(1-e_1)\sin(\omega t-\gamma) \end{aligned}\right\}. \qquad (3.4.9)$$

From expressions (3.4.9) it follows that oscillation/vibration with elliptical polarization (elliptic coefficient is equal to e_1),

actually, it is possible to represent circular polarization in the form of two orthogonal oscillations/vibrations E* and E**, having circular polarization. In this case circular polarization for each oscillation/vibration (E* and E**) is determined by the sum of two oscillations/vibrations, which have linear polarization in the mutually orthogonal planes. By other words, the principle of obtaining the elliptically polarized vibration, by analogy with Fig. 3.4.2, can be represented in the form of the diagram, given in Fig. 3.4.5.

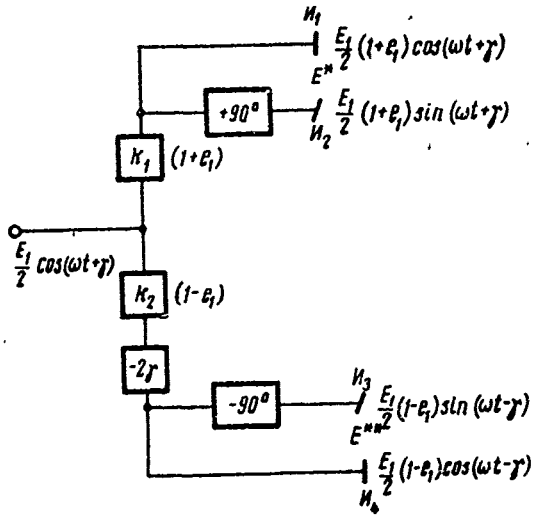

Fig. 3.4.5. Principle of obtaining vibrations with elliptical polarization.

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To the input of this diagram is fed/conducted the vibration with an amplitude of $E_1/2$, which through the elements κ_1 and κ_2 with the transmission factors respectively $1+e_1$ and $1-e_1$ proceeds to emitters $И_1$-$И_4$ directly or passing through the devices/equipment, which turn the phase of vibration on +90° (for component E*) or on 2γ and -90° (for component E**). Let us note that the emitters $И_1$ and $И_4$ excite vibrations with the linear polarization, but different amplitude in one and the same plane, orthogonal to the plane, in which the vibrations are excited by emitters $И_2$ and $И_3$.

As per theorem of reciprocity receiving antenna with elliptical

polarization can be represented in the form of analogous diagram, given in Fig. 3.4.6. Receiving dipoles Д$_1$ and Д$_4$ are located in one plane, and dipoles Д$_2$ and Д$_3$ - in other, orthogonal plane. Components E*, which arrive from the emitters И$_1$ and И$_2$ (see Fig. 3.4.5) on the dipoles Д$_1$ and Д$_2$ of receiving antenna, induce emf:

$$E^{*}_{Д_1} = k\frac{E_1}{2}(1+e_1)\cos(\omega t+\gamma),$$

$$E^{*}_{Д_2} = k\frac{E_1}{2}(1+e_1)\sin(\omega t+\gamma).$$

Here k - proportionality factor determined by attenuation between transmitting and receiving antennas. After the passage through phase inverter of 90° at the output of element with the transmission factor κ' we will obtain the total vibration (summation it is realized according to the power):

$$E^{*}_{Д_1Д_2} = k\kappa'\frac{E_1}{\sqrt{2}}(1+e_1)\cos(\omega t+\gamma). \qquad (3.4.10)$$

With arrival of component E** from emitters И$_4$ and И$_3$ (see Fig. 3.4.5) on dipoles Д$_1$ and Д$_2$ of receiving antenna we will obtain emf:

$$E^{**}_{Д_1} = -\frac{E_1}{2}k(1-e_1)\sin(\omega t-\gamma),$$

$$E^{**}_{Д_2} = \frac{E_1}{2}k(1-e_1)\cos(\omega t-\gamma).$$

From output of element with transmission factor κ' after passage through phase inverter of 90° in this case we will obtain $E^{**}_{Д_1Д_2}=0$.

Latter is explained by the fact that due to phase inverters vibrations, passing through И$_3$ and Д$_2$ differ in phase from vibrations, which go through И$_2$ and Д$_2$ on 180° (see Fig. 3.4.5 and 3.4.6).

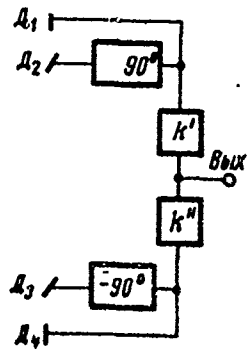

Fig. 3.4.6. Principle of receiving antenna with elliptical polarization.

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After leading analogous considerations for the dipoles Д$_3$ and Д$_4$ let us find that after passing through the phase inverter and the element κ''

$$\left.\begin{aligned} E^{*}_{Д_3Д_4} &= 0 \\ E^{**}_{Д_3Д_4} &= k\kappa'' \frac{E_1}{\sqrt{2}}(1-e_1)\cos(\omega t - \gamma) \end{aligned}\right\}. \qquad (3.4.11)$$

Analogous with diagram in Fig. 3.4.5 for diagram in Fig. 3.4.6 let us assume

$$\left.\begin{aligned} k\kappa' &= \alpha(1+e_2) \\ k\kappa'' &= \alpha(1-e_2) \end{aligned}\right\}, \qquad (3.4.12)$$

where α - constant coefficient, and e_2 - eccentricity of ellipse of polarization of receiving antenna.

Taking into account (3.4.10), (3.4.11) and (3.4.12), we will obtain expression, which determines power of oscillations adopted on

single resistor/resistance

$$P_{пр}=\frac{a^2 E_1^2}{2}\left[(1-e_1)^2(1-e_2)^2+(1+e_1)^2(1+e_2)^2+2\cos 2\gamma(1-e_1^2)(1-e_2^2)\right]. \tag{3.4.13}$$

Amount of transmitted power with single output resistance and axis alignment of ellipse of polarization of transmitting antenna with axes of coordinates ($\gamma=0$) in accordance with (3.4.9) will be

$$P_{п}=E_1^2(1+e_1^2). \tag{3.4.14}$$

Comparing two latter/last expressions, let us find transmission factor of power between two antennas taking into account polarization of oscillations/vibrations

$$\kappa_{пол}=\frac{P_{пр}}{P_{п}}=\frac{a^2}{2}\left[\frac{(1-e_1)^2(1-e_2)^2}{1+e_1^2}+\frac{(1+e_1)^2(1+e_2)^2}{1+e_1^2}+\right.$$
$$\left.+2\frac{(1-e_1^2)(1-e_2^2)}{1+e_1^2}\cos 2\gamma\right].$$

Coefficient a is determined from conditions that with $e_1=e_2$ and $\gamma=0$ value $\kappa_{пол}=1$. In this case

$$a_2=\frac{1}{2(1+e_2^2)}.$$

Therefore

$$\kappa_{пол}=\frac{1}{4}\left[\frac{(1-e_1)^2(1-e_2)^2}{(1+e_1^2)(1+e_2^2)}+\frac{(1-e_1^2)(1-e_2^2)}{(1+e_1^2)(1+e_2^2)}2\cos 2\gamma+\right.$$
$$\left.+\frac{(1+e_1)^2(1+e_2)^2}{(1+e_1^2)(1+e_2^2)}\right], \tag{3.4.15}$$
$$\kappa_{0пол}=10\lg\kappa_{пол}.$$

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Results of calculation $\kappa_{пол}$ according to expression (3.4.15) for different e_1 and e_2 are given in Fig. 3.4.7. If one of the antennas

has circular polarization, and the second - linear (for example, $e_1=1$, $e_2=0$ or vice versa), value $\kappa_{поп}$ will prove to be equal to 0.5 that it corresponds to the decrease of the power adopted two times. Reduction in magnitude $\kappa_{поп}$ with 2 $\gamma=\pi$ is especially noticeable. Thus, for example, with $e_1=e_2=0$, $\gamma=90°$ (antenna with the linear polarization of oscillations/vibrations in the orthogonal planes) magnitude $\kappa_{поп}$ proves to be equal to zero. This means that in this case at the output of receiving antenna the power of signal will prove to be equal to zero. During calculations of the line of communications through ISZ [ИСЗ - artificial earth satellite] it is very important to know not only the form of the polarization of the emitted oscillations/vibrations, but also the direction of rotation of field.

It is customary to assume that field of clockwise rotation for "observer", who looks" in direction of propagation of oscillations/vibrations is characterized by rotation of vector E in time clockwise in fixed plane, perpendicular to direction of propagation.

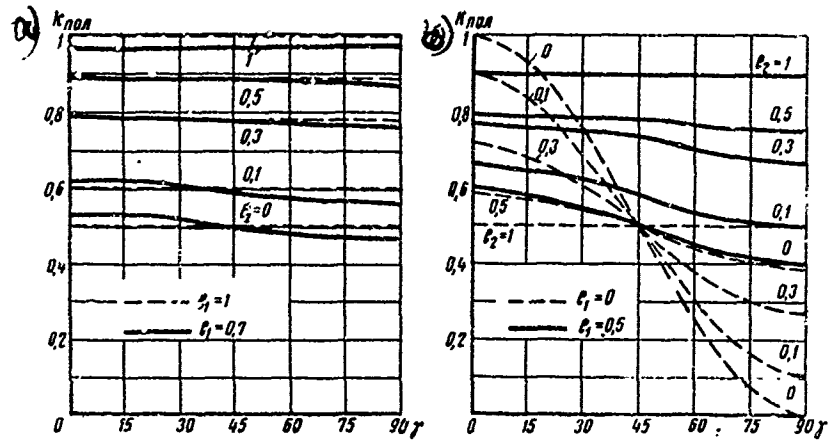

Fig. 3.4.7. Change $k_{пол}$ from different parameters.

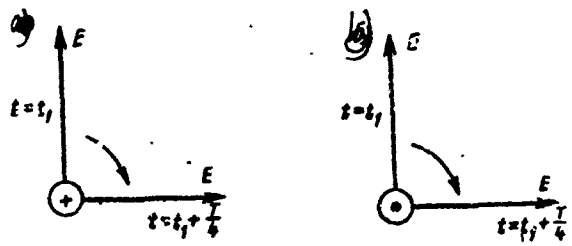

Fig. 3.4.8. Field with polarization of right (a) and left (b) of directions of rotation.

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This clarifies Fig. 3.4.8a, where the position of the vector of electric field E for two moments of time, which differ to value T/4 during the propagation of oscillations/vibrations in the direction from the reader to the plane of drawing, is shown. The latter is marked by sign "+". Thus, Fig. 3.4.8a relates to the polarization of oscillations/vibrations with the right-hand direction of rotation.

Fig. 3.4.8b gives case, when oscillations/vibrations are

propagated in direction from plane of drawing to reader (conditionally markedly by point). This corresponds to the left direction of the rotation of the plane of polarization.

Should be focused attention on the fact that with reflection of electromagnetic energy from ideally reflecting surface phase of vector E is changed on π, i.e., direction of polarization of oscillations/vibrations is changed to opposite. The latter fact must be considered during the selection of the antenna systems of the communication systems, which use the passive communications relays, and also during the use of various antennas (parabolic and of antennas of Cassegrain, etc.).

Rotation of the plane of polarization of oscillations/vibrations in communication systems through ISZ in essence is determined by motion of satellite, as a result of which is changed mutual orientation of onboard and terrestrial antennas. At the same time the rotation of the plane of polarization can be produced by the Faraday effect. This effect appears as a result of the fact that when the passage of the electromagnetic vibration through the ionosphere and the magnetic field of the Earth is present, the oscillation/vibration can be represented by two components, which are propagated through the ionosphere with different phase rates. After passage the phase shift, which leads to the rotation of the polarization of total

signal, appears by these components of a certain path in the ionosphere between them.

If ionized gas is uniform, then value of angle of rotation can be determined [3.15] according to formula

$$\varphi = \frac{ANH_{np}l}{f^2}, \text{ rad}, \quad (3.4.16)$$

where $A=e^3\mu_0/4\pi m^2\epsilon_0$ - proportionality factor;

N - electronic concentration,

H_{np} - longitudinal component of magnetic field, a/m;

l - path length ionized gas, m;

e - electron charge, k;

m - mass of electron, kg;

$\mu_0=4\pi 10^{-7}$ - magnetic permeability, H/m;

$\epsilon_0=10^{-9}/36\pi$ - electrical breakdown of free space, F/m;

f - frequency, Hz.

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Rotation of the plane of polarization of received signal leads to additional energy losses. Actually, in the absence of the fluctuation of the ionosphere it would be possible to orient receiving antenna so that the vector of the electric field of received signal would create greatest power input of receiver. However, due to continuous changes in the electronic concentration the position of the plane of polarization continuously varies. When receiving and transmitting antennas they are fixed, they have linear polarization and they are oriented equally, magnitude of losses due to the rotation of the plane of polarization with the passage of the ionosphere can be determined according to the expression

$$\kappa'_{\text{пол}} = 20\lg\cos\varphi. \qquad (3.4.17)$$

Results of calculation, carried out according to given formulas, are represented in Fig. 3.4.9 [3.1]. In this case it was assumed that $N=2.8\cdot10^{12}$ e/m^3 in the interval of the heights/altitudes of 230-370 km and N=0 at all other heights/altitudes, $H_{\text{пр}}=30$ a/m. As it is marvellous from Fig. 3.4.9, the effect of Faraday it is necessary to consider at frequencies below 3-4 GHz.

Values, entering expression (3.4.16), are characteristic for quiet ionosphere.

Studies of conditions for passage of radio waves through disturbed ionosphere showed [3.16] that aurorae polares increase speed of fadings and are caused rapid and considerable fluctuations of polarization; value of angle φ at frequency of 0.5 GHz exceeded 1 rad. In this case the average/mean of the signal levels accepted remained constant in limits of accuracy of measurements (±2 db). for decreasing the losses, caused by the rotation of polarization, at transmitting and receiving ends of the line of communications antenna with the circular polarization are used.

3.5. Doppler effect.

Onset of Doppler effect due to motion of satellite relative to terrestrial station is one of special features of communication systems through ISZ. As a result of this at the point of reception/procedure the strain of the spectrum of signal and change in the carrier frequency is observed.

Change in carrier frequency will lead to the fact that level of received signal with insufficient width of band of filters Φ will be reduced, since value $b_{f\text{пр}}$ (or $b_{u/\text{пр}}$), entering expressions (3.1.2) and (3.1.4), will be changed.

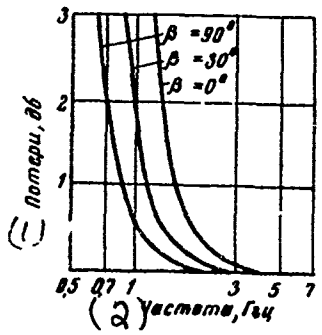

Fig. 3.4.9. Dependence of losses, caused by effect of Faraday.

Key: (1). Losses, db. (2). Frequency, GHz.

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Therefore width of band of filter must be chosen taking into account the Doppler effect; the latter must be taken into consideration and during the circuit design APCh of receiver.

Let us consider phenomena, caused by Doppler effect. It is known that during the motion of the source of signal to the side of reception/procedure with the speed v, the frequency f adopted is connected with the frequency of the emitted f_0 with the relationship/ratio

$$f = \frac{f_0}{1 - \frac{v}{c_0}}. \qquad (3.5.1)$$

Here $c_0 = 3.10^{10}$ cm/s speed of light.

Condition $\frac{v}{c_0} \ll 1$, usually always is satisfied therefore during

motion of signal to side of reception/procedure we will obtain

$$f = f_0\left(1 + \frac{v}{c_0}\right). \qquad (3.5.2)$$

In expression (3.5.2) it is taken into consideration, that velocity v of satellite is directed along radio communication link. In the general case, when velocity vector can be directed at any angle toward the radio communication link, the change in the frequency, caused by the Doppler effect, will be determined only by that component of velocity vector, which is parallel to radio communication link. Let us designate this component of rate through v_r and let us agree, that with a decrease in the distance between the source of signal and the point of the reception/procedure of component v_r has negative value, and with an increase in the distance of the component of velocity v_r indicated it is positive. Then from (3.5.2) we obtain the following expression for determining of change in frequency $\Delta f = f - f_0$:

$$\Delta f = -f_0 \frac{v_r}{c_0} = -\frac{v_r}{\lambda_0}, \qquad (3.5.3)$$

where $\lambda_0 = c_0/f_0$ - wavelength of the emitted oscillation/vibration.

From expression (3.5.3) it follows that for stationary satellites change in frequency, caused by Doppler effect, is equal to zero.

For evaluation/estimate of change in value Δf let us consider simplified case: satellite travels at constant speed $\overline{v}$ relative to

fixed point O. Let us replace the small phase of trajectory of the motion of the satellite of straight line. Then in accordance with Fig. 3.5.1, counting after t=0 moment of time, which corresponds to the shortest distance of r_0, we obtain

$$v_r = v \sin \Theta, \tag{3.5.4}$$

where

$$\sin \Theta = \pm \sqrt{\frac{r^2 - r_0^2}{r^2}}, \tag{3.5.5}$$

r_0 corresponds to the shortest distance.

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In expression (3.5.5) positive sign corresponds to values of t<0, and sign "minus" to values t>0. In accordance with (3.5.3) let us find

$$\Delta f_r = \pm \frac{v \sin \Theta}{\lambda_0} = \pm \frac{v}{\lambda_0 r} \sqrt{r^2 - r_0^2}. \tag{3.5.6}$$

Taking into account that according to Fig. 3.5.1 $\operatorname{tg} \Theta = \frac{vt}{r_0}$, from expression (3.5.6) we will obtain

$$\frac{\Delta f}{\frac{v}{\lambda_0}} = \pm \sin\left(\operatorname{arc\,tg} \frac{vt}{r_0}\right). \tag{3.5.7}$$

According to this relationship/ratio is constructed graph Fig. 3.5.2, on the basis of which, knowing velocity of motion v and wavelength λ_0, it is easy to determine value of Doppler frequency shift.

From latter/last expression and Fig. 3.5.2 it follows that rate

of change in frequency, caused by Doppler effect with uniform motion of satellite relative to observation point, will be greatest with $r=r_0$.

If relative to observation point satellite moves not with constant velocity, more complicated relationships/ratios are obtained. For the representation of calculation procedure the determination of a Doppler change in frequency [3.17] for the simplest case is given below. In this determination the orbit of ISZ is selected circular and the rotation of the Earth is not considered that it is considered permissible for ISZ, which move along the orbits with a comparatively low altitude.

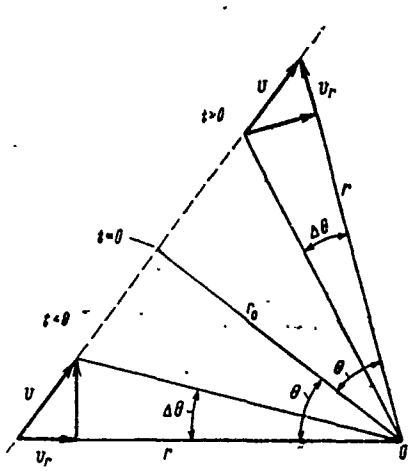

Fig. 3.5.1.

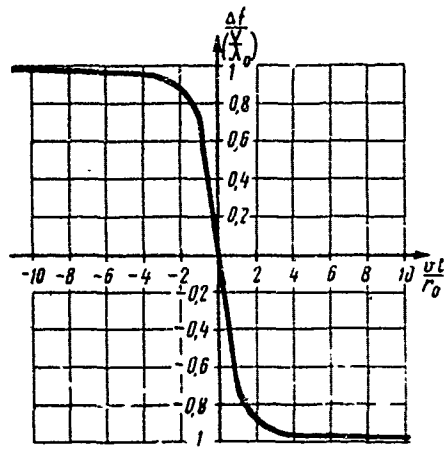

Fig. 3.5.2.

Fig. 3.5.1. To determination of Doppler change in frequency.

Fig. 3.5.2. Dependence of change in frequency on velocity of motion.

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Fig. 3.5.3 gives geometric constructions, necessary for calculations of Doppler change in frequency, emitted by satellite, moving along section of orbit TT, from which it follows

$$L^2 = R^2 + r^2 - 2Rr\cos D, \qquad (3.5.8)$$

where L - distance between satellite C and terrestrial station, which is located at point A;

r - distance from center of Earth to ISZ;

R - radius of Earth;

D - angle between lines of centers of Earth with ISZ and terrestrial station, which is located out of orbital plane.

After taking derivative for time from (3.5.8), we will obtain

$$LL' = rr' - Rr' \cos D + RrD' \sin D. \qquad (3.5.9)$$

For circular orbit r'=0; therefore expression (3.5.9) is converted in

$$L' = \frac{RrD' \sin D}{L}. \qquad (3.5.10)$$

Doppler shift according to (3.5.3)

$$\Delta f = -f_0 \frac{v_r}{c_0} = -f_0 \frac{L'}{c_0}. \qquad (3.5.11)$$

Substituting (3.5.10) in (3.5.11), we obtain

$$\Delta f = -f_0 \frac{RrD' \sin D}{c_0 L}. \qquad (3.5.12)$$

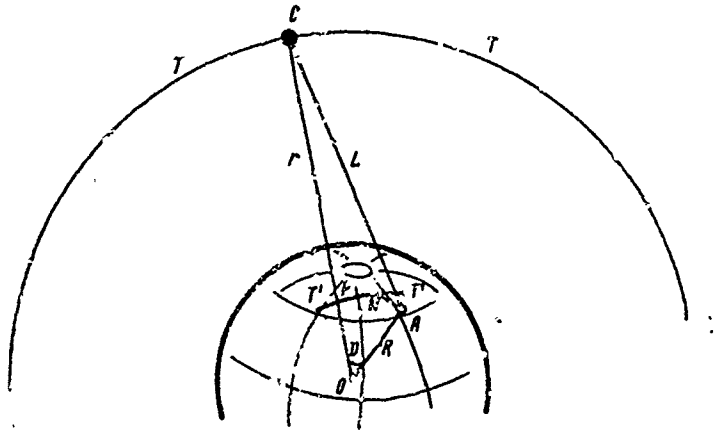

Fig. 3.5.3. On calculation of Doppler effect.

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Thus, for determining Doppler shift it suffices to know change in time of angle D. We will use the auxiliary drawing Fig. 3.5.4. Here arc T'T ' corresponds to the projection to the earth of the section of orbit TT, shown in Fig. 3.5.3, angle M determines the minimum angular distance between the projection of satellite and the point A, and points P and N - projection ISZ on the surface of the Earth.

From spherical triangle APN it follows that

$$\cos D = \cos\theta \cos M. \qquad (3.5.13)$$

It is obvious that during motion of satellite angle M remains constant/invariable; therefore time derivative of (3.5.13) is equal to

$$D' \sin D = \theta' \sin\theta \cos M. \qquad (3.5.14)$$

Since $\theta=\omega_3 t$, $\theta'=\omega_к$ where $\omega_к$ - angular satellite velocity. Therefore after substitution (3.5.14) in (3.5.12) we will obtain the value of Doppler frequency shift

$$\Delta f = -f_0 \frac{Rr\omega_к \cos M \sin \omega_к t}{c_0 L}. \qquad (3.5.15)$$

Let us note that to value of t=0 corresponds shortest distance of satellite. Thus, with t=0 in Fig. 3.5.4 projection of satellite to the earth will coincide with point N.

From (3.5.15) it follows that change in frequency will be large, when angle M to vanish, i.e., in the case, when projection of motion of satellite (arc T'T ' in Fig. 3.5.4) passes through observation point.

Let us consider motion of satellite along orbit 1 relative to terrestrial points/items A and B Fig. 3.5.5. At any point of orbit (for example, at point 1 or 2) the orbital satellite velocity can be decomposed on two orthogonal components, one of which coincides with the radio communication link, i.e., with the straight line, which connects terrestrial point/item with the point of the orbit, where is located satellite at the moment of time in question.

This resolution for points 1 and 2 orbits I (see Fig. 3.5.5) is shown in Fig. 3.5.5a and 3.5.6b respectively.

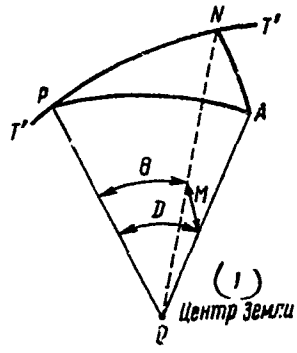

Fig. 3.5.4. Determination of angles.

Key: (1). Center of the earth.

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From the examination of Fig. 3.5.6a it follows, that complete total variation in the frequency on the line of communications at the moment of passage by the satellite of point 1 of orbit must be determined by a difference in the frequency drift, caused by the Doppler effect, for the interval the Earth-ISZ and the frequency drift in the interval an isz-Earth, i.e., according to (3.5.3)

$$\Delta f_{\text{поля}} = -\frac{|\bar{v}_{rA1}|}{\lambda_0} + \frac{|\bar{v}_{rB1}|}{\lambda_0}. \qquad (3.5.16)$$

When satellite is located at point 2 (see fig 3.5.6), components of velocity $\bar{v}_r$ have identical directions (in direction of removal/distance of satellites from points/items A and B).

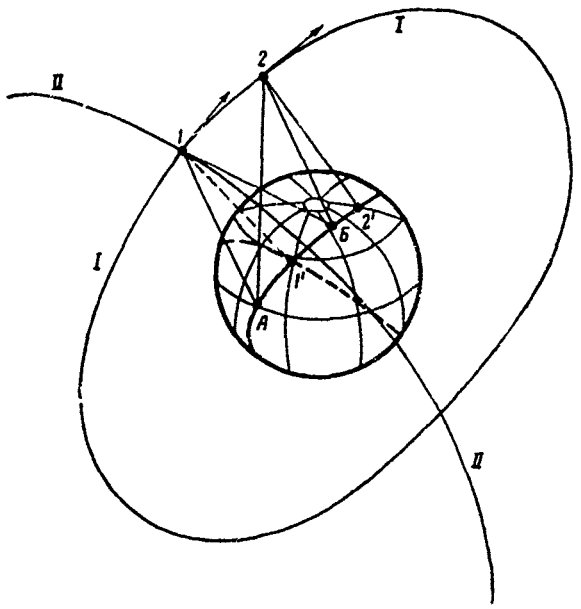

Fig. 3.5.5. To determination of change of frequency in communication system.

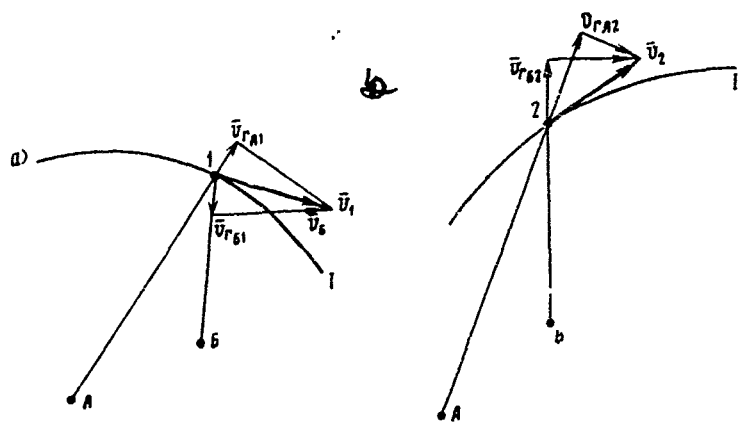

Fig. 3.5.6. Versions, which determine sum and difference in change of frequency in communication system.

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Therefore at the moment of passage by the satellite of point 2 total variation of the frequency in two intervals will be determined by the sum of frequency drift in the individual sections of the line of communications, i.e.,

$$\Delta f_{\text{полн}} = -\frac{|\bar{v}_{rA2}|}{\lambda_0} - \frac{|\bar{v}_{rB2}|}{\lambda_0}. \qquad (3.5.17)$$

Value of components v_{rA} and v_{rB} can be found from formulas (3.5.4) and (3.5.5).

Let us note that during motion of satellite along orbits, whose projections are arranged/located perpendicular to line, which connects points/items A and B on surface of Earth (for example, along orbit II in Fig. 3.5.5), total variation in frequency on route will be determined always by sum of change in frequency, caused by Doppler effect, in individual sections.

For solving question about that, will be full deflection of frequency in different cases determined by sum or difference in frequency deviations in individual sections, should be considered direction of motion of projection of satellite on surface of Earth relative to each ground station.

Thus, for example, in the case of moving satellite along orbit I (see Fig. 3.5.5) with the passage by the satellite of point its 1

projection - point 1 ' - will be moved in the direction from one point/item A to the next B. Thus, in this example the projection of satellite will recede from point/item A and approach point/item B, as a result of which the full deflection of frequency will be determined by a difference in the frequency deviations in the intervals of the line of communications in question [see expression (3.5.16)].

With passage by satellite of point its 2 projection on surface of Earth - point 2 ' - will recede from point/item A and point/item B; therefore full deflection of frequency will be determined by sum of frequency deviations in sections of line of communications [see expression (3.5.17)].

Doppler effect will lead not only to change in carrier frequency, as this follows from expressions (3.5.7) or (3.5.15), but also to strain of transmitted spectrum. For the explanation of the aforesaid let us consider three different forms of modulation.

We will consider that with OBP is transmitted signal, whose spectrum is limited by frequencies F_1 and F_2, moreover $F_1>F_2$. Then as a result of the Doppler effect at the point of reception/procedure instead of the frequencies indicated are accepted others, that differ from those transmitted in $(1\pm v/c_0)$ once.

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This means that the width of the spectrum, which in the place of transfer was in the place of reception/procedure it will prove to be

$$\Delta F = F_2 - F_1, \qquad (3.5.18)$$

in the place of reception/procedure it will prove to be deformed and equal to

$$\Delta F' = F_2\left(1 \pm \frac{v}{c_0}\right) - F_1\left(1 \pm \frac{v}{c_0}\right) = \Delta F\left(1 \pm \frac{v}{c_0}\right). \qquad (3.5.19)$$

Let us consider now case ChM by sinusoidal oscillation with frequency of F with low index of modulation. Then it is possible to consider that the modulated signal consists of oscillations with a carrier frequency of f_0 and oscillations with lower and upper lateral frequencies of f_0+F and f_0+F_0. As a result of the effect of the Doppler effect the frequencies of the modulated signal will take the following values:

- the carrier frequency

$$f_0\left(1 \pm \frac{v}{c_0}\right) = f_0 \pm f_0 \frac{v}{c_0} = f_0 \pm \Delta f_0, \qquad (3.5.20)$$

- the lower lateral

$$(f_0 - F)\left(1 \pm \frac{v}{c_0}\right) = f_0 - F \pm f_0 \frac{v}{c_0} \mp F \frac{v}{c_0} = f_0 \pm \Delta f_0 - \left(F \pm F \frac{v}{c_0}\right),$$

- the upper lateral

$$(f_0 + F)\left(1 \pm \frac{v}{c_0}\right) = f_0 \pm \Delta f_0 + \left(F \pm F \frac{v}{c_0}\right).$$

From these expressions it follows that difference between carrier frequency and any of lateral as a result of effect of Doppler effect is equal to $F(1+v/c_0)$. Therefore after detection we will

obtain a change of the modulating frequency into the number of times, equal to $(1\pm v/c_0)$. If we consider ChM with large the index of modulation, and also with modulation serrated signal it is possible to make the conclusions that Doppler's phenomenon will lead to a change in the modulating frequencies and width of the spectrum in $(1\pm v/c_0)$ once. These conclusions/outputs are valid also for AM and FM.

Examination of such forms of pulse modulation as amplitude, latitudinal or phase AIM, ShIM and FIM), shows that also in these cases effect Doppler effect leads to strain of spectrum of transmitted signal and produces change of modulating frequencies in $(1\pm v/c_0)$ once for all components of spectrum (including for clock frequency, which determines sequence of unmodulated sequence of pulses/momenta).

One should note that in communication systems through ISZ value v_0/c is equal to approximately 10^{-5} and it is less.

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Described change in spectrum of signal, which is developed as a result of Doppler effect for enumerated forms of modulation, means that duration in time of transmitted signal will be changed in $(1\pm v/c_0)$ times. This means, for example that the duration of moving picture at the point of reception/procedure will prove to be

different, than at the point of transfer. Difference in the duration, certainly, will be very low and in the case of the transfer of moving picture virtually by anyone will not be noticed; however, during the transfer, for example, the impulses/transmissions of standard duration or standard frequencies this phenomenon it is necessary to consider.

With KIM Doppler effect will produce change in clock frequency.

For decreasing effect of Doppler shift of carrier frequency it is possible to use two methods. In the first of them in the receivers is used APCh and installation at the input of sufficiently broadband filters. The second possible method consists in the fact that with the previously known parameters of orbit IZS and the coordinates of terrestrial station it is possible to preliminarily compute the dependence of Doppler shift in the time and to realize the appropriate shift of the carrier frequency of transmitter. In this case, obviously, it is necessary to correspondingly expand the filter pass band of transmitter.

Let us note that any of methods described above is not removed strain of spectrum of signal.

3.6. Efficiency of feeders.

Ratio of active power, which is isolated in load at end/lead of feeder, is called efficiency of feeder η_ϕ to total active power, conducted/supplied to its beginning. The efficiency of feeder it is possible to express through the traveling-wave ratio kbv on the voltage/stress (κ), the linear attenuation and Np/m and the length feeder l m as follows [3.5]:

$$\eta_\phi = \frac{1}{\operatorname{ch} 2\alpha l + 0{,}5\left(\kappa + \frac{1}{\kappa}\right)\operatorname{sh} 2\alpha l}. \qquad (3.6.1)$$

If $2\alpha l \ll 1$, then it is possible to count $\operatorname{sh}2\alpha l \approx 2\alpha l$ and $\operatorname{ch}2\alpha l \approx 1$; in this case

$$\eta_\phi = \frac{1}{1 + \left(\kappa + \frac{1}{\kappa}\right)\alpha l}. \qquad (3.6.2)$$

Dependence η_ϕ on value αl with different values kbv. constructed according to expression (3.6.1), is given in Fig. 3.6.1.

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Let us note that the value of kbv on the voltage/stress is connected with the coefficient of reflection p with the following dependence:

$$\kappa = \frac{1 - |\rho|}{1 + |\rho|}. \qquad (3.6.3)$$

For rectangular waveguide with usually utilized wave $H_{0,1}$ decay constant can be determined according to following expression [3.5]:

$$\alpha = \frac{\sqrt{\frac{\pi f \mu}{\gamma}}\left[1+2\frac{b}{a}\left(\frac{\lambda_0}{2a}\right)^2\right]}{bW\sqrt{1-\left(\frac{\lambda_0}{2a}\right)^2}}, \text{ Np/m. (3.6.4).}$$

Here λ_0 - wavelength free space, m;

γ - conductivity of metal, S/m (for copper $\gamma=5.8\cdot10^7$ S/m);

b and a - height/altitude and the width of the cross section of waveguide, m;

f - frequency, Hz;

μ - magnetic permeability, H/m, moreover for the free space $\mu=\mu_0=4\pi10^{-7}$ H/m.

For circular waveguide during use of wave H_{11} decay constant can be found from expression

$$\alpha = \frac{\sqrt{\frac{\pi f \mu}{\gamma}}}{aW\sqrt{1-0{,}0087\left(\frac{\lambda}{a}\right)^2}}\left[0{,}087\left(\frac{\lambda}{a}\right)^2+0{,}43\right], \frac{\text{неп}}{\text{м}}. \quad (3.6.5)$$

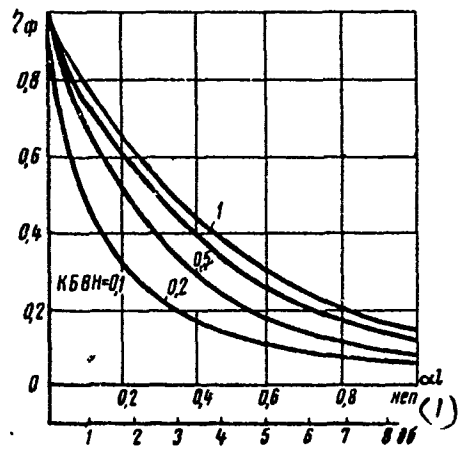

Fig. 3.6.1. Dependence of efficiency of feeder on αl with different KBVN.

Key: (1). Np.

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Here a - radius or waveguide, m.

$$W=120\ \pi.$$

Remaining designations coincide with those accepted for expression (3.6.4). Expressions (3.6.4) and (3.6.5) are derived under the assumption that the conductivity of the walls of waveguide does not have a noticeable effect on the structure of field in it. This assumption is justified in such a case, when waveguide is made from the materials, which have high conductivity in the range of shf/SVCh

(copper, brass, aluminum, silver, etc.).

Formulas (3.6.4) and (3.6.5) do not consider roughness of walls of waveguides. It is obvious that its account will lead to an increase of the decay constant, as a result of which the real value of coefficient α will differ from the values, calculated according to formulas (3.6.4) and (3.6.5), on 10^{-3}-$5 \cdot 10^{-4}$ Np/m.

In number of cases coefficients α are expressed in decibels, in reference to feeder by length of 100 m. Taking into account that 1 nep=8.68 db, it is possible to obtain the following relationship/ratio: α[Np/m]=$8.68 \cdot 10^{-2}\alpha'$; here α' has the dimensionality of db/100 m.

Table 3.6.1 gives values of linear attenuation of some waveguides, calculated with λ=8 cm [3.18].

Table 3.6.1.

(1) Материал	(2) Проводимость по постоянному току γ, $\left(\frac{сим}{м}\right)$	(3) Погонное затухание $\alpha'\left(\frac{дб}{100\ м}\right)$ при λ=8 см, для волноводов			(4) Глубина проникновения тока (мм)
		58×25 мм H_{01}	72×34 мм H_{01}	Ø70 мм H_{11}	
Медь (5)	$5,8\cdot10^7$	2,86	1,68	1,25	$1,1\cdot10^{-3}$
Алюминий (6)	$3,48\cdot10^7$	3,68	2,17	1,62	$1,41\cdot10^{-3}$
Латунь с содержанием меди (7)					
96%	$4,07\cdot10^7$	3,4	2,0	1,49	$1,31\cdot10^{-3}$
90%	$2,52\cdot10^7$	4,34	2,55	1,89	$1,67\cdot10^{-3}$
70%	$1,65\cdot10$	5,38	3,16	2,32	$2,07\cdot10^{-3}$
60%	$1,51\cdot10^7$	5,6	3,3	2,44	$2,14\cdot10^{-3}$
Серебро (8)	$6,27\cdot10^7$	2,72	1,6	1,2	$1,05\cdot10^{-3}$

Key: (1). Material. (2). Conductivity for direct current γ, (S/m). (3). Linear attenuation α'(db/100 m) with λ=8 cm, for waveguides. (4). Depth of penetration of current (mm). (5). Copper. (6). Aluminum. (7). Brass with contenc of copper. (8). Silver.

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Chapter 4.

NOISE LEVEL AT THE INPUT OF A RECEIVER.

4.1. Initial positions.

The noise level at the input of a receiver is defined both by the value of its own thermal noises of the receiver $P_{шпр}$ and by intensity of noises of different sources and circuits external with respect to the receiver. Ambient noises include: the radio emission of the atmosphere, which has thermal character, thermal noises of the Earth and antenna, and also thermal noises created by different circuits connected to the input of the receiver (by feeders, by filters, etc.). Furthermore, considerable noise level at the input of a receiver can be created by extraterrestrial sources - radio emissions of the sun, moon, and planets and by space radio emissions.

Radio emissions of the sun, which, as show radio-astronomical observations, in remaining on the average constants, have both short-term periods of an increase in level during 1-2 min (repeating after 20-30 min) and prolonged periods of an increase in radiation

level, are most intense. The latter are approximately ten days per annum in the per ds of the maxima of solar activity, repeating, approximately, in 11 years.

Cosmic radio-frequency radiation is determined by two forms of sources: by sources, which are located within limits of our galaxy (galactic radio emission), and by sources, which are located beyond its limits (metagalactic radio emission). Cosmic rad.o-frequency radiation in the space is distributed unevenly, moreover angular dimensions and intensity of the emission of the sources, observed from the Earth, prove to be different: together with the very extended sources with the low power of radio emission it is possible to observe comparatively powerful/thick sources, whose sizes/dimensions comprise several angular minutes (Cassiopeia-A, Cygnus-A, etc.).

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By nature galactic radio emission can be subdivided into two groups: to thermal and nonthermal.

Nonthermal radio emission is determined by oscillations/vibrations of inhomogeneous plasma and by emission of relativistic electrons in presence of magnetic field. Nonthermal radio emission distributed in outer space almost evenly and predominates on the longer waves than thermal (approximately

increasing with an increase in the wavelength proportionally $\lambda^{0,3\div0,8}$).

Thermal radio emission predominates on waves of shorter than 1-1.5 m and virtually it does not depend on frequency.

Solar noise, Moon, planets and discrete/digital sources, distributed cosmic radio-frequency radiation they are characterized by continuous spectrum and it is very weakly polarized.

Knowing character of motion of satellite relative to sources of cosmic radio-frequency radiation, it is possible to previously predict moments of time, when the receiving antenna proves to be simultaneously directed toward the source of the signal and source of cosmic radio-frequency radiation. In this case the ratio of the received signal to the total value of noises must not be lower than the certain assigned magnitude, determined by the parameters of the system of communications (by form of modulation, with the power of transmitters, by antenna gain, by receiver circuit, etc.). This in the equal measure relates both to the section of communication the Earth-ISZ and to the section of the ISZ-Earth. On the basis of the given considerations it is possible to consider that the total power of the noises at the input of the receiver is determined by the following components:

$$P_{ш} = P_{ш\,пр} + P_{А} + P_{Ф} + P_{косм}. \qquad (4.1.1)$$

Here $P_{ш\,пр}$ - inherent noise of the receiver;

P_A - antenna noises taking into account thermal noises of atmosphere, noises of Earth and effect of radome;

P_Φ - noises, created by feeder and other circuits, connected to the input of the receiver;

$P_{космш}$ - noises created by radio emissions of the sun, moon and planets and by space sources.

In subsequent sections of this chapter the determination of the values and enumerated components of the noises is examined.

4.2. Noise temperature of quadrupoles and receivers.

Noisiness of quadrupoles it is accepted to characterize by noise factor or by efficient noise temperature.

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Ratio of total power of noise at the output of the quadrupole, determined in a single frequency band, is called the factor of noise of quadrupole at certain frequency to power, caused by noises of input resistance, at standard noise temperature, which usually take as equal to $T_0=290°K$ [4.1]. Thus, considering that the transmission factor of quadrupole according to the power is equal to M, in accordance with the determination we obtain

$$Ш = \frac{P'_{ш\,вых}}{kT_0 M \Delta f}. \qquad (4.2.1)$$

Here $k=1.38\cdot10^{-23}$ J/deg - Boltzmann constant.

Let us establish the dependence between factor of noise and efficient noise temperature of quadrupole. If on the input resistance of the quadrupole with the transmission factor M a certain source in the band of frequencies Δf there develops the nominal power of the noise

$$P'_{ш\,вх}=kT_0\Delta f, \qquad (4.2.2)$$

then at the output of the quadrupole in the same frequency band the nominal power of noise $P'_{ш\,вых}$ will be equal to the sum of the power of noises, obtained during the amplification of noises, subject from source $P'_{ш\,вх}$, and the inherent noise of quadrupole $P_{ш\,вых}$. Thus, $P'_{ш\,вых}=P'_{ш\,вх}M+P_{ш\,вых}$. Hence follows that the power of the inherent noise of quadrupole, in reference to its input, upon consideration of (4.2.1) and (4.2.2), can be expressed thus:

$$P^{*}_{ш\,вх}=\frac{P_{ш\,вых}}{K\,M}=\frac{P'_{ш\,вых}}{M}-P'_{ш\,вх}=kT_0\Delta fШ-kT_0\Delta f=kT_0\Delta f(Ш-1). \qquad (4.2.3)$$

Quadrupole examined with respect to noises can be substituted by certain two-terminal network, which has such efficient noise temperature $T_ш$, at which two-terminal network on matched load in frequency band, equal band of frequencies of quadrupole, develops power $P_ш$, equal to power of inherent noise of quadrupole, converted

to its input. Keeping in mind, that for the two-terminal network

$$P_{ш.вх} = kT_0\Delta f \qquad (4.2.4)$$

and, equalizing (4.2.3) and (4.2.4), let us find

$$T_ш = T_0(Ш - 1). \qquad (4.2.5)$$

Hence

$$Ш = 1 + \frac{T_ш}{T_0}. \qquad (4.2.6)$$

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From expression (4.2.5) it follows that efficient noise temperature, equal to zero, corresponds to noise factor, equal to one. Let us note that into expression (4.2.5), just as into expression (4.2.1), the noise factor enters together with temperature T_0, standard for determining the noise factor. Without the indication of this temperature (usually it it is taken as the equal to 290°K) the value of noise factor does not have a sense.

Let us note that in all given considerations it was assumed that factor of noise Ш and value of transmission factor of quadrupole according to power M in band of frequencies Δf in question were constant/invariable and identical with any value of frequency f, which lies at band Δf.

In examination of quadrupoles very wide passband it is necessary to consider with fact that values M and Ш, found for the maximally narrow bands of frequencies df, which lie within the cut-off

frequencies f_1 and f_2, can prove to be different (Fig. 4.2.1). Thus, in the general case these values can be the functions of frequency, i.e., M=M(f) and Ш=Ш(f), where value f is taken within the band of frequencies $f_1 \leq f \leq f_2$ in question.

It is obvious that in general case power of noises at the output of the quadrupole in the band of frequencies $\Delta f=f_2-f_1$ will be determined by expression

$$P'_{\text{ш вых}} = kT_0 \int_{f_1}^{f_2} Ш(f) M(f)\, df. \qquad (4.2.7)$$

Let us note that the concept "broad band" of transmission of quadrupole implies that difference f_2-f_1 proves to be of the same order as maximum cut-off frequency of f_2. In practice frequently it is necessary to deal concerning the selective networks, which have a comparatively narrow passband, on boundaries of which with values of $f=f_1$ and $f=f_2$ value $M(f)\approx 0$ (Fig. 4.2.2). In this case the concept "narrow band" means that

$$|f_2-f_1| \ll f_1. \qquad (4.2.8)$$

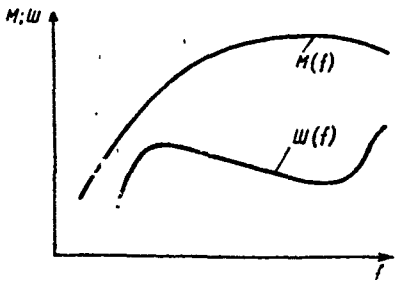

Fig. 4.2.1.

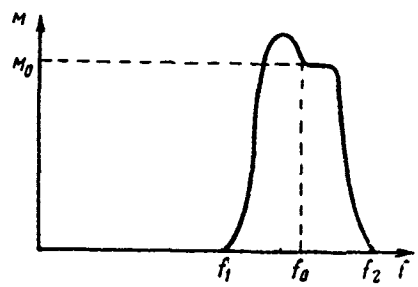

Fig. 4.2.2.

Fig. 4.2.1. Derivation of formula 4.2.7.

Fig. 4.2.2. Derivation of formula 4.2.9.

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Specifically, to such narrow-band quadrupoles, which satisfy conditions (4.2.8), the input and tuned amplifiers of radio receiving equipment can be referred.

With satisfaction of condition (4.2.8) usually it is possible to consider that in entire band of frequencies, where M(f) noticeably differs from zero, factor of noise Ш(f) proves to be virtually constant/invariable and therefore it can be carried out as integral sign. Consequently, instead of (4.2.7) we will obtain

$$P_{\text{ш max}} = kT_0 Ш \int_{f_1}^{f_2} M(f)\,df = kT_0 Ш \int_0^{\infty} M(f)\,df. \qquad (4.2.9)$$

A change in integration limits is determined by the fact that with $f<f_1$ or $f>f_2$, usually value $M(f)=0$. Introducing the concept of the amplitude-frequency characteristic F(f) of quadrupole, it is

possible to register

$$M(f) = M_0 F^2(f), \qquad (4.2.10)$$

where M_0 - transmission factor of quadrupole according to the power, found for the certain "central" frequency f_0. Value

$$F^2(f) = \frac{M(f)}{M_0} \qquad (4.2.11)$$

characterizes the ratio of the power of oscillations, removed from the output of the quadrupole during the successive feed to its input of identical with respect to value of voltages to frequency f and to a "central" frequency of f_0. Then after substitution (4.2.10) in (4.2.9) we obtain

$$P_{\text{вх max}} = kT_0 Ш \Delta f^* M_0, \qquad (4.2.12)$$

where

$$\Delta f^* = \int_0^\infty F^2(f)\, df. \qquad (4.2.13)$$

Expression (4.2.13) determines the value of resulting noise bandwidth Δf^*, which is equal to the foundation of rectangle with single ordinate. The area of this rectangle is equivalent the area, formed by the axis of abscissas and curved, obtained by plotting the dependence of the square of the relation of ordinates, the amplitude-frequency characteristic of quadrupole at frequencies f and f_0 (Fig. 4.2.3).

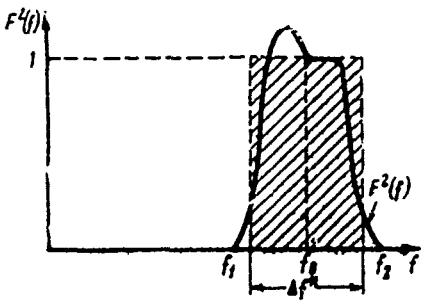

Fig. 4.2.3. Determination of noise bandwidth.

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It is possible to show [4.2] that if we serially connect n amplifiers (quadrupoles), each of which is characterized by its own factor of noise $Ш_m$ and by transmission factor according to power at "central" frequency $M_m \geqslant 1$ (index m it indicates reference number of one of the n amplifiers), then the common noise factor is connected with factor of noise of separate amplifiers by the following ratio:

$$Ш = Ш_1 + \frac{Ш_2 - 1}{M_1} \frac{\Delta f^*(2,3,\ldots,n)}{\Delta f_n} + \\ + \frac{Ш_3 - 1}{M_1 M_2} \frac{\Delta f^*(3,4,\ldots,n)}{\Delta f_n} + \ldots + \frac{Ш_n - 1}{M_1 M_2 \ldots M_{n-1}} \frac{\Delta f^*(n)}{\Delta f_n}. \quad (4.2.14)$$

Here Δf^* (m, m+1,..., n) is determined resulting noise bandwidth of series-connected amplifiers (quadrupoles), beginning from the number m; Δf_n - resulting noise bandwidth of all n series-connected amplifiers (quadrupoles).

Let us establish connection/communication between resulting noise bandwidth of the n-cascade amplifier and passband of this

amplifier $\Delta f_{0,7}$, which is determined it is determined on detuning, which corresponds to decay in amplitude-frequency characteristic to level 0.7 from maximum value. It is obvious that in accordance with (4.2.13) the noise bandwidth of the n-cascade amplifier

$$\Delta f_n^* = \int_0^\infty F_n^2(f)\,df.$$

Results of calculating [4.3] this integral for tuned amplifiers with circuits tuned for one frequency and with band circuits are given in Table 4.2.1.

Table 4.2.1.

Число каскадов (1)	1	2	3	4	5	6	∞
Отношение $\frac{\Delta f_n^*}{\Delta f_{0,7}}$ для резонансных контуров (2)	1,57	1,22	1,16	1,13	1,11	1,10	1,06
Отношение $\frac{\Delta f_n^*}{\Delta f_{0,7}}$ для полосовых контуров (3)	1,11	1,04	1,02	1,014	1,00	1,00	1,00

Key: (1). Number of cascades. (2). Relation ... for resonant circuits. (3). Ratio ... for band circuits.

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From Table 4.2.1 it follows that when n≥4 with an accuracy sufficient for practical calculations, it is possible to consider that $\Delta f_n^* = \Delta f_{0,7}$, that noise bandwidth of multistage amplifier is equal to passband of amplifier.

Let us turn to an examination of expression (4.2.14). Since for amplifiers $M_m > 1$, from (4.2.14) it follows that:

1) the factors of the noise of the last amplifier stages affect the common noise factor considerably less than the factors of the noise of the first cascades;

2) for reduction in the common noise factor it is desirable so that first stages of the amplifier would have largers possible

factors of amplification and narrower frequency band so that the resulting band of frequencies of the network of amplifiers would be determined precisely by the first cascades.

Let us note that in the literature relationship/ratio (4.2.14) sometimes is given in the form

$$\text{Ш} = \text{Ш}_1 + \frac{\text{Ш}_2 - 1}{M_1} + \frac{\text{Ш}_3 - 1}{M_1 M_2} + \ldots + \frac{\text{Ш}_n - 1}{M_1 M_2 \ldots M_{n-1}}. \quad (4.2.15)$$

From a comparison of the last expression (4.2.14), it follows that this recording is valid when

$$\frac{\Delta f^*(2,3,\ldots,n)}{\Delta f_n} = \frac{\Delta f^*(3,4,\ldots,n)}{\Delta f_n} = \ldots = \frac{\Delta f^*(n)}{\Delta f_n} = 1.$$

This is possible if all series-connected amplifiers have an identical, strictly rectangular amplitude-frequency characteristic.

When the band of frequencies of the amplifiers is determined by second cascade/stage, i.e.,

$$\Delta f^*(2,3,\ldots,n) \approx \Delta f,$$

and the amplification factors are so great that in expression (4.2.14) it is possible to disregard all terms beginning from the third, we will obtain

$$\text{Ш} \approx \text{Ш}_1 + \frac{\text{Ш}_2 - 1}{M_1}. \quad (4.2.16)$$

For calculating the efficient noise temperature of

series-connected amplifiers we will use expression (4.2.6), after registering it in the form:

$$Ш = 1 + \frac{T_э}{T_0},$$
$$Ш_1 = 1 + \frac{T_{э1}}{T_0},$$
$$Ш_n = 1 + \frac{T_{эn}}{T_0}.$$

Then with substitution into (4.2.14) after conversions we obtain

$$T_э = T_{э1} + \frac{T_{э2}}{M_1}\frac{\Delta f^*(2,3,\ldots,n)}{\Delta f} + \frac{T_{э3}}{M_1 M_2}\frac{\Delta f^*(3,4,\ldots,n)}{\Delta f} + \ldots \quad (4.2.17)$$

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Hence follows that total efficient noise temperature is determined to a considerable degree by value of transmission factors according to power of first stages of amplifier.

In the case of satisfaction of conditions, formulated in obtaining of expression (4.2.16), we find

$$T_э = T_{э1} + \frac{T_{э2}}{M_1}. \quad (4.2.18)$$

Values of $T_{э1}$, $T_{э2}$ and M_1 are determined by schematic of first amplifier stages. The efficient noise temperature of radio receiving equipment in the dependence on the diagrams of the first cascades/stages can take the different values, which differ into hundreds of times. Fig. 4.2.4 gives approximate values $T_э$ for receivers with different input cascades.

One should note, that low-noise input cascades are complicated in manufacture and operation. Therefore, the selection of the receiver, for example, with the laser, must precede the technical-economic comparison of this receiver with other possible versions. At the same time the selection of the diagram of the input device of the receiver must be determined by gain in the value of total noises. Thus, for example, in accordance with expression (4.1.1), the installation of the low-noise input devices of the receiver (for example, laser) is appropriate only then, when the sum of the remaining components of noises at the input of the receiver $(P_A + P_\Phi + P_{косм})$ proves to be of the same order as the inherent noise of the receiver. The case, when $P_A \pm P_\Phi + P_{косм} \gg P_{ш\,пр}$, characterizes the incorrect selection of the diagram of the input device of the receiver, which causes excessive complication and rise in price of the receiver. In this case the transition/junction to the simpler input device is advisable, although the amount of the total power of noises at the input of the receiver in this case somewhat will increase.

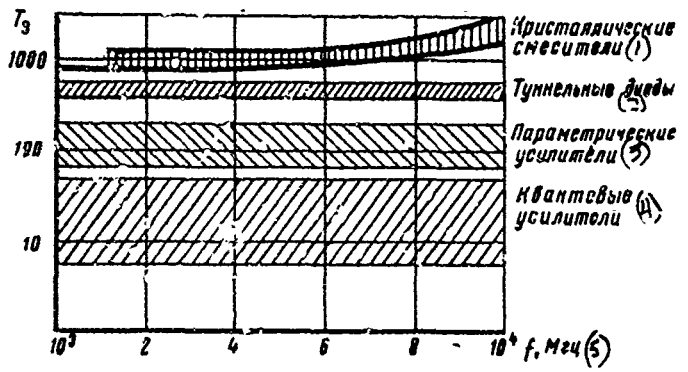

Fig. 4.2.4. Values $T_ш$ of receivers.

Key: (1). Crystal mixers. (2). Tunnel diodes. (3). Parametric amplifiers. (4). Lasers. (5). MHz.

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Let us consider now the case of series connection of attenuator, i.e., quadrupole with transmission factor according to power $M_{ат}<1$, or with coefficient of weakening

$$B_{ат}=\frac{1}{M_{ат}}>1. \qquad (4.2.19)$$

Let us represent attenuator in the form of overall equivalent diagram, formed by the T-connection of resistors of R_1, R_2 and R_3 (Fig. 4.2.5), coordinated with the input resistance $R_{вх}$. If the weakening due to losses in the attenuator is equal to $B_{ат}$, then under the matching condition the value of resistances is determined as follows:

$$\left.\begin{aligned} R_1&=R_{вх}\frac{B_{ат}+1}{B_{ат}-1}-R_3\\ R_2&=R_{вых}\frac{B_{ат}+1}{B_{ат}-1}-R_3\\ R_3&=\frac{2\sqrt{B_{ат}R_{вх}R_{вых}}}{B_{ат}-1}\end{aligned}\right\}. \qquad (4.2.20)$$

We will consider that the attenuator has a temperature $T_{ат}$, i.e., each of the resistors of R_1, R_2, R_3 is heated to temperature $T_{ат}$ and, consequently, for each of them certain noise emf

$$e_k = 4kT_{ат}\Delta f R_k, \qquad (4.2.21)$$

where index k determines index of resistors/resistances of attenuator, is created.

In this case power of noises at the output of the attenuator $P_{ш\,вых}$ will be:

$$P_{ш\,вых} = \frac{e_1^2 R_3^2 R_{вых} + e_2^2 (R_{вх} + R_1 + R_3)^2 R_{вых} + e_3^2 (R_{вх} + R_1)^2 R_{вых}}{[(R_{вх} + R_1 + R_3)(R_{вых} + R_3 + R_3) - R_3^2]^2} =$$

$$= 4kT_{ат} R_{вых} \Delta f \frac{R_1 R_3^2 + R_3 (R_{вх} + R_1 + R_3) + R_3 (R_{вх} + R_1)^2}{[(R_{вх} + R_1 + R_3)(R_{вых} + R_3 + R_3) - R_3^2]}.$$

After substitution of expressions (4.2.20) and conversions

$$P_{ш\,вых} = kT_{ат}\Delta f \frac{B_{ат} - 1}{B_{ат}} = kT_{ат}\Delta f (1 - M_{ат}). \qquad (4.2.22)$$

Let us find the efficient noise temperature of matched attenuator in the absence at its input of signals. For this we use the already used previously procedure: we will consider that power $P_{ш\,вых}$ is equal to the power of noises $P_ш$ of a certain resistor/resistance, heated to temperature $T_{э\,вых}$. i.e.

$$P_ш = kT_{э\,вых}\Delta f.$$

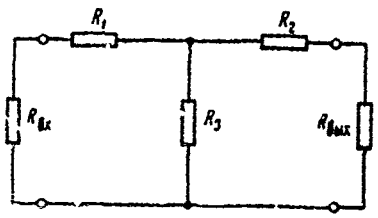

Fig. 4.2.5. Equivalent diagram of attenuator.

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Equalizing this expression to value $P_{ш\,вых}$, found from (4.2.22), we will obtain

$$T_{э\,вых} = T_{ат}\frac{B_{ат}-1}{B_{ат}} = T_{ат}(1-M_{ат}). \qquad (4.2.23)$$

From this expression it follows that when transmission factor of attenuator $M_{ат}\to 0$ (coefficient of weakening $B_{ат}\to\infty$), effective temperature at its output proves to be equal to temperature, to which it is heated. When the attenuator does not introduce the attenuation ($B_{ат}\to 1$), its effective temperature becomes equal to zero.

Let us note that expressions (4.2.22) and (4.2.23) are identical to results, obtained in [4.3] by another method.

In number of cases it is necessary to know effective temperature of the attenuator, in reference to its input. Let us fulfill for this the appropriate conversions, taking into account that transmission factor $M_{ат}$ and the coefficient of weakening of the attenuator are

connected by the relation (4.2.19).

It is obvious that the power of inherent noise of the attenuator, converted to its input, can be determined upon consideration of expression (4.2.22) as follows:

$$P_{ш\,вх}=\frac{P_{ш\,вых}}{M_{ат}}=kT_{ат}\Delta f(B_{ат}-1)=kT_{ат}\Delta f\left(\frac{1-M_{ат}}{M_{ат}}\right).$$

Introducing into the examination the two-terminal network, equivalent according to power of noise with effective temperature $T_{э\,вх}$ equating, let us find

$$T_{э\,вх}=T_{ат}(B_{ат}-1)=T_{ат}\left(\frac{1-M_{ат}}{M_{ат}}\right). \qquad (4.2.24)$$

On the basis of expression (4.2.24) it is possible to draw conclusion that switching on/inclusion of attenuator leads to an increase in the efficient inlet temperature.

We use expression (4.2.24) for calculating efficient noise temperature of feeder, which combines the antenna with the receiver. Keeping in mind that the value of efficiency of the feeder, introduced in Section 3.6, is connected with the coefficient of division with the relationship/ratio

$$B_{ат}=\frac{1}{\eta},$$

upon consideration of expressions (4.2.24) and (3.6.1) with the replacement of subscripts "ат" (attenuator) by "ф" (feeder), we will

obtain

$$\frac{T_{э\,вх}}{T_{ф}} = \operatorname{ch} 2\alpha l + 0{,}5\left(k + \frac{1}{k}\right)\operatorname{sh} 2\alpha l - 1. \qquad (4.2.25)$$

In this expression $T_{э\,вх}$ determines the efficient noise temperature of the feeder, converted to its input, and $T_{ф}$ - physical temperature of the feeder. In Fig. 4.2.6 in accordance with expression (4.2.25) the graphic dependences with the different values of the кбв [kbv - travelling-wave ratio] are constructed.

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From expressions (4.2.24) and (4.2.25) it follows that switching on/inclusion of attenuator (or feeder with certain attenuation) unavoidably leads to increase in effective temperature, converted to its input. In this case, if temperature $T_{ат}$=290°K when $B_{ат}$=1,26 (1 dB), efficient noise inlet temperature $T_{э\,вх}$=75,5°K; when $B_{ат}$=1,023 (0.1 dB) — $T_{э\,вх}$=6,7°K.

Thus, introduction of attenuation to 1 dB leads to an increase in the inlet temperature of 75.5°K, and each tenth of a decibel attenuation increases the inlet temperature by approximately 7°K. Hence it follows that if the noise temperature of the antenna, directed to a certain point of the firmament, is equal to 20°K, then in the case when the antenna is connected with the receiver by feeder with k=1 and attenuation 1 dB, this will be equivalent so that at the input of the feeder (at the output of the antenna) the total

efficient noise temperature will be 20°+75.5°=95.5°K. But if between the antenna and the receiver feeder will be without the attenuation, efficient noise inlet temperature of it will be 20°K. Hence it follows that the attenuation of the feeder will lead to a considerable increase in the value of noises at its input (in the examined case 95.5/20≈5 times).

4.3. Intensity of the emission of space.

With the reception of signals emitted by the moving communication satellite, terrestrial antenna system in process of "tracking" satellite must be directed for different regions of celestial sphere. Consequently, picking up signal, emitted satellite, antenna will accept the radio emissions of different space sources, which have both discrete/digital, and distributed character. Space sources usually characterize by using concepts of the spectral density of flow and spectral intensity of radio emission.

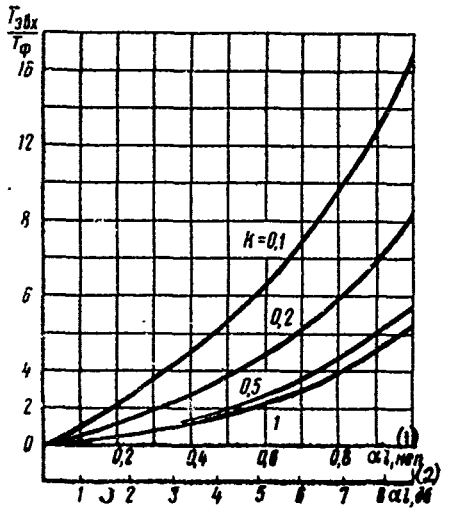

Fig. 4.2.6. Change in $T_э$ at the input of the feeder with different attenuation and кбв.

Key: (1). Np. (2). dB.

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Spectral density of flux of radio emission Φ is determined by the power emitted by the source in single frequency band and passing through single area/site in direction of normal to this area/site. The spectral intensity J or the brightness of the source of radio emission defines as the limit of the spectral density of power flux, in reference to infinitesimal solid angle. Therefore J is connected with the flow Φ by the dependence

$$J = \lim_{\Delta\Omega \to 0} \frac{\Delta\Phi}{\Delta\Omega}, \quad \text{W/m}^2 \cdot \text{Hz} \cdot \text{sr}, \qquad (4.3.1)$$

where $\Delta\Omega$ - solid angle (in the steradians). From this expression it

follows

$$\Phi = \int_{\Omega_u} J d\Omega, \qquad (4.3.2)$$

where Ω_u - solid angle of source.

In number of cases to more conveniently use concept of brightness temperature T_B, which is equivalent temperature of blackbody (°K) of the same geometry, as observed source, and having at this frequency in this direction the same spectral intensity. The functional dependence of the temperature of blackbody and spectral intensity is determined by Planck law:

$$J = \frac{2hc_0}{\lambda^3} \frac{1}{e^{\frac{hc_0}{k\lambda T_B}} - 1}, \quad \text{Т/м}^2\text{ср}.$$

Hence, knowing J, it is possible to find temperatures T_B of blackbody. Here h - Planck's constant; $k=1.38\cdot10^{-23}$ J/deg - Boltzmann constant; λ - wavelength of the radio emission adopted; $c_0=3\cdot10^{10}$ cm/s - speed of light.

In radio-frequency range usually $\frac{hc_0}{\lambda} \ll kT_B$ and spectral intensity of blackbody can be described by approximate expression of Rayleigh-Jeans:

$$J = \frac{2kT_B}{\lambda^2}. \qquad (4.3.3)$$

Taking into account (4.3.2), let us find that the flow in the limits

of solid angle Ω_A at a uniform temperature of source will be

$$\Phi = J\Omega_A = \frac{2kT_A}{\lambda^2}\Omega_A. \qquad (4.3.4)$$

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Let us determine power on matched termination of ideal antenna A (i.e. this antenna, which loss-free and minor lobes), directed in direction of blackbody. We will consider that it is located at a distance of L from the antenna (Fig. 4.3.1) and relative to antenna has angular dimension Ω_M. If blackbody has a temperature of T_M, then in the single solid angle in the direction, which composes angle Θ_0 to standard/normal $\bar{n}$, intensity of radio emission, the numerically equal to the spectral power density, which emerges from the unit of the surface of this body (see Fig. 4.3.1), according to [4.4]:

$$J = \frac{2kT_M}{2\pi\lambda^2}\cos\Theta_0. \qquad (4.3.5)$$

If the antenna is designed for reception of oscillations with linear polarization, then value J it is expedient to represent in the form two components $J=J^*+J^{**}$, one of which completely corresponds to polarization of oscillations, taken by antenna, and the second - orthogonal to it. In the case if the source creates nonpolarized radiation $J^*=J^{**}=J/2$. It is obvious that during the nonpolarized emission, acting on the antenna will be the spectral intensity or the spectral power density, which emerges from the single surface

$$J' = \frac{J}{2} = \frac{kT_M}{2\pi\lambda^2}\cos\Theta_0. \qquad (4.3.6)$$

The spectral power density, determined by expression (4.3.6), will be accepted as antenna with efficient surface $S_{A\Phi\Phi}$ in solid angle

$\frac{S_{\Theta\Phi\phi}}{L^2}$ (see Fig. 4.3.1). Consequently, the antenna power adopted, which emits the blackbody from surface element $d\sigma$ (see Fig. 4.3.1) it will be

$$dP = \frac{kT_{\text{я}}}{2\pi\lambda^2}\cos\Theta_0 d\sigma\frac{S_{\Theta\Phi\phi}}{L^2}.$$

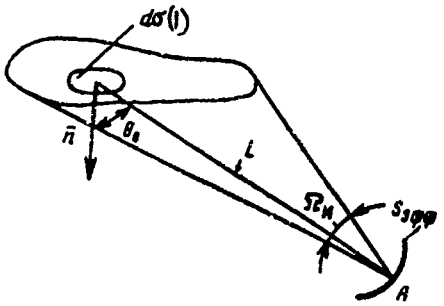

Fig. 4.3.1. Determination of equivalent temperature of the source.

Key: (1). dB.

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Since the value $(1/L_2)\cos\theta_0 d\sigma=d\Omega$ - solid angle of area/site $d\sigma$ from direction of the antenna, for determining the total power taken by the antenna, should be led integration for solid angle $\Omega_и$, in limits of which the blackbody proves to be "visible" from the location of the antenna. Thus,

$$P = \int_{\Omega_и} \frac{kT_a}{2\pi\lambda^2} S_{эфф} d\Omega_и. \qquad (4.3.7)$$

Let us take into account that the efficient surface of the antenna can be expressed through directive gain $D_и$ [see (3.3.9)], and the latter is determined by expression (3.3.3).

We convert expression (3.3.3), first of all, let us pass from integration for spherical surface to integration in limits of solid angle Ω_a, out of which reception to ideal antenna is impossible.

Taking into account the aforesaid, let us find

$$S_{\text{эфф}} = \frac{\lambda^2}{4\pi} D_{\text{н}} = \frac{\lambda^2}{4\pi} \frac{F^2(\theta, \varphi) F}{\int\limits_F F^2(\theta,\varphi)\, dF} = \frac{\lambda^2}{4\pi} \frac{4\pi F^2(\theta, \varphi)}{\int\limits_{4\pi} F^2(\theta,\varphi)\, d\Omega_{\text{а}}} = \frac{F^2(\theta, \varphi)\lambda^2}{\int\limits_{\Omega_{\text{а}}} F^2(\theta, \varphi)\, d\Omega_{\text{а}}}. \tag{4.3.8}$$

Substituting (4.3.8) into (4.3.7), let us find

$$P = \frac{k}{2\pi} \frac{\int\limits_{\Omega_{\text{н}}} T_{\text{н}} F^2(\theta,\varphi)\, d\Omega_{\text{н}})}{\int\limits_{\Omega_{\text{а}}} F^2(\theta,\varphi)\, d\Omega_{\text{а}}}. \tag{4.3.9}$$

When the blackbody, equivalent on spectral power density to space source, has in individual sections with coordinates θ and φ different brightness temperature, the last expression can be registered in more general view

$$P = \frac{k}{2\pi} \frac{\int\limits_{\Omega_{\text{н}}} T_{\text{н}}(\theta,\varphi) F^2(\theta, \varphi)\, d\Omega_{\text{н}}}{\int\limits_{\Omega_{\text{а}}} F^2(\theta,\varphi)\, d\Omega_{\text{а}}}. \tag{4.3.10}$$

The power at the output of antenna as a result of reception of radio emissions of space source or equivalent to it blackbody can be characterized with equivalent temperature of source, led to antenna $T_{\text{а}}$. By value $T_{\text{а}}$ it is accepted to understand the temperature, to which it is necessary to heat resistor/resistance of R (equal to output antenna resistance) so that the increment in the power of noise, created by this resistor/resistance on the matched load, would be equal to an increment in the power of noise from the radio emission of source (on the same load).

If resistor/resistance of R has temperature T_a, then the power of the noise in equivalent band of frequencies Δf on matched load (equal R) is determined by expression $P_ш = kT_a\Delta f$; upon transfer to angular frequencies $\Delta\omega = 2\pi\Delta f$ let us find

$$P'_ш = \frac{kT_a}{2\pi}\Delta\omega.$$

For single band we will obtain

$$P''_ш = \frac{kT_a}{2\pi}. \qquad (4.3.11)$$

Equalizing (4.3.10) and (4.3.11), let us find

$$T_a = \frac{\int\limits_{\Omega_и} T_я(\theta,\varphi)F^2(\theta,\varphi)\,d\Omega_и}{\int\limits_{\Omega_a} F(\theta,\varphi)\,d\Omega_a}. \qquad (4.3.12)$$

If angular dimensions of the source of radio emission are much greater width of main lobe of radiation of antenna, i.e., $\Omega_и \gg \Omega_a$, and within limits of Ω_a the value of brightness temperature can be considered constant $T_я(\theta,\varphi) = T_я$, from (4.3.12) let us find

$$T_a = \frac{T_я\int\limits_{\Omega_и} F^2(\theta,\varphi)\,d\Omega_и}{\int\limits_{\Omega_a} F^2(\theta,\varphi)\,d\Omega_a} = T_я, \qquad (4.3.13)$$

i.e. equivalent temperature of source, led to antenna, proves to be to equal brightness temperature of observed section of radio emission and does not depend on parameters of ideal antenna.

If the angular dimension of source $\Omega_и$ is considerably less than width of antenna radiation pattern Ω_a, then it is possible to consider

that in limits $\Omega_и$ value $F(\Theta, \varphi)=F$ is constant/invariable. Then

$$T_а = T_и \frac{\Omega_и}{\Omega_а}. \qquad (4.3.14)$$

When $\Omega_и \geqslant \Omega_а$ expression (4.3.14) will pass into expression (4.3.13). By other words, an increase in the sizes/dimensions of the antenna (and as the corollary of this, the decrease of the width of its radiation pattern) increases equivalent temperature $T_а$ of any source, converted to the input of the antenna, only until the width of the antenna radiation pattern $\Omega_а$ becomes equal to the angular dimensions of source $\Omega_и$.

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From an examination of (4.3.13) it follows that with large dimensions of the source and small changes in its brightness temperature the intensity of the radio emission it is expedient to characterize by equivalent temperature, led to antenna, since in this case this index of intensity of emission of source does not depend on parameters of antenna. Completely opposite conclusions can be drawn in the examination of expression (4.3.14), into which enters the index of the width of the antenna radiation pattern. Therefore, the sources of radio emission with the relatively low angular dimensions it is accepted to characterize not by equivalent temperature of the source, reduced to the antenna, but by another index - usually by the radiant flux, per unit of area ($м^2$) on the surface of the Earth.

It must be noted that expression (4.3.10) is correct, when

radiating body is located relative to antenna in remote zone, for which concept of directivity and antenna radiation pattern can be considered correct.

The case when the radiating body is arbitrarily located relative to antenna is examined in detail in several works: [4.5], [4.6], whence it follows that intensity of emission can be characterized by equivalent temperature of source, led to antenna. With the large sizes/dimensions of the radiating body (for example, in the examination of the thermal radiation of the atmosphere, which surrounds antenna), equivalent temperature of source, led to the matched load of ideal antenna, does not depend on the parameters of antenna.

One should note that emissions of space sources and atmosphere to a considerable degree determine the sensitivity of the receiver.

Let us pass to examination of special features of noises of different sources.

Galactic noise, as has already been mentioned in Section 4.1, it is possible to subdivide into two groups: noises, caused by discrete sources, and distributed galactic noises (background noise), which do not have specific sources.

Table 4.3.1 gives information about noises of most powerful/thickest galactic sources (based on materials [4.7]).

For calculation of power of noise of that taken by antenna, directed toward source, with width of band 1 Hz it is necessary to multiply the values given in Table 4.3.1 by the efficient surface $S_{эфф}$ of antenna.

Table 4.3.1.

Частота Мгц (1)	Спектральная плотность мощности вт/м², гц (2)		
	(3) Кассиопея-А	(4) Лебедь-А	(5) Телец-А
200	$6{,}5\cdot10^{-23}$	$4{,}3\cdot10^{-23}$	$1{,}5\cdot10^{-23}$
1000	$3\cdot10^{-23}$	$2\cdot10^{-23}$	10^{-23}
3000	10^{-23}	$7{,}5\cdot10^{-24}$	$8\cdot10^{-24}$
10 000	$4\cdot10^{-24}$	$1{,}7\cdot10^{-24}$	$5\cdot10^{-24}$

Key: (1). Frequency MHz. (2). Spectral power density W/m², Hz. (3). Cassiopeia-A. (4). Cygnus-A. (5). Taurus-A.

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Intensity of distributed galactic emission is maximum in direction of Milky Way and is minimum in direction of galactic poles. This noise virtually does not depend on the efficient surface of antenna, since the angular dimension of noise source almost is always greater major lobe of the antenna radiation pattern. The noise intensity can be expressed in W/Hz or in the form of the equivalent temperature source reduced to the antenna. The approximate numerical values can be found in Table 4.3.2, from which it is clear that with an increase in the frequency the noise temperature sharply decreases.

The intensity of radiation of the sun (constant component and average value of slowly varying component) can be illustrated by data of Table 4.3.3. [4.7]. In the numerator there are cited data which correspond to the maximum, while in the denominator, to the minimum

of solar activity.

Radiant fluxes of Moon and planets, measured at frequency 3000 MHz, are equal approximately to $7.6 \cdot 10^{-22}$ W/m²Hz. Flow at the frequency of 9000 MHz comprised for Mars $6.5 \cdot 10^{-24}$, for Saturn - 4.10^{-24}, for Venus from 4 to $9 \cdot 10^{-24}$ W/m²Hz [4.8]. The averaged, on the disk of the Moon, Venus and Jupiter, brightness temperature at the frequency of 3000 MHz respectively comprises 218, 600 and 700°K, and at the frequency of 9400 MHz - 210, 580 and 180°K [4.7].

A comparison of flow values of emission shows that sun - most intense radiation source.

One should note that angular diameter of sun for a "terrestrial" observer is 32', and the angular diameter of the moon in perigee and apogee is, respectively, 33.7' and 29.4'. Therefore, the probability of the direction of the receiving antenna to one or the other planet accurately proves to be very low. But it is necessary to consider the presence of the minor lobes of the antenna radiation pattern, which can coincide with the direction to the sun, which in the periods of the intensive solar radiation can lead to disturbance/breakdown or deterioration in the connection/communication.

Table 4.3.2.

Частота, Мгц (1)	Эквивалентная температура шумов, °К (2)	
	(3) минимальная	(4) максимальная
200	100	1500
1000	<1	30
3000	<1	1

Key: (1). Frequency MHz. (2). Equivalent noise temperature, °K. (3). minimum. (4). maximum.

Table 4.3.3.

Частота, Мгц (1)	10 000	3000	1200	600	200
Спектральная плотность мощности, $\frac{вт}{м^2 \cdot гц} \cdot 10^{-21}$ (2)	$\frac{32}{27}$	$\frac{13}{6,5}$	$\frac{7}{3,5}$	$\frac{5}{2,5}$	0,85

Key: (1). Frequency, MHz. (2). Spectral power density, $W/m^2 \cdot Hz\ 10^{-21}$.

As show given numerals, power of noises, caused by emission of planets, in comparison with power of noises of other sources is very low and it can be disregarded/neglected.

Noise of water vapor and molecular oxygen. Computation of intensity of emission (thermal noises) of atmospheric oxygen and water vapor, and also dependence of value of these noises from angle of elevation of antenna and frequency are given in [4.9].

Fig. 4.3.2 [4.10] shows dependence of equivalent temperature of atmosphere, given in antenna, on frequency and angle of elevation β, and also are given the straight lines showing exemplary/approximate limits of temperature of cosmic noises. Equivalent temperature of the atmosphere, led to the antenna, or as it is sometimes called, the temperature of antenna, depending on angle of elevation β it can be found from the approximation formula

$$T_a \approx \frac{2,5}{\sin\beta}, \text{ °K.} \qquad (4.3.15)$$

This formula is valid in the range 1-8 GHz at β>3-5°.

With value $\beta \to 0$ value T_a in frequency band indicated in accordance with graphs/curves Fig. 4.3.2 reaches 90-140°.

Theoretical curves of Fig. 4.3.2 coincide well with results, obtained experimentally [4.18].

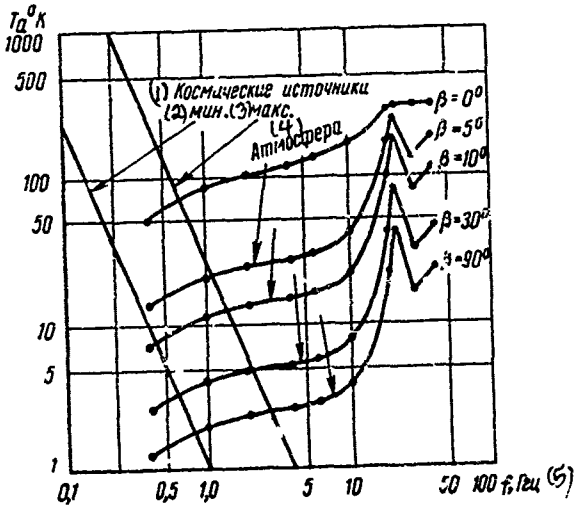

Fig. 4.3.2. Equivalent temperature of space sources and atmosphere.

Key: (1). Space sources. (2). min. (3). max. (4). Atmosphere. (5). f, GHz.

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Examination of curves Fig. 4.3.2 show that in the range to 4 GHz it is necessary to consider both the cosmic noises and the noises, caused by atmospheric absorbtion. At higher frequencies (from 4 to 10 GHz) only the noise, connected with the atmospheric absorbtion, is developed. With the decrease of angles of elevation the value of noises grows so rapidly that the use of angles of elevation of less than 5-7° is unsuitable. One should note that at small angles of increase the probability of interferences from ground radio engineering equipment, industrial units and reflections from all possible ground features increases also.

During calculations of curves, shown in Fig. 4.3.2, it was assumed that absorption, and consequently, also emission of atmosphere are completely determined by water vapor and oxygen at frequencies 22.5 and 60 GHz respectively. During calculations the dependence of temperature and pressure of the atmosphere on the height/altitude was taken in accordance with the international standard, and the content of water vapor was received as that changing according to the linear law - from 12 g/m$_3$ on the earth's surface, to zero at the height/altitude of 5 km.

With other content of water vapors, as show theoretical and experimental works, equivalent temperature of atmosphere changes.

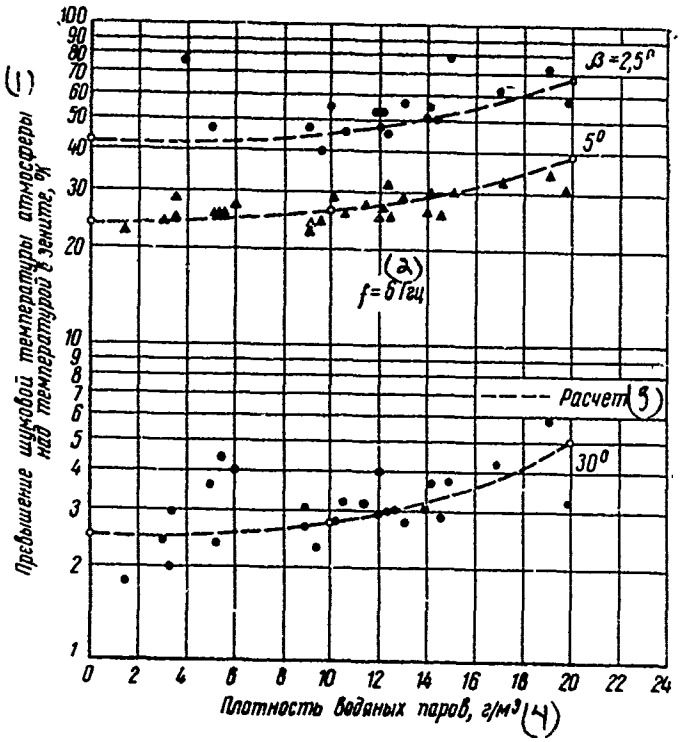

Fig. 4.3.3. density effect of water vapor on equivalent temperature of the atmosphere.

Key: (1). Excess of noise temperature of the atmosphere above the temperature in the zenith, °K. (2). GHz. (3). Calculation. (4). Water vapor density g/m³

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Fig. 4.3.3 [4.11] gives the result of calculating the excess of value T_a above equivalent temperature of the atmosphere in the zenith from the density of water vapor for different angles of elevation β (dotted curves) and the measurement at the frequency of 6 GHz. The curves of Fig. 4.3.3 attest to the fact that a change in the water vapor density especially strongly is manifested with the values, which exceed 12- 14

g/m³; in this case an identical change in the water vapor density in the case of small angles of increase, a considerably larger absolute increase in equivalent temperature is caused, than in the case of large angles of elevation. The calculation of the effect of rain on the noise temperature of the atmosphere is difficult.

Fig. 4.3.4 presents comparative data about the noise temperature of the atmosphere in the range 6 GHz during the rainy and dry days depending on angle of elevation β.

Besides water vapor oxygen, in the atmosphere, especially at high altitudes, there are also other absorbing media - ozone and different ionic formation/educations. However, the concentration of these media is low; therefore at frequencies, which exceed 100 MHz, by the absorption (and consequently, and by emission) of these media it is possible to disregard/neglect.

Of the aforesaid it is above and the graphs/curves Fig. 4.3.2 it follows that for increasing the reliability of connection/communication and decrease of noises the frequency band, intended for the communication systems through ISZ, must be within the limits from 3 to 8 GHz. At higher frequencies the noise due to the absorption in the water vapor (especially with the strong rains) increases, while lower frequencies affects the effect of cosmic noises. Precisely these considerations, and also the account of the already available radio equipment, they were assumed as the basis of

the international distribution of radio-frequency bands for the communication systems through ISZ accepted in Geneva in 1963. This distribution given in Appendix 3.

Intensity of emission of Earth. The equivalent noise temperature of the Earth, led to matched output antenna resistance, is taken as the equal to 290°K.

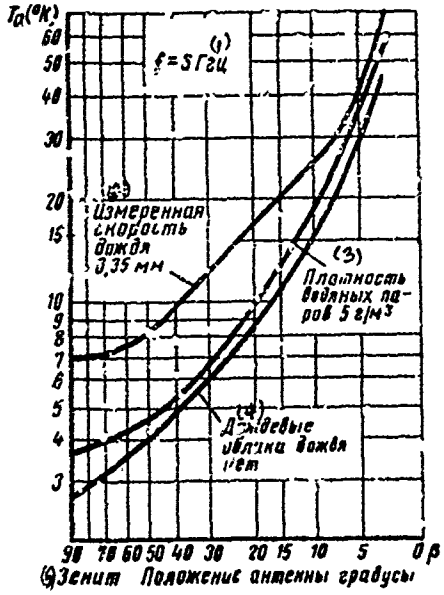

Fig. 4.3.4. Noise temperature of atmosphere for dry and damp/raw day.

Key: (1). GHz. (2). Measured speed of rain 6.35 mm. (3). Charge of water vapor of 5 g/m³. (4). Rain clouds there is no rain. (5). Zenith. Antenna position degrees.

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4.4. Noise temperature of antenna.

Noise temperature of real antenna is determined by two factors: by thermal radiation of atmosphere and by emission of Earth. In the communication systems through ISZ are utilized terrestrial antennas with the larger factor of amplifications, which have the width of radiation pattern of approximately one degree or even less. Such antennas, taking into account Fig. 4.3.2, for decreasing the equivalent noise temperature of the atmosphere are utilized at the

angles of elevation, greater than by 5-7°.

Consequently, it is possible to consider that emissions of Earth will act on ground-based antenna only through minor lobes. If we consider that at the angles of elevation of $\beta>7°$ in the range of the frequencies of 3-8 GHz equivalent temperature of the atmosphere in the absence of rain according to Fig. 4.3.2 does not exceed 20°K, but equivalent temperature of the Earth is equal to 290°K, then becomes obvious the effect of the side-lobe level of antenna on its total temperature. let us explain this by the curves of Fig. 4.4.1 of work [4.11].

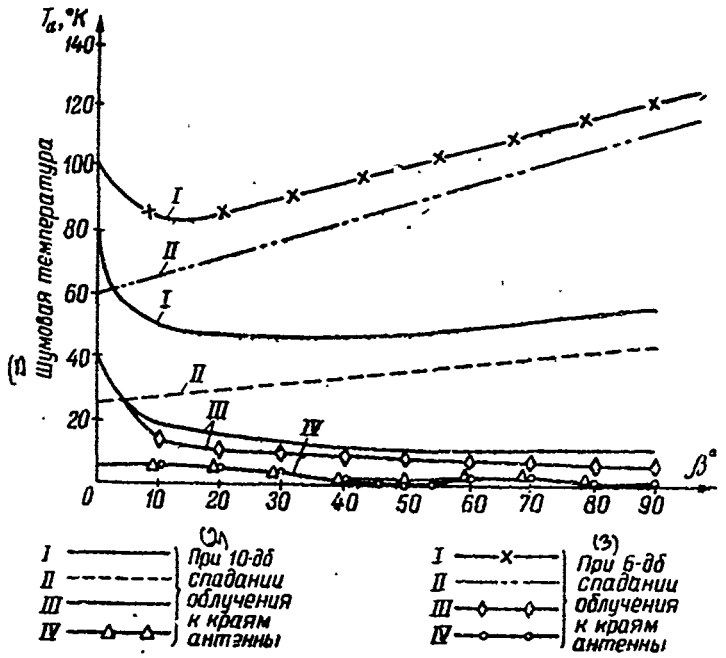

Fig. 4.4.1. dependence of the noise temperature of antenna on angle of elevation with 10 and 6 dB the drop of irradiation to the edges of antenna, f=2 GHz; I - total noise temperature; II - portion of the minor lobes, III - portion of major lobe; IV - portion of rear lobes/lugs.

Key: (1). Noise temperature. (2). With 10 dB drop of irradiation to edges of antenna. (3). With 6 dB the drop of irradiation to the edges of antenna.

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In this figure it is shown the dependence of the noise temperature of antenna at the frequency of 2 GHz on the angle β with two versions of irradiation, and also relative values of the noises, which fall to major lobe of diagram and minor lobes of front/leading and rear

hemispheres. It turns out that the noise, which arrive on the minor lobes, is most significant and precisely these noise determines the inherent noise level of antenna. In this case is very important the fact that the noise, which arrives through the minor lobes, to a considerable extent depends on the method of irradiating the antenna dish: with the sharper drop of irradiation to the edges of antenna smaller minor lobes are obtained and noise temperature decreases. One should note that simultaneously with this deteriorates the use of a surface of antenna, which leads to reduction in the amplification factor with constant dimensions of antenna dish.

They usually consider that the optimum drop of field to the edges of aperture is equal to 10 dB. However, in this case minor lobes prove to be still considerable and in the case of using the low-noise input receivers the value of antenna noises can prove to be very noticeable. Transition/junction to larger decay in the irradiation at the edges of mirror, although it will lead to the decrease of the antenna gain, nevertheless, due to a decrease in the side-lobe level, it is possible to lead to an improvement in the resulting signal-to-noise ratio.

In connection with this, since in practice condition of drop of irradiation to edges of mirror usually is satisfied, at least, on 10 dB, in accordance with Fig. 4.4.1 we will approximately consider that due to minor lobes temperature of ground-based antenna increases by value

$$\Delta T_3 = 23 + 0{,}2\,(90 - \beta), \qquad (4.4.1)$$

where β - determines angle of elevation in degrees.

Application for terrestrial stations of antenna fairings (protective hoods) leads to additional increase in total temperature of antenna.

Fig. 4.4.2 gives theoretical dependence of the component of noise temperature of antenna at different frequencies, caused by fairing, covered with water layer of different thickness [4.12]. Dotted lines showed the total component of noise temperature antennas upon consideration not only of the layer of water, by the thickness κ, which covers fairing, but also the fairing itself, made from the dielectric material, for which ϵ=3+i0.045. in work [4.12] is noted that the thickness of water layer, equal to 0.125 mm and 0.25 mm, with the spherical fairings with a radius of 15 m it is formed at the intensity of rainfall 3 mm/h and 25 mm/h respective

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For the spherical fairings about the radius of 30 m the thickness of water layer, equal to 0.125 mm and 0.25 mm, it is formed during precipitation with an intensity of 1.5 mm/h and 12.5 mm/h respectively. Precipitation with an intensity of 3 mm/h in Tokyo on the average are observed approximately by 2% of time in the year, in Paris - 0.5%, and in Halifax (Canada) and Washington - about 1% of time in the year. Precipitation with an intensity of 25 mm/h can

occur on the average in the year in 0.1 periods in Tokyo and less than 0.01% time in Paris.

One should note that dry protective hoods of antennas create comparatively small supplementary noise temperature, whose value proves to be equal to approximately 3-10°K.

On the basis of aforesaid total of temperature of ground-based antenna taking into account protective hood can be determined according to expression

$$T_{a\Sigma} = T_a + \alpha T_з + T_{зк} = T_a + 23 + 0{,}2(90-\beta) + T_{зк}.$$

Here T_a is determined according to graphs/curves Fig. 4.3.2 in the case of dry weather or according to graphs/curves Fig. 4.3.3 and 4.3.4 in presence of precipitation, β is determined angle of elevation (in degrees), and $T_{зк}$ is located through curves of Fig. 4.4.2.

For onboard antennas, oriented to the Earth, it is possible to consider that $\Omega_a > \Omega_з$, and $T_з > T$ (here Ω_a - aperture angle of radiation pattern of onboard antenna, $\Omega_з$ - angular dimension of Earth, "observed" from onboard of satellite, $T_з$ - temperature of Earth, equal to 290°, T - temperature of medium and elements of construction/design, which surround onboard antenna). Taking into account that, besides the thermal radiation of the Earth, on onboard antenna the thermal radiation of the atmosphere will still act, moreover this effect will occur at the angles of increase of $\beta \cong 90°$, in

accordance with expression (4.3.14) we will obtain

$$T_{\text{в}} = [T_{\text{з}} + T_{\text{а}}(90)]\,\frac{\Omega_{\text{в}}}{\Omega_{\text{а}}} = [290 + T_{\text{а}}(90)]\frac{\Omega_{\text{в}}}{\Omega_{\text{а}}}. \qquad (4.4.2)$$

Here $T_{\text{а}}$ (90) are located through the curve on Fig. 4.3.2 at β=90°.

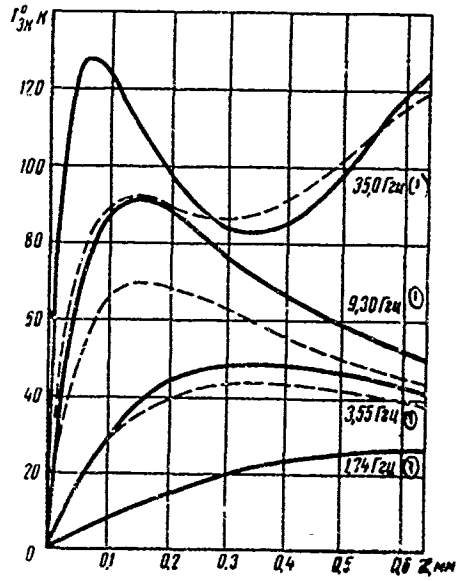

Fig. 4.4.2. Dependence of noise temperature of radome on thickness of water layer.

Key: (1). GHz.

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4.5. Changes in the level of noise inlet temperature of the receivers of terrestrial stations.

As was noted in Section 4.1, noise level at input of terrestrial receivers is determined by expression (4.1.1) and depends on many factors. Expression (4.1.1) can be represented in the form

$$P_{ш} = k \Delta f T_{\Sigma}, \qquad (4.5.1)$$

where Δf - resulting band, T_{Σ} - total effective inlet temperature of receiver, which depends both on the environmental factors (effect of galactic noise, the effect of precipitation, etc.) and from the noise

temperature of antenna, fairing and elements of receiver.

During calculations of communication systems through ISZ, correct estimate of the magnitude T_a has vital importance.

Fig. 4.5.1 gives dependence T_a on precipitation intensity for two different systems. Curve 1 relates to the receiving system in Raisting (FRG) at the angles of elevation from 25° to 45°. Horn-parabolic antenna in Raisting has a fairing with a diameter about 49 m, made from the dielectric with a thickness of 1.8 mm, which has $\epsilon=3$, and $tg\delta=0.015$ (δ - loss angle). Curve 2 relates to the receiving system, which is located in Andover (USA), where horn-parabolic antenna with the fairing is utilized. Antennas in Raisting and Andover work in one and the same frequency band - about 4000 MHz during reception, and about 6000 MHz during the transfer.

Fig. 4.5.2 gives dependences T_a on angle of elevation under different climatic conditions.

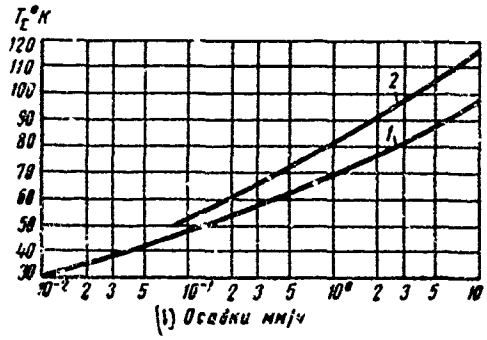

Fig. 4.5.1. Dependence T_a on precipitation intensity.

Key: (1). Precipitation mm/h.

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Curves 1, 2, 3 and 4 [4.13] show change T_a of the antenna in Raisting; moreover curve 1 is taken in dry fairing and cloudless sky (5 hours of 30 min.); curve 2 - with the weak rain (intensity of 0-1.2 mm/h 5 hours of 50 min.); curve 3 - with the rain with an intensity of 1.21-6 mm/h (6 hours of 15 min.); curve 4 - after the termination of rain (6 hours of 25 min.). From the comparison of the enumerated curves it is possible to make a conclusion, first, about substantial change T_a upon the appearance of precipitation (especially at the large angles of elevation), and secondly, about the sufficiently good coincidence of curves 1 and 4, taken prior to the beginning and after the termination of rain.

Fig. 4.5.2 by dotted line shows curved 5, which corresponds to change T_a from angle of elevation, obtained in dry weather for receiving system in Andover [4.14]. The comparison of curves 1 and 5

shows that the character of change $T_{\text{в}}$ from the angles of elevation for different receiving systems coincides sufficiently well. Curve 6, given in Fig. 4.5.2, it is constructed according to the theoretical curves, which determine the noises of the dry atmosphere, which are given in Fig. 4.3.2 for the oscillations/vibrations with a frequency of 4000 MHz.

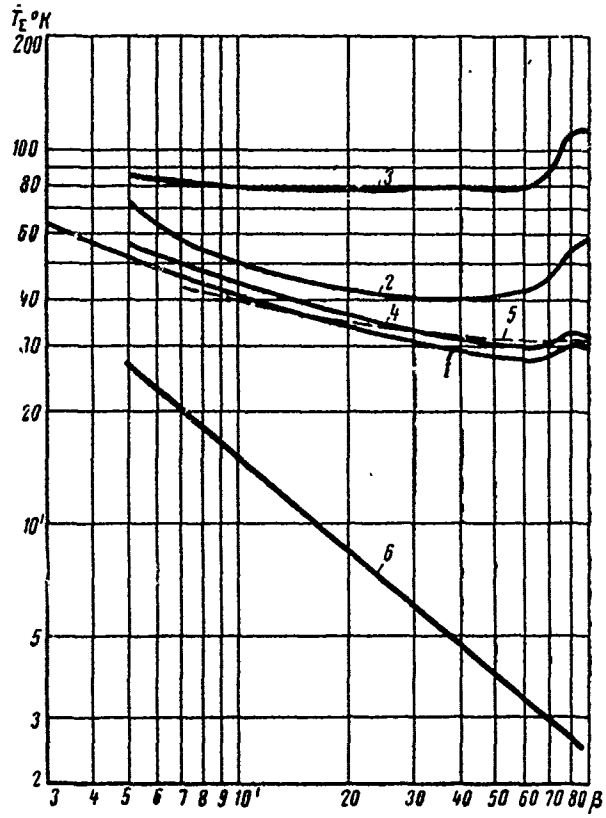

Fig. 4.5.2. Change T_a from angle of elevation under different climatic conditions.

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The comparison of curves 1, 4, 5 and 6 indicate the validity of theoretical data, used for Fig. 4.3.2, and also the fact that the inclination/slope of curves 1-5 is determined by atmospheric noises. The disagreement between curved 6 and 1, and also 5 and 1 is determined by the intrinsic noise of receiver, by noises of the radome, feeder, etc.

It is interesting to examine values of different components of

noises, which determine value of total temperature T_Σ at angle of elevation, equal to 90°, during clear day. The values of the equivalent noise temperature of separate components with these conditions for the receiving systems in Andover [4.14] and Raisting [4.19] are given in Table 4.5.1.

Appearance of precipitation leads not only to increase in noise inlet temperature of receiver, but also to certain weakening of signal. As a result reduction in the ratio of the power of signal to the intensity of noises at the input of receiver is observed. This can be shown by the curves of Fig. 4.5.3, where curve 1 characterizes a change in the ratio of the power of signal to the noise for the receiving system in Raisting [4.13] at the angles of elevation from 25° to 45°. Curve 2 is constructed for the case, when the power of signal does not change, but only the power of noises in accordance with curve 1, Fig. 4.5.1 changes. Thus, a difference in the ordinates between curves 1 and 2 in Fig. 4.5.3 determines fading signal in the presence of precipitation. Curve 3 characterizes change in the relation in question at the angle of elevation, equal to 28°, for the receiving system in Mill Village (Canada), where is established/installed two-mirror antenna (Cassegrain), by the diameter of 25.9 m, the covered by fairing [4.15] antenna works in the range of approximately 4000 MHz and 6000 MHz and has amplification by 59 and 60 dB respectively.

Table 4.5.1.

(1) Наименование компонента, определяющего шумовую температуру, °К	(2) Приемные системы	
	(3) в Андовере	(4) в Райстинге
Атмосфера при сухой погоде (5)	2,4°	2,0°
Рассеянне обтекателя и боковые лепестки антенны (6)	8,4°	9°
Поглощение обтекателя (7)	3°	3°
Потери антенны (8)	не приведена	(9) 0,5°
Потери в волноводных цепях (10)	14,2°	10,2°
Шумы приемника (квантовомеханический усилитель с последующим каскадом)	(11) 4°	4,3°
Суммарная шумовая температура (12)	32,0°	29,0°

Key: (1). Designation of the component, which determines the noise of temperatures, °K. (2). Receiving systems. (3). in Andover. (4). in Raisting. (5). Atmosphere in dry weather. (6). Scattering fairing and minor lobes of antenna. (7). Absorption of fairing. (8). Losses of antenna. (9). it is not given. (10). Losses in waveguide circuits. (11). Receiver noise (quantum-mechanical amplifier with subsequent cascade/stage). (12). Total noise temperature.

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Curve 4 in Fig. 4.5.3 characterizes a change in the ratio of the power of signal to the power of noise in the case of precipitation of different intensity for the receiving system in Andover [4.14].

Together with given graphs/curves, during calculations of communication system through ISZ and evaluations/estimates of qualitative indices of such systems, it is very important to know distribution in time of change in total noise temperature of entire

receiving system T_z and ratio of power of signal to power of noises.

Fig. 4.5.4 shows curves, which characterize distribution of value in time: along axis of abscissas is plotted probability (in percentages), with which value T_z - is less than values, indicated along axis of ordinates. Curves 1, 2, 3, 4 for different angles of elevation are obtained at the receiving station in Ploemeur-Baud (France) [4.16].

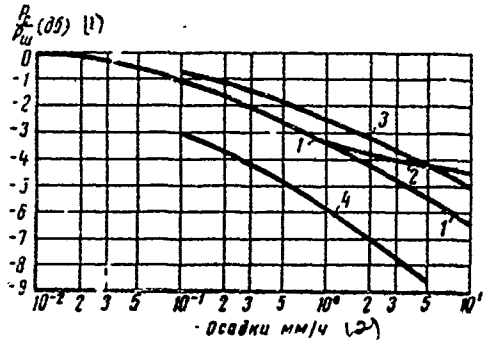

Fig. 4.5.3. change in the ratio of the power of signal to the noise from the intensity of the precipitation.

Key: (1). (dB). (2). Precipitation, mm/h.

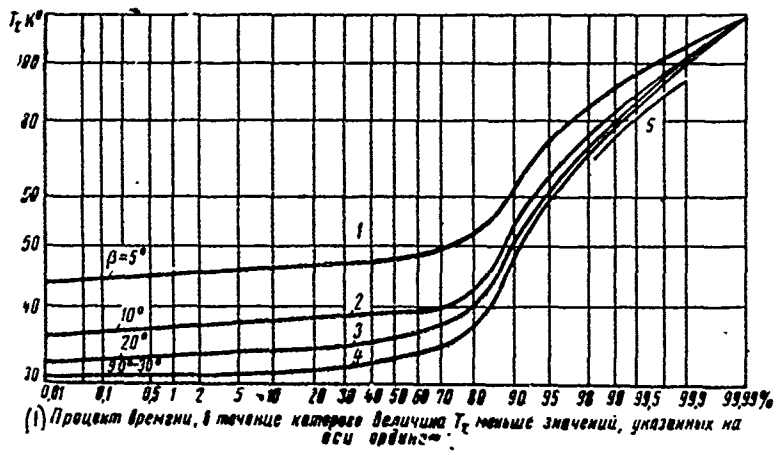

Fig. 4.5.4. Distribution T_Σ

Key: (1). Percentage of time, during which value ... less than the values, indicated on the axis of ordinates.

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Let us note that statistical change T_Σ in the curves of Fig. 4.5.4 is determined by meteorological conditions. It follows to indicate that in Ploemeur-Baud reception is realized with the horn-parabolic antenna

system, analogous antenna to the system, established/installed in Andover. Curve 5 in Fig. 4.5.4 characterizes distribution in the time of value $T_в$ for the receiving installation in Raisting [4.13]. The distribution in the time of a change in the relation of the power of signal and power of noise (in dB), measured in the band 8 kHz for the receiver in Ploemeur-Baud, is given in Fig. 4.5.5. On this graph/curve is shown the distribution of the relation indicated for the case, when the reception of the signals of the synchronous satellite HS-303, whose orbit had an inclination of 18°, was realized, and the ascending node - 30° the western longitude/length.

Certainly, during the reception of emissions of satellites, which move along the orbits with other parameters, distribution curve will take another form, than in Fig. 4.5.5. It is natural that relocation of reception or equipment will also give to with changes in the distribution curve. From the examination of the given curves very essential conclusion/output can be made - in the communication systems through ISZ the distribution in the time of the ratio of the power of signal to the power of noises has smooth character, the range of a change in this relation lying/resting within the limits of several dB.

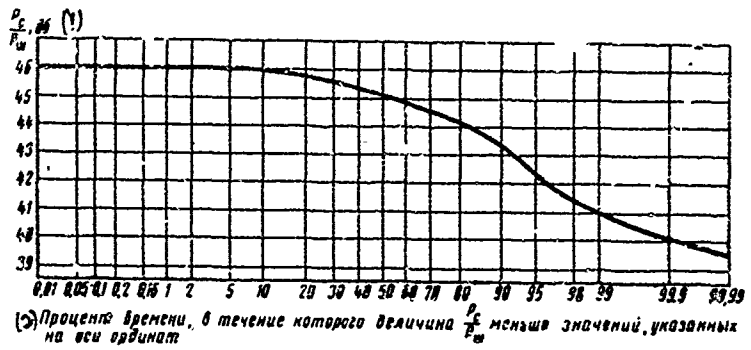

Fig. 4.5.5. Distribution of ratio of power of signal to noise.

Key: (1). dB. (2). Percentage of time, during which value ... less than values, indicated on axis of ordinates.

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Chapter 5.

Calculation of the ratio of the power of signal to the thermal noises at the output of the communication system.

5.1. Initial relationships/ratios.

Signals, transmitted by line of communications, undergo effect of interferences. It is accepted to distinguish two forms of the interferences: additive and multiplicative.

Additive interferences can act on receiver independent of presence of signal and are developed in the form of interferences, which are totalled with signal. Additive interferences are caused by different reasons: by the thermal agitation of molecules, by the emissions of different stations, which affect the input of receiver, but its not changing parameters, by man-made interferences, etc.

Multiplicative interferences lead to random changes in parameters of communication channel and they are developed only in presence of signal. The multiplicative interferences include different phenomena, which occur in the communication channel, including the phenomenon of the fadings, caused by multiple-beam propagation of signal.

In communication systems through ISZ at sufficiently large angles

of elevation (β>3-5°) with phenomenon of multiple-pronged propagation virtually it is possible not to be considered. The parameters of the communication channel through ISZ it is possible to consider time-constant: the change of the signal level at the input of receiver, caused by the variable distance during the motion of satellite relative to the Earth or by other reasons (for example, by supplementary attenuation in precipitation), it occurs in the small limits relative to the value of fundamental attenuation. These changes are very slow - signal level varies from maximum to minimum value for several hours. Therefore in the communication systems through ISZ should be considered only additive interferences. This is confirmed by experimental investigations. Let us note that the additive interferences are frequently called noises; this term will be used with the following presentation.

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In Chapters 3 and 4 were obtained relationships/ratios, on the basis of which, knowing fundamental indices of channel of communication (orbital parameters, power of transmitter, radiation pattern of antennas, temperature of inherent noise of receivers, etc.), it is possible to determine signal-to-noise ratio at input of receivers. In this case by noises at the input of receivers the thermal noises was implied.

Let us switch over to determination of signal-to-noise ratio at output of receiver and effectiveness of communication system with different methods of modulation: single-band (OBP), frequency

modulation (ChM) and pulse-code (KIM).

In communication systems through ISZ, as in other systems, radio channel must satisfy the requirements of high noise immunity and effectiveness of use. By noise immunity is understood the ability of system to resist the effect of interferences. In the practical calculations they usually operate with relative noise immunity, which makes it possible to compare between themselves the diverse variants of the technical solutions, selected both for twist the systems (for example, the form of modulation) and for its separate components/links.

Quantitative measure of noise immunity is degree of conformity of communication/report accepted to that transmitted (correctness).

During transmission of continuous communications/reports by criterion of correctness it is possible to count rms error [5.1]

$$\varepsilon^2 = \overline{[v_1(t) - v(t)]^2}. \qquad (5.1.1)$$

Here v(t) and $v_1(t)$ represent respectively the totality of the transmitted and taken signals. Difference $v_1(t)$-v(t) can be considered as interference (noise) at the output of receiver.

Relative error E is defined as:

$$E = \frac{\varepsilon^2}{P_{\text{с вых}}} = \frac{\int_0^F \sigma^2_{\text{вых}}(f)df}{\int_0^F P_c(f)\,df} = \frac{P_{\text{ш вых}}}{P_{\text{с вых}}} = \frac{1}{q_{\text{вых}}}, \qquad (5.1.2)$$

where $\sigma_{вых}(f)$ - jamming intensity at frequency f at output of receiver,

F - cut-off frequency of passband of receiver after detector; since integration is conducted from zero, value F is numerically equal to bandwidth at output of receiver,

$P_{с\ вых}$ - average/mean power of communication report at output of receiver in band F,

$P_{ш\ вых}$ - average/mean power of interference at output of receiver in band F.

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From (5.1.2) it follows that square of rms error is completely determined by average/mean power of interference at output of receiver, and relative error - by ratio of average/mean power of interference to average power of signal $P_{с\ вых}$. During the practical calculations as the parameter, which characterizes noise immunity of system, take the value of gain β or generalized gain β' the systems, which are determined on the noise levels, small with respect to signal level, as follows:

$$\beta = \left(\frac{P_с}{P_ш}\right)_{вых} : \left(\frac{P_с}{P_ш}\right)_{вх} = \frac{q_{вых}}{q_{вх}}, \qquad (5.1.3)$$

$$\beta' = \left(\frac{P_с}{\frac{P_ш}{\Delta F}}\right)_{вых} : \left(\frac{P_с}{\frac{P_ш}{\Delta f}}\right)_{вх} = \frac{q'_{вых}}{q'_{вх}}. \qquad (5.1.4)$$

Here ΔF - frequency band, in which is measured power of interference at output of receiver (usually this band it coincides with band of signal), Δf - equivalent frequency band, in which it is

measured power of interference at input of receiver (see Section 4.2). From the comparison of two latter/last expressions it follows that the connection/communication between the gain of system and the generalized gain is determined by the following relationship/ratio:

$$\nu = \frac{\delta}{\delta'} = \frac{\Delta f}{\Delta F}. \qquad (5.1.5)$$

Thus, generalized gain of system is determined by ratios of signal not to power of noises, but to average/mean spectral noise density at input and output: $\sigma^2_{вх} = \frac{P_{ш\ вх}}{\Delta f}$, $\sigma^2_{вых} = \frac{P_{ш\ вых}}{\Delta F}$.

During transmission of discrete communications, accuracy of reproduction is determined by probability of error $p_{ош}$ in elements of communication/report accepted. As the criterion of correctness S accept the logarithm of the value of reverse/inverse $p_{ош}$ [5.2]:

$$S = 10\lg\frac{1}{p_{ош}} = -10\lg p_{ош}. \qquad (5.1.6)$$

Effectiveness of use of channel can be evaluated by coefficient η, which characterizes nearness of speed of transmission R to channel capacity C, i.e.

$$\eta = \frac{R}{C}. \qquad (5.1.7)$$

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As is known [5.3], maximum channel capacity of communications with bandwidth of frequencies, equal to Δf, in presence of statistically independent variables of signal and noise with average/mean power, respectively equal to P_c and $P_ш$, it is determined

as follows:

$$C = \Delta f \log_2 \frac{x^2_{c+ш}}{x^2_ш} \left(1 + \frac{P_c}{P_ш}\right). \qquad (5.1.8)$$

Here $x_{c+ш}$ and $x_ш$ - peak factor of the total process, which consists of the signal and the noise (additive interferences) and the peak factor of noise, determined by their statistical properties. If the statistical properties of communication/report and noise are identical and are characterized by the normal distribution law, total process is also similar/such to noise. Therefore

$$x^2_{c+ш} = x^2_ш. \qquad (5.1.9)$$

In this case channel capacity of communications in accordance with (5.1.8)

$$C_{вх} = \Delta f \log_2 \left[1 + \left(\frac{P_c}{P_ш}\right)_{вх}\right] = \Delta f \log_2 [1 + q_{вх}]. \qquad (5.1.10)$$

If communication/report has uniform spectrum, i.e., are satisfied condition (5.1.9), speed of transmission

$$R = \Delta F \log_2 \left[\left(\frac{P_c}{P_ш}\right)_{вых} + 1\right] = \Delta F \log_2 [q_{вых} + 1]. \qquad (5.1.11)$$

Therefore in accordance with (5.1.7), (5.1.3) and (5.1.5) it is possible to obtain following two equivalent expressions:

$$\eta_1 = \frac{1}{\nu} \frac{\log_2 [\beta q_{вх} + 1]}{\log_2 [q_{вх} + 1]}, \qquad (5.1.12)$$

$$\eta_1 = \frac{1}{\nu} \frac{\log_2 [q_{вых} + 1]}{\log_2 \left[\frac{1}{\beta} q_{вых} + 1\right]}. \qquad (5.1.13)$$

For high-quality communication systems condition

$$q_{вх} = \left(\frac{P_c}{P_ш}\right)_{вх} \gg 1. \qquad (5.1.14)$$

usually is satisfied.

In this case with certain approximation/approach instead of (5.1.12) we will obtain

$$\eta_1 = \frac{1}{\nu} \frac{\log_2 q_{вых}}{\log_2 \left(\frac{1}{\beta} q_{вых}\right)}. \qquad (5.1.15)$$

From expressions (5.1.13) and (5.1.15) it follows that with given value $q_{вых}$ value of effectiveness is determined by gain of system β and by value ν.

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Let us note that since into expressions for computing effectiveness of system enters relation of logarithms, their foundation can be any.

Ideal communication system it is possible to consider system, in which is provided maximum of channel capacity C, determined by expression (5.1.10), at speed R=C and by how conveniently small probability of errors. Therefore, equating (5.1.10) and (5.1.11) for the ideal system, we will obtain

$$\Delta f \log_2 [q_{вх} + 1] = \Delta F \log_2 [q_{вых} + 1]. \qquad (5.1.16)$$

Hence upon $q_{вх} \gg 1$ and consideration (5.1.3)-(5.1.5) we will obtain:

$$q_{вых} = q_{вх}^{\nu}, \qquad (5.1.17)$$

$$\beta = (q_{вх})^{\nu-1}, \qquad (5.1.18)$$

$$\beta' = (q_{вх})^{\nu-2}. \qquad (5.1.19)$$

Furthermore, with fulfillment of inequality $q_{вх} \gg 1$ and

substitution (5.1.3), (5.1.18) in (5.1.12) let us find that for ideal system

$$\eta_{\text{и}} = \frac{1}{\nu} \frac{\log_2 [\beta q_{\text{вх}} + 1]}{\log_2 [q_{\text{вх}} + 1]} = \frac{1}{\nu} \frac{\log_2 [q_{\text{вх}}^{\nu} + 1]}{\log_2 [q_{\text{вх}} + 1]} \cong 1. \qquad (5.1.20)$$

Should be focused attention on the fact that according to (5.1.18) in ideal communication system gain β grows/rises with increase ν according to power law.

For discrete/digital communication systems it is convenient to introduce concept of transmission rate during ideal coding R_0. In this case expression (5.1.7) can be recorded in the form

$$\eta_{\text{и}} = \frac{R}{R_0} \frac{R_0}{C}. \qquad (5.1.21)$$

Ratio R/R_0 can be named effectiveness of coding, and R_0/C - effectiveness of modulation.

For system with L by discrete/digital levels, in which all levels have identical probability of transmission and identical probability of error $p_{\text{ош}}$, speed of transmission is determined by expression [5.1]

$$R = \frac{1}{T} [\log_2 L - p_{\text{ош}} \log_2 (L-1) + p_{\text{ош}} \log_2 p_{\text{ош}} + (1 - p_{\text{ош}}) \log_2 (1 - p_{\text{ош}})]. \qquad (5.1.22)$$

Here T - duration of signal.

Let us pass to examination of threshold characteristics of communication channel. From expression (5.1.3) it follows

$$q_{\text{вых}} = \beta q_{\text{вх}}. \qquad (5.1.23)$$

The value of the gain of system, entering this expression, is the constant value, which characterizes the properties of the communication systems. This definition is correct only for the systems, in which the threshold is absent; for the threshold systems value β proves to be constant (and greatest) only with values $q_{вх}$, which exceed a certain threshold value.

Aforesaid is explained by graphs/curves: Fig. 5.1.1 for threshold system, and for nonthreshold - Fig. 5.1.2. In the case, when transition/junction from the region, which lies is below threshold value, into the region, which lies it is higher than it, it occurs very smoothly, as the threshold point is taken that, with with the approach to which from the direction of large values $q_{вх}$ value $q_{вых}$ decreases by one dB in comparison with the linear change (see Fig. 5.1.1).

Comparing expressions (5.1.3) and (5.1.4), it is possible to say that threshold values of values β and β ' will coincide. Taking into account the aforesaid, since the value of gain β, in (5.1.23), is constant, this expression is correct only for the region, which lies above threshold, i.e., when $q_{вх} \geqslant q_{вх\,пор}$ and respectively when $q_{вых} \geqslant q_{вых\,пор}$.

For determination of relationship/ratio $q_{вых}=f(q_{вх})$ in region, where $q_{вх} \leqslant q_{вх\,пор}$ and $q_{вых} \leqslant q_{вых\,пор}$, it is necessary to proceed from other

positions.

Let us examine channel capacity of communications at input of receiver (in front of demodulator) and at output (after demodulator). In accordance with (5.1.8) let us find:

$$C_{вх} = \Delta f \log_2 \left(\frac{x^2_{с+ш}}{x^2_{ш}} \right)_{вх} [1 + q_{вх}], \qquad (5.1.24)$$

$$C_{вых} = \Delta F \log_2 \left(\frac{x^2_{с+ш}}{x^2_{ш}} \right)_{вых} [1 + q_{вых}]. \qquad (5.1.25)$$

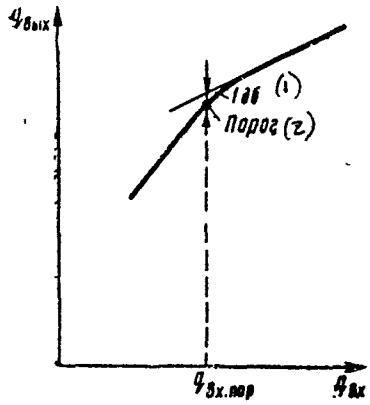

Fig. 5.1.1.

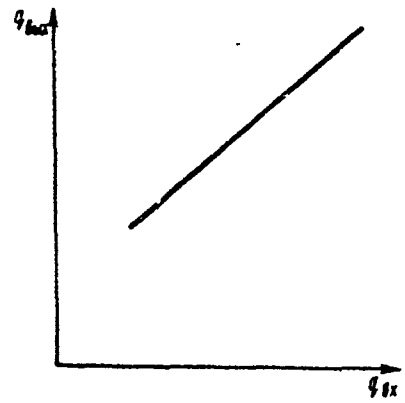

Fig. 5.1.2.

Fig. 5.1.1. Dependence, which characterizes threshold system.

Key: (1). dB. (2). Threshold.

Fig. 5.1.2. Dependence, which characterizes nonthreshold system.

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As a result of passage through demodulator of receiver quantity of information cannot increase; therefore it is possible to consider that $C_{вых} \leqslant C_{вх}$ and in accordance with latter/last expressions let us find

$$q_{вых} \leqslant \frac{(\delta_{вх})^{\nu}}{\delta_{вых}} [1 + q_{вх}]^{\nu} - 1. \qquad (5.1.26)$$

Here:

$$\delta_{вх} = \left(\frac{x^2_{с+ш}}{x^2_{ш}} \right)_{вх}, \qquad (5.1.27)$$

$$\delta_{вых} = \left(\frac{x^2_{с+ш}}{x^2_{ш}} \right)_{вых}. \qquad (5.1.28)$$

Expression (5.1.26) with threshold values of relations of average/mean power of signal and noise will pass into equality.

Taking into account this, and also expression (5.1.23), for determining threshold, we obtain system of equ.:

$$\left.\begin{aligned} q_{\text{вых пор}} &= \frac{(\delta_{\text{вх}})^{\nu}}{\delta_{\text{вых}}}\left[1+q_{\text{вх пор}}\right]^{\nu}-1 \\ q_{\text{вых пор}} &= \beta\, q_{\text{вх пор}} \end{aligned}\right\}. \qquad (5.1.29)$$

Solution of system of equ. (5.1.29 is found usually by graphic methods with known values $\delta_{\text{вх}}$ and $\delta_{\text{вых}}$, which characterize statistical properties of signal and noise.

Being based on given relationships/ratios, in subsequent sections of this chapter let us examine values and interconnection of fundamental parameters of communication system at different methods of modulation.

At present frequency modulation extensively is used in communication systems through ISZ. Pulse-code modulation (KIM), which in combination with the amplitude, frequency or phase modulation of the oscillations of the carrier frequency makes it possible to considerably increase noise immunity and effectiveness of system, is of great interest for the newly developed/processed systems. Furthermore, in the technical literature is discussed a question about the possibility of the transmission of one sideband (OBP), at least in the section from the Earth to ISZ. This interest in OBP is explained by the tendency to shorten the occupied band.

Reasons indicated determine selection of enumerated forms of modulation in subsequent examination.

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5.2. Transmission of one sideband (OBP).

During transmission of OBP obtained as a result of using different diagrams of modulation, transposition of spectrum of initial signal occurs. Therefore width the bands of signal after modulation Δf_c and initial signal ΔF prove to be equal. If the band of receiver Δf is equal to the band of signal, i.e., when

$$\Delta f = \Delta f_c = \Delta F \qquad (5.2.1)$$

in accordance with (5.1.5) we have

$$\nu = 1. \qquad (5.2.2)$$

Analysis, carried out in [5.1] and [5.3] for condition (5.2.1), shows that during transmission of OBP of relationship/ratio between signal and noise it remains without change and therefore

$$\left.\begin{aligned} \beta_{\text{ОБП}} &= 1 \\ \beta'_{\text{ОБП}} &= 1 \\ \eta_{\text{ОБП}} &= 1 \end{aligned}\right\}. \qquad (5.2.3)$$

In actuality, since band of real receiver Δf always is somewhat more than band of signal Δf_c, gain and effectiveness, obtained with OBP, prove to be somewhat less than one. Therefore value (5.2.3) should be assigned to an idealized system during the transmission of OBP.

Comparison (5.2.3) with (5.1.17) and (5.1.20) shows that system with OBP satisfies conditions of ideal system with $\nu=1$. Furthermore, since in the ideal case for the system with OBP $\nu=1$, and $\delta_{вх}=\delta_{вых}$ expression (5.1.26) coincides with (5.1.23); this means that in the system with OBP the threshold is absent.

5.3. Frequency modulation (ChM).

ChM signal at input of receiver in the absence of interferences can be recorded in the form

$$u(t)=U_0\cos[\omega_0 t+\Delta\omega_m S(t)], \qquad (5.3.1)$$

$$S(t)=\int V(t)\,dt. \qquad (5.3.2)$$

Here V(t) - initial signal, transmitted by system of communications (communication/report),

$\Delta f_m=\frac{\Delta\omega_m}{2\pi}$ - maximum change (deviation) in frequency, caused by initial signal, moreover $\Delta\omega_m=K_{чм}V_m(t)$, where $K_{чм}$ - coefficient, which characterizes the rate of rise of modulation characteristic (transmission factor of modulator), and $V_m(t)$ - maximum (peak or quasi-peak) value of initial signal V(t).

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During transmission of harmonic oscillation $1\cdot\cos\Omega t$ with frequency of $F=(\Omega/2\pi)$ value $V(t)=\cos\Omega t$:

$$\Delta\omega_m=K_{чм}, \qquad (5.3.3)$$

$$S(t)=\frac{1}{\Omega}\sin\Omega t. \qquad (5.3.4)$$

Therefore:

$$U(t) = U_0 \cos[\omega_0 t + m \sin \Omega t], \qquad (5.3.5)$$

$$m = \frac{\Delta\omega_m}{\Omega} = \frac{\Delta f_m}{F}. \qquad (5.3.6)$$

Introducing value of peak factor κ, equal to ratio of maximum (peak or quasi- peak) value of transmitted signal to its effective (rms) value, can be obtained $\Delta\omega_m = \Delta\omega_э\kappa$; $m = m_э\kappa$, where $\frac{\Delta\omega_э}{2\pi} = \Delta f_э$ and $m_э$ which are effective (rms) values of deviation of frequency and index of modulation. Values Δf_m and m determine the maximum (peak or quasi-peak) values of the deviation of frequency and index of modulation.

Selection of peak or quasi-peak values Δf_m or m is determined by specific problems. By such, for example, during calculations, connected with the thermal noises, usually are utilized quasi-peak values (values of signal, exceeded during 1% of time), and in the examination of sine waves - peak values. Consequently, in the case of modulation by the sine wave, for which $\kappa = \sqrt{2}$, we will obtain $\Delta f_m = \sqrt{2}\,\Delta f_э$; $\Delta\omega_m = \sqrt{2}\,\Delta\omega_э$; $m = \sqrt{2}\,m_э$.

In considerable excess of signal above noise at input of receiver and absence of distortions instantaneous frequency at output of circuit as derivative of phase, can be found from (5.3.1) and (5.3.2)

$$\omega = \frac{d}{dt}\left[\omega_0 t + \Delta\omega_m \int V(t)dt\right] = \omega_0 + \Delta\omega_m V(t). \qquad (5.3.7)$$

Instantaneous value of signal at output of ideal frequency discriminator, calibrated on deviation of frequency, is

written/recorded in the form

$$U_{вых}(t) = K_{чд} \frac{\omega}{\Delta\omega_m}\Big|_{\omega_0=0} = K_{чд} V(t). \qquad (5.3.8)$$

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Here $K_{чд}$ characterizes mutual conductance of frequency discriminator (transmission factor of frequency discriminator).

As a result of effect on input of receiver of signal and additive noises it is possible to consider that noise will lead to change in parameters of modulation. At the same time in the case of ChM the spurious modulation must be considered by the noises of the phase of oscillations. Parasitic amplitude modulation can be disregarded as noises with the ideal limiter and operation higher than the threshold.

If spectral power (intensity) of fluctuation noises at input of receiver on single resistor/resistance is equal to $\sigma^2_{вх}$, then at output of receiver at frequency F noise intensity on single resistor/resistance [5.1], [5.3]

$$\sigma^2_{вых}(F) = \frac{\sigma^2_{вх}}{U_c^2}\left(\frac{2\pi F}{\Delta\omega_m}\right)^2. \qquad (5.3.9)$$

Here U_c - actual voltage of input signal on single resistor/resistance; therefore $U_c^2 = P_{с вх}$. From (5.3.9) it follows that value $\sigma^2_{вых}$ is proportional to the square of a change in the frequency and can be by the curve of Fig. 5.3.1. The output power of noise in the band of signal frequencies $\Delta F = F_2 - F_1$ (see Fig. 5.3.1),

$$P_{ш\,вых} = \int_{F_1}^{F_2} \sigma^2_{вых}(F)\,dF = \frac{4\pi^2\sigma^2_{вх}}{3U_g^2\Delta\omega_m^2}(F_2^3 - F_1^3). \qquad (5.3.10)$$

Power of noises at input of receiver with band Δf is determined by expression

$$P_{ш\,вх} = \int_{f}^{f+\Delta f} \sigma^2_{вх}\,df = \sigma^2_{вх}\Delta f. \qquad (5.3.11)$$

Since bands of frequencies of receiver Δf and communication/report ΔF are connected with relationship/ratio $\Delta f = \nu\Delta F$, with $F_1=0$ and upon consideration (5.3.11) and (5.3.10) we will obtain

$$P_{ш\,вых} = P_{ш\,вх}\frac{\Delta F^3}{3U_g^2\Delta f_{m\nu}^2} = \left(\frac{P_ш}{P_c}\right)_{вх}\frac{\Delta F^3}{3\Delta f_{m\nu}^2}. \qquad (5.3.12)$$

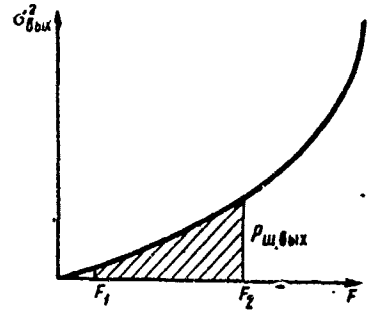

Fig. 5.3.1. Dependence of output power of noise on frequency.

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Expressions (5.3.9), (5.3.10) and (5.3.12) determine power of noises on single output resistance; therefore power of output signal on the same resistor/resistance with single amplitude U_m

$$P_{с\,вых} = \frac{U_m^2}{\varkappa^2} = \frac{1}{\varkappa^2}.$$

After dividing left side of equality (5.3.12) on $P_{с\,вых}$ and right side on $1/\kappa^2$ after conversion we will obtain:

$$\beta_{чм} = 3\frac{\Delta f_m^2 \nu}{\Delta F^2 \varkappa^2} = 3\frac{\Delta f_m^2}{\Delta f^2 \varkappa^2}\nu^3, \qquad (5.3.13)$$

$$\beta'_{чм} = 3\frac{\Delta f_m^2}{\Delta F^2 \varkappa^2} = 3\frac{\Delta f_m^2}{\Delta f^2 \varkappa^2}\nu^2. \qquad (5.3.14)$$

Let us note that for transition/junction from maximum value of deviation of frequency to effective $(\Delta f_э)$ it is possible to utilize

$$\Delta f_m = \Delta f_э \varkappa. \qquad (5.3.15)$$

From expression (5.3.13) and (5.3.14) it follows that value of gain grows/rises proportional to square of deviation. Consequently, obtaining considerable gains with ChM depends on the value of the deviation of frequency, which is defined by both the technical

capabilities of different diagrams and methods of obtaining ChM and by frequency band, permitted or allowed for the transmission of ChM signal.

We will perform a certain transformation of obtained expressions. In the case, when the maximum deviation of frequency Δf_m exceeds the band of frequencies of the transmitted communication/report, i.e., in the case of broadband ChM, it is possible to approximately consider that the optimum band of the receiver

$$\Delta f \cong 2\Delta f_m. \qquad (5.3.16)$$

In this case with substitution in (5.3.13) and (5.3.14) we will obtain:

$$\beta_{\text{чм}} = \frac{3}{4\varkappa^3}\nu^3, \qquad (5.3.17)$$

$$\beta'_{\text{чм}} = \frac{3}{4\varkappa^3}\nu^2. \qquad (5.3.18)$$

Comparing (5.3.17) and (5.3.18) with expressions (5.1.18) and (5.1.19), it is possible to say that during use of frequency modulation communication system on noise immunity is distant from ideal system, since in latter gain from change ν grows/rises considerably sharper than with ChM. Let us note that the gain with ChM in the real receivers proves to be sufficiently to close ones to the values, determined by expressions (5.3.13) and (5.3.17).

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Let us determine gain and generalized gain, obtained during use of ChM for two cases:

1. Transmission of harmonic oscillation with a frequency of F_2 in the band $\Delta F=F_2-F_1$ with $F_1=0$.

On the basis (5.3.13) we will obtain

$$\beta_{\text{ЧМ}}=3\frac{m^2}{\varkappa^2}\nu=\frac{3}{2}m^2\nu. \qquad (5.3.19)$$

Here $m=\frac{\Delta f_m}{F_2}$ index of modulation, $\varkappa=\sqrt{2}$.

From (5.3.14) it follows

$$\beta'_{\text{ЧМ}}=\frac{1}{\nu}\beta_{\text{ЧМ}}=\frac{3}{2}m^2. \qquad (5.3.20)$$

2. Transmission in band ΔF of communications/reports with identical bands, shifted relative to each other in frequency, moreover band of each message $\Delta F_{\text{к}}=\frac{\Delta F}{N}$ (see Fig. 5.3.2). This case corresponds to the frequency modulation of the oscillations of transmitter by multichannel telephone signal in frequency division multiplex N of telephone channels. If we designate total maximum voltage of all N channels at the output of the receiver as U_{mN}, and the maximum voltage in any channel through $U_{m\text{к}}$, then noise intensity at the output of receiver at frequency F, it is analogous (with 5.3.9), it can be recorded as

$$\sigma^2_{\text{вых}}(F)=\frac{\sigma^2_{\text{вх}}}{U_{\text{с}}^2}\left(\frac{U_{mN}}{U_{m\text{к}}}\right)^2\left(\frac{2\pi F}{\Delta\omega_{mN}}\right)^2. \qquad (5.3.21)$$

Output power of noise in band of frequencies of channel k with central frequency of $F_{\text{к}}$ will be

$$P_{\text{ш вых}}=\int_{F_{\text{к}}-\frac{\Delta F_{\text{к}}}{2}}^{F_{\text{к}}+\frac{\Delta F_{\text{к}}}{2}}\sigma^2_{\text{вых}}(F)\,dF=\frac{\sigma^2_{\text{вх}}}{U_{\text{с}}^2}\left(\frac{U_{mN}}{U_{m\text{к}}}\right)^2\frac{4\pi^2}{\Delta\omega^2_{mN}}F_{\text{к}}^2\Delta F_{\text{к}}\left\{1+\frac{\Delta F_{\text{к}}^2}{12F_{\text{к}}^2}\right\}. \qquad (5.3.22)$$

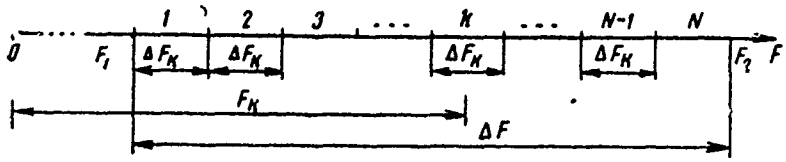

Fig. 5.3.2. Arrangement/position N of channels in band ΔF.

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Usually $\Delta F_к \ll F_к$, therefore value $\frac{\Delta F_к^2}{12F_к^2}$ in comparison with one can be disregarded/neglected. This approximation/approach indicates that spectral noise density in the limits of the band of channel it is considered uniform and equal spectral density at frequency $F_к$, which is located in the middle of the band of channel.

Taking into account that

$$\frac{U_{mN}}{U_{mк}} = \frac{\Delta\omega_{mN}}{\Delta\omega_{mк}} = \frac{\Delta f_{mN}}{\Delta f_{mк}}, \qquad (5.3.23)$$

where $\Delta f_{mк} = \frac{\Delta\omega_{mк}}{2\pi}$ - maximum deviation of frequency, which falls to one channel, let us find value of gain for one of channels

$$\beta_{чм\,к} = \frac{\Delta f_{mк}^2 \Delta f}{\Delta F_к F_к^2 \cdot 3}. \qquad (5.3.24)$$

Taking into account (5.3.15), it is possible to pass from maximum value of deviation in one telephone channel $\Delta f_{mк}$ to effective value of deviation $\Delta f_к$ for one channel, then

$$\beta_{чм\,к} = \frac{\Delta f_к^2 \Delta f}{\Delta F_к F_к^2}. \qquad (5.3.25)$$

From (5.3.24) it follows that with identical values of deviation of frequency channels have not identical gain. The higher the number

of channel, i.e., the higher the central frequency $F_{ц}$, the less the gain, obtained using ChM. For the equalization of gain is introduced the "predistortion" of the spectrum, i.e., for each channel different deviation of frequency is established/installed, in consequence of which a change in the gain in different channels they prove to be in the limit inferiors. Let us recall that the resulting expressions, which determine gain with ChM, prove to be valid only then, when the power of signal at the input of receiver or the relation of the average/mean power of signal and noise at the input of receiver ($q_{вх}$) prove to be above threshold value. In the case, if this is not fulfilled, the gain, given during the use of frequency modulation, substantially is reduced. Consequently, during the construction of the communication systems with ChM the value of threshold level plays the decisive role during the selection of energy relationships/ratios (power of transmitter, the gain of antennas, noise temperature of receiver, etc.).

Computation of threshold values for relations of average/mean power of signal and noise at input and output of receiver ($q_{вх}$ and $q_{вых}$), can be carried out according to expressions (5.1.29). Let us note that during the computations appear the considerable difficulties during the determination of values $\delta_{вх}$ and $\delta_{вых}$, determined by the statistical characteristics of random processes.

Energy spectrum of real process at input of receiver depends on such parameters as signal-to-noise ratio, width of band of receiver,

value of deviation of frequency, and also on statistical properties of communication/report and noise.

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Energy spectrum at the output of receiver, furthermore, it depends on the form of the amplitude characteristic of limiter and its frequency characteristic. The numerous investigations of threshold values $q_{вх}$ and $q_{вых}$, carried out in different works, among which one should note [5.4], [5.6] and [5.1], are shown to the sufficiently good coincidence of theoretical and practical results.

Fig. 5.3.3 gives experimental threshold curves during the reception of ChM of signals [5.6], which characterize dependence $q_{вых}$ on $q_{вх}$, moreover noise at output is measured in the absence of modulation of signal, and noise at input of receiver is measured in entire band of circuit, which precedes frequency discriminator.

Value $q_{вых}$ is calculated for value of index of modulation m=1 and with different values ν. If value $\nu>2$, then system can work with values of m>1; however, in this case, to numerical values $q_{вых}$, calculated along the axis of ordinates, one should add 10 lg m^2. The threshold, with which the relationship/ratio between $q_{вх}$ and $q_{вых}$ differs from linear more than by 1 dB, in Fig. 5.3.3 is designated by points. Let us note that more to the right these points proves to be valid expression (5.3.13) The examination of Fig. 5.3.3 shows that depending on value ν the value of threshold value $q_{вх\,пор}$ varies from

3.5 to 12 dB, i.e., from 2.5 to 16 times, moreover the large threshold signal will correspond to large values of the index of modulation m. In order to ensure the gain, given by frequency modulation, it is necessary to satisfy the condition

$$q_{вх} \geqslant q_{вх\ пор}. \qquad (5.3.26)$$

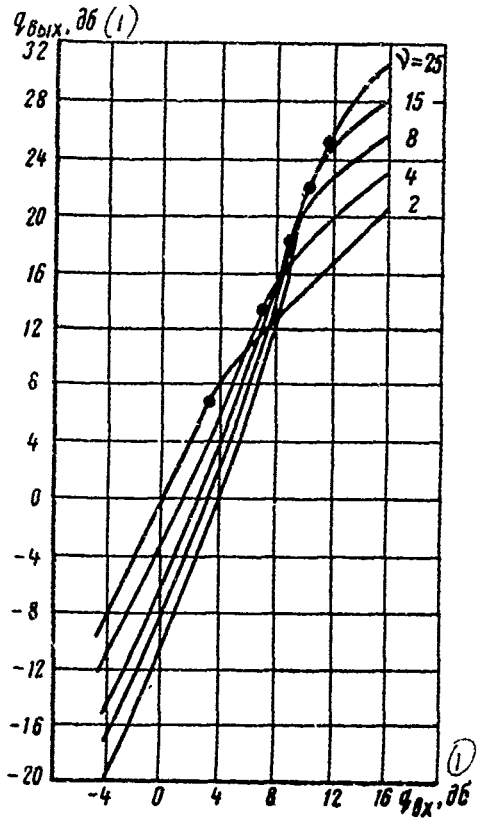

Fig. 5.3.3. Dependence $q_{вых}$ on $q_{вх}$ with ChM. Values along the axis of ordinates are indicated with the index of modulation m=1; for other m to the values along the vertical axis to add 10 lg m^2,

Key: (1). dB.

Taking into account given above threshold values, and also the fact that according to introduced designations $P_{\text{с вх}}=q_{\text{вх}}P_{\text{ш вх}}$ we obtain following condition, determining the power of signal at input of receiver, with which is ensured calculated gain β:

$$P_{\text{с вх}} \geqslant (2{,}5 \div 16)\, P_{\text{ш вх}} = (2{,}5 \div 16)\, \kappa T_{\text{э}}\, \Delta f. \qquad (5.3.27)$$

Here $T_{\text{э}}$ - total temperature of receiver (see Chapter 4), and Δf is width of band of receiver at input and with accuracy sufficient for practice is determined by expression

$$\Delta f = 2F(1+m). \qquad (5.3.28)$$

In the case, when $\Delta f_m \gg F$, expression (5.3.28) passes in (5.3.16).

It is obvious that during design of system they attempt to ensure its normal operation with possible smaller values of signal at input of receiver $P_{\text{с вх}}$. From the graphs, given in Fig. 5.3.3, and expression (5.3.28) it is possible to do the conclusion that in this case is desirable to have low values ν. However, with the assigned band of initial signal reduction in value ν will lead to the decrease of the band of receiver Δf, consequently, to the smaller values of deviation, which according to expression (5.3.13) will lead to reduction in the value of gain β. Thus, the selection of the low values of the parameter ν (or narrow-band modulation), although it permits confidently to work with the low values of signal in the hyperthreshold region, nevertheless will lead to reduction in the gain and, in the final analysis, to reduction in qualitative indices of the

system of communications - value q_{max}.

For operational provisions in hyperthreshold region at least possible power of signal at input of receiver, in accordance with expression (5.3.27), it is possible to plan following paths:

a) reduction in equivalent noise-ache temperature $T_{ш}$, using low-noise input devices of receiver, antenna systems with small side-lobe level, high-quality feeders and the like;

b) change in coefficients, which stand in brackets in expression (5.3.27), due to redistribution of energy of noise along spectrum with subsequent filtration. This can be realized by the limiters, which have the amplitude characteristic, close to the ideal, after which is established/installed the filter with the bandwidth, equal to the band of the signal frequencies and noise on the input of limiter. The numerical ratios between the signal and the noise at input and output of this system limiter - filter, proposed by Bagdadi [5.7], were established/installed in work [5.8] by Davenport. The efficiency of this method depends substantially on the frequency characteristic of the filter, established/installed after the limiter: the stronger the limitation, the wider must be the uniform part of the characteristic.

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This is necessary so that the effect of limitation would not be distorted by selection in the frequency, since precisely the high-frequency components of the spectrum after limiter, subjects to the input of discriminator (device/equipment of the realizing of

conversion of ChM oscillations/vibrations into AM oscillation/vibration), characterize the quality of limitation. Therefore filter with too narrow a passband or nonuniform frequency characteristic weakens/attenuates these components and worsens/impairs the action of limiter. Similarly must appear the frequency characteristic of discriminator. As show experiments [5.7], the series connection of several limiters with the filters makes it possible to additionally decrease the numerical coefficient, entering expression (5.3.27);

c) considerable quantity of the methods of reduction in the threshold value of signal is based on the decrease of the band of frequencies of receiver Δf. With the large indices of modulation this is realized with the help of the special diagrams in such a way that the instantaneous value of the frequency of ChM signal at any moment of time would prove to be within this band. Thus, for example, if the broadband frequency modulation of oscillations is realized, then the reception of this oscillation can be carried out by means of the narrow-band receiver, whose band is moved according to frequency dial ("follows") in accordance with the instantaneous value of the frequency of received signal. This displacement/movement of the band of frequencies of the receiver can be realized by an application of voltage from the output of frequency discriminator to the heterodyne or the control system, which reconstructs the narrow-band filter of receiver. The phase of this voltage/stress in entire frequency band of the transmitted communication/report must be such, in which is ensured a precise sequence of the frequency of heterodyne or band of

frequencies of the retunable filter after the instantaneous frequency of the taken ChM oscillation. Under these conditions is feasible the undistorted reception of ChM oscillation even if $\Delta f < \Delta f_m$.

Consequently, reception of ChM oscillations by narrow-band receiver with "servo" heterodyne (this method is called method of reception with feedback in frequency - OSCh) or receiver with "servo" filter, established/installed in high or intermediate frequency, is feasible to special receivers with known band of frequencies of transmitted communication/report or, in other words, at known maximum speed of change in frequency of ChM. For reduction in the threshold with the reception of broadband ChM oscillations, besides methods with OSCh and servo filter, are used some other methods of reception - with frequency divider, with the addition of cophasal or orthogonal oscillation relative to the carrier of the taken ChM oscillation, etc. The examination of different methods of reduction in the threshold is given in works [5.9], [5.11].

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Application of OSCh or other methods makes it possible to sufficiently substantially reduce threshold value of signal. In particular, the application of OSCh, examined in work [5.10], made it possible to lower threshold to 9 db. In this case in the absence of OSCh the index of modulation was equal to m=10, which corresponded to value $\nu \cong 22$. In work [5.11] it is shown that the application of OSCh makes it possible to obtain further gain from reduction in the

threshold

$$\beta_{доп} = 3{,}5m^{0{,}3}. \qquad (5.3.29)$$

Here m - index of modulation. With m=5, 10 and 20 respectively are obtained the gains, with respect to threshold equal to 6 db, 7.5 db and 9 db.

In section 10.2 examination of schematics of receivers, which make it possible to carry out reduction in threshold, is given.

Let us pass to examination of efficiency of communication systems during use of ChM. For this into expression (5.1.12) should be substituted the value of gain from (5.3.13) or, if is examined broadband ChM, from (5.3.17). Being limited to the examination of the case of the transmission of harmonic oscillation with broadband ChM, when $\Delta f \cong 2\Delta f_m$, and value $\nu \cong 2m$, after substitution (5.3.19) we find

$$\eta = \frac{1}{2m} \frac{\log_2(3m^3 q_{вх} + 1)}{\log_2(q_{вх} + 1)}. \qquad (5.3.30)$$

On the basis of this expression it is possible to draw conclusion that efficiency of system is raised with decrease of $q_{вх}$. If we take into account that value $q_{вх}$ compulsorily must exceed threshold value, it is possible to conclude that with the work of system near the threshold value the efficiency of system will be more. Thus, reduction in the threshold value will contribute to an increase in the efficiency of the communication system.

Usually value $q_{вых}$ for system is that specified or by norms, presented to communication channel. For increasing the efficiency of system its parameters and, in particular, the gain β should be chosen so that would be satisfied the following conditions:

$$q_{вых\ зад} = q_{вых\ пор} = \beta\, q_{вх\ пор}. \qquad (5.3.31)$$

In this case it is necessary to take measures, which ensure reduction threshold value of system. If the parameters of system are such, that $q_{вых\ зад} \gg q_{вых\ пор}$, then reduction in the threshold values will not lead to an increase in the efficiency of system.

Let us construct graphic dependence of efficiency of system on $q_{вх}$. With ChM the following expressions, which determine width of band of signal and index of modulation are valid in general form:

$$\Delta f = 2F(m+1),$$
$$m = \frac{\Delta f_m}{F}, \qquad (5.3.32)$$

where F - frequency of the transmitted.

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Considering $F=F_2$, and band of optimum receiver equal to band of signal on its input, let us find

$$\nu = \frac{\Delta f}{F} = 2(m+1). \qquad (5.3.33)$$

After determining from this expression value m and taking into account (5.3.19), after substitution in (5.1.12) we will obtain

$$\eta = \frac{1}{\nu} \frac{\log_2\left[\frac{3\nu}{2}\left(\frac{\nu}{2}-1\right)^2 q_{вх}+1\right]}{\log_2[q_{вх}+1]}. \tag{5.3.34}$$

Fig. 5.3 gives dependence $\eta=f(q_{вх})$.

5.4. Pulse-code modulation (KIM).

With KIM at transmitting end/lead of line of communications initial signal is converted into code combinations of groups of pulses, and then by these groups is realized modulation of oscillations of carrier frequency. Depending on the form of modulation let us agree to designate subsequently through KIM-AM amplitude modulation (manipulation) of oscillations with words of pulses/momenta, through KIM-ChM - the frequency modulation of oscillations with words of pulses/momenta, and through KIM-FM - phase.

Clock period T_i, through which follow words of pulses/momenta, in accordance with V. A. Kotelnikov's theorem, generalized to stationary random functions with limited energy spectrum [5.4], is determined by band of transmitted communication/report ΔF as follows:

$$T_i \leqslant \frac{1}{2\Delta F}. \tag{5.4.1}$$

Since $\Delta F=F_2-F_1$, where F_2 and F_1 respectively greatest and smallest frequency of transmitted communication/report, with $F_1=0$ we will obtain

$$T_i \leqslant \frac{1}{2F_2}. \tag{5.4.2}$$

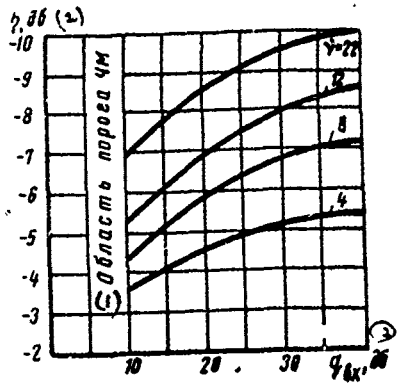

Fig. 5.3.4. Efficiency of ChM.

Key: (1). Region of threshold of ChM. (2). db.

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Each character corresponds to specific quantized level of communications/reports and ccnsists of n isochronous pulses/momenta. During the binary coding it is possible to transmit a number of levels

$$L = 2^n. \qquad (5.4.3)$$

Thus, with KIM initial signal is replaced by quantified, which only approximately reflects initial. Because of this quantization noises appears. The relation of the average/mean power of communication/report P_c and quantization noises $P_{шL}$ they depend on a number of levels of quantization [5.1]

$$q_L = \frac{P_c}{P_{шL}} = \frac{3L^2}{\varkappa^2} = \frac{3 \cdot 2^{2n}}{\varkappa^2}. \qquad (5.4.4)$$

Here κ - peak factor of transmitted signal. During the transmission of sinusoidal oscillation $\kappa=\sqrt{2}$; therefore

$$q_L = 3 \cdot 2^{2n-1}. \qquad (5.4.5)$$

From (5.4.4) and (5.4.5) it follows that quantization noises will be sharply reduced with an increase of the number of pulses/momenta in word. In this case, however, will increase the band of frequencies $\Delta F_{кв}$, occupied by the quantized communication/report. The latter can be determined as follows. We will consider that along the communication system with the help of code combinations, which contain n pulses/momenta with duration τ, it is necessary to transmit N of the initial signals, each of which has a band of frequencies Δf. Therefore the band of frequencies of the quantized signal, which requires for the transmission of pulses/momenta by duration τ, is determined by the expression

$$\Delta F_{кв} = \frac{a}{\tau}. \qquad (5.4.6)$$

Here a>1 is coefficient, which depends on permissible distortion of shape of pulse.

$$\tau = \frac{T_l}{nNQ}. \qquad (5.4.7)$$

Duration of pulses τ is determined by obvious relationship/ratio:

Here Q>1 is determined by porosity of sequence of pulses/momenta. Taking into account (5.4.1) and (5.4.6), let us find width of band of the quantized signal

$$\Delta F_{кв} \geqslant 2\Delta FanNQ. \qquad (5.4.8)$$

Consequently, during transmission of one initial signal (N=1) frequency band after quantization increases more than in 2n times.

During transmission along channel of communication of quantized signal at output of receiver, besides signal, thermal noises appear. The relation of the average/mean power of signal and noise at the output with any form of modulation of the oscillations of transmitter characters is determined by following expression [5.1]:

$$q_{вых} = \frac{3}{4} \frac{(L-1)^2}{x^2(L^2-1)\Delta F T_t} \frac{1}{1-\Phi(\alpha)}. \qquad (5.4.9)$$

Here $\Phi(\alpha) = \frac{2}{\sqrt{2\pi}} \int_0^{\alpha} e^{-\frac{x^2}{2}} dx$ - probability integral, which through value α is found from tables. The parameter α^2 determines the ratio of the medium energy of one pulse/momentum to the specific power of noise, consequently, α^2 is determined by the form of modulation of the oscillations of the carrier frequency. Let us note that entering expression (5.4.9) value $\frac{1}{2}[1-\Phi(\alpha)] = p(\alpha)$ determines the probability of the erroneous reception of the elementary impulses/transmissions of code combination.

Let us now move on to refinement of values α with different types of carrier modulation. Accordingly given determination

$$\alpha^2 = \frac{P_{с вх} \tau}{2} : \frac{P_ш}{\Delta f} = q_{вх} \frac{\tau \Delta f}{2}. \qquad (5.4.10)$$

With KIM-AM optimum band of receiver Δf, equal to band of frequencies of radio pulses, it twice exceeds band of quantized signal. Therefore upon consideration (5.4.6) for KIM-AM let us find

$$\Delta f = 2\Delta F_{кв} = 2\frac{a}{\tau}. \qquad (5.4.11)$$

Keeping in mind (5.4.8) with Q=1, we will obtain

$$\Delta f = 4\Delta FnNa. \qquad (5.4.12)$$

Substituting (5.4.11) in (5.4.10) with a=1, we find that with KIM-AM

$$\alpha^2 = q_{вх}. \qquad (5.4.13)$$

With analogous reasonings it is possible to obtain for case KIM-ChM

$$\alpha^2 = 2q_{вх}. \qquad (5.4.14)$$

With KIM-FM let us find

$$\alpha^2 = 4q_{вх}. \qquad (5.4.15)$$

Being assigned by different values $q_{вх}$ according to given formulas it is possible to determine value α for different cases and to find probability of erroneous reception $p(\alpha)=\frac{1}{2}[1-\Phi(\alpha)]$.

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This dependence is given in Fig. 5.4.1. the gain, depending on the method of modulation, can be determined from (5.4.9) upon consideration (5.4.7) and (5.4.10) according to the expression

$$\beta_{КИМ} = \frac{q_{вых}}{q_{вх}} = \frac{3(L-1)^2\nu}{8\lambda^2(L^2-1)\alpha^2 nN}\frac{1}{1-\Phi(\alpha)}. \qquad (5.4.16)$$

Here parameter ν is located, on the basis of band of signal, i.e.

$$\nu = \frac{\Delta f}{F}. \qquad (5.4.17)$$

In the case, when $\alpha >> 1$,

$$\Phi(\alpha) \cong 1 - \frac{2}{\sqrt{2\pi}} \frac{e^{-\frac{\alpha^2}{2}}}{\alpha}. \qquad (5.4.18)$$

Taking into account (5.4.8) with L>>1 and equality in expression (5.4.1), instead of (5.4.9) we will obtain

$$q_{вых} \cong \frac{2}{\chi^2} \alpha e^{\frac{\alpha^2}{2}}.$$

Taking into account latter/last expression and equalities (5.4.13), 5.4.14), 5.4.15), let us find value of gains $\beta = \frac{q_{вых}}{q_{вх}}$, with different methods of modulation.

With KIM-AM

$$\beta = \frac{2}{\chi^2 \sqrt{q_{вх}}} e^{\frac{q_{вх}}{2}}. \qquad (5.4.19)$$

With KIM-ChM

$$\beta = \frac{2\sqrt{2}}{\chi^2 \sqrt{q_{вх}}} e^{q_{вх}}. \qquad (5.4.20)$$

With KIM-FM

$$\beta = \frac{4}{\chi^2 \sqrt{q_{вх}}} e^{q_{вх}}. \qquad (5.4.21)$$

Corresponding values of generalized gains can be obtained, if resulting expressions are divided into value ν, which is determined by expression (5.4.17).

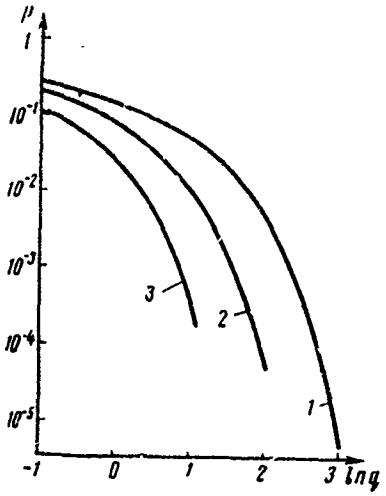

Fig. 5.4.1. Dependence of probability of erroneous reception on 1 - KIM-AM ; 2 - KIM-ChM; 3 - KIM-FM.

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According to expressions (5.4.19)-(5.4.21) are constructed the graphic dependences, given in Fig. 5.4.2, on which it is possible to do conclusions about the value of the gains, which are obtained in the various forms of modulation by code combinations of groups of pulses.

Real gain of systems with KIM proves to be less than designed according to expression (5.4.16) or according to expressions (5.4.19)-(5.4.21). For KIM-AM the ratio of real gain to the ideal ξ on the level of pulse clipping in the receiver, equal to 0.5, is given in Table 5.4.1 [5.3].

Let us consider efficiency of system in the case of KIM. Substituting (5.4.19)-(5.4.21) into expression (5.1.12) and keeping in

mind the transmission of the sine wave, for which $\kappa^2=2$, we obtain for KIM-AM

$$\eta = \frac{1}{\nu} \frac{\log_2\left[q_{вх} e^{\frac{q_{вх}}{2}} + 1\right]}{\log_2[q_{вх} + 1]}, \quad (5.4.22)$$

for KIM-ChM

$$\eta = \frac{1}{\nu} \frac{\log_2[\sqrt{2} q_{вх} e^{q_{вх}} + 1]}{\log_2[q_{вх} + 1]}, \quad (5.4.23)$$

for KIM-FM

$$\eta = \frac{1}{\nu} \frac{\log_2[2q_{вх} e^{2q_{вх}} + 1]}{\log_2[q_{вх} + 1]}. \quad (5.4.24)$$

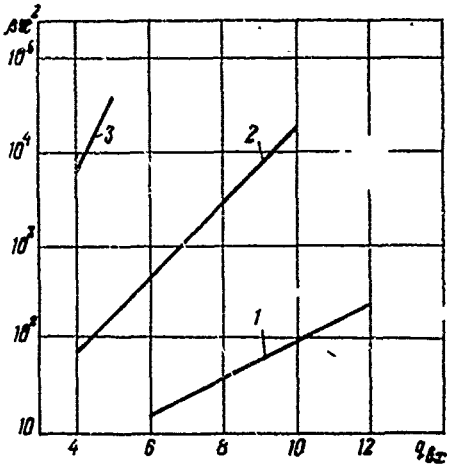

Fig. 5.4.2.

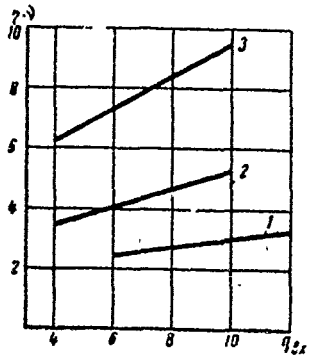

Fig. 5.4.3.

5.4.2. Dependence $\beta\kappa^2$ on $q_{эх}$: 1 - KIM-AM , 2 - KIM-ChM, 3 - KIM-FM.

Fig. 5.4.3. Dependence $\eta\nu$ on $q_{эх}$: 1 - KIM-AM , 2 - KIM-ChM, 3 - KIM-FM.

Table 5.4.1.

α	1	2	3	4	5
	1	4	9	16	25
ξ, дб	—3,6	—5,5	—7	—9,75	—19,2

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Results of calculations of value $\eta\nu$ from $q_{эх}$ according to given formulas are given in Fig. 5.4.3, whence it follows that efficiency of system little is changed with constant value ν.

5.5. Relationships/ratios for the selection of the fundamental parameters of the communication system.

In communication system through ISZ are two sections of

connection/communication: section Earth-isz, which subsequently we will call the first and designate by index 1 and section isz-Earth, determined by index 2.

Thermal noises, which appear for each of these sections are statistically independent; therefore on outlet of channel powers of noises of individual sections will store/add up. It is obvious that the total noise must not exceed the assigned or permissible value.

Passing from permissible amount of power of thermal noises to relations of power of noise to signal in individual sections, it is possible to record

$$\frac{1}{q}=\frac{1}{q_{\text{вых1}}}+\frac{1}{q_{\text{вых2}}}. \qquad (5.5.1)$$

Here $q_{\text{вых2}}$ - ratio of average/mean power of signal to noise respectively at output of terrestrial receiver with work only of second section, $q_{\text{вых1}}$ - at output of airborne receiver, q - relationship at output of terrestrial receiver with work of entire circuit.

We will consider that

$$\left.\begin{aligned} q_{\text{вых1}} &= aq \\ q_{\text{вых2}} &= bq, \end{aligned}\right\}, \qquad (5.5.2)$$

where a and b - coefficients, which consider relationship/ratio between noises, which are introduced in different sections. Substituting (5.5.2) in (5.5.1), let us find

$$\frac{1}{a}+\frac{1}{b}=1.$$

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Graph, constructed according to latter/last expression, is given in Fig. 5.5.1, on it it is possible to select relationship/ratio a and b. Since the power of onboard transmitter is always lower than the power of terrestrial, it is expedient so that basic part of the noises would be determined by the second section. From this it follows that values for b should be chosen close to one, for example, within limits of 1.05-1.2; then the value of coefficient a will be in limits of 6-20.

Taking into account (5.1.3) and (5.5.1), we will obtain following relationships/ratios:

$$\frac{1}{q}=\frac{1}{q_{вх1}\beta_1}+\frac{1}{q_{вх2}\beta_2}, \tag{5.5.3}$$

$$q_{вх1}=\frac{aq}{\beta_1}. \tag{5.5.4}$$

$$q_{вх2}=\frac{bq}{\beta_2}. \tag{5.5.5}$$

Here $q=\left(\frac{P_c}{P_ш}\right)_{вых}$ - relation of average/mean power of signal and noise at output of communication channel. With normal conditions should be implemented inequality $q\geqslant q_{доп}$. where $q_{доп}$ - allowed value q.

Entering latter/last expressions values β_1, β_2 - gains, which are obtained respectively at output of terrestrial receiver (with work only of section satellite - Earth) and at output of onboard receiver. On the basis of expressions (3.1.1) and (4.5.1), it is possible to

record (5.5.3) in the form

$$\left(\frac{P_{ш}}{P_{с}}\right)_{max} = \frac{kT_{э6}\,\Delta f_{1} B_{Ш1}}{P_{пз}\,\beta_{1}} + \frac{kT_{э_{3}}\,\Delta f_{2} B_{Ш2}}{P_{п6}\,\beta_{2}}. \qquad (5.5.6)$$

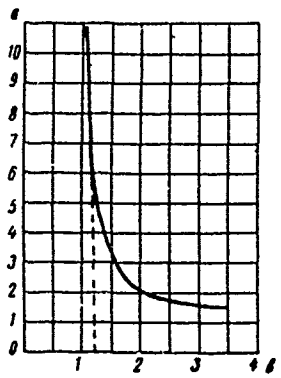

Fig. 5.5.1. Distribution of relationship/ratio between shims on sections of line of communications.

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Here B_{Σ_1} and B_{Σ_2} designate total attenuations between transmitter and receiver, determined from expression (3.1.2) respectively for sections Earth-satellite (index 1) and satellite-Earth (index 2), and indices in all other designations determine parameters of onboard (b) or terrestrial (3) equipment. Usually

$$B_{\Sigma 1} \cong B_{\Sigma 2}. \tag{5.5.7}$$

Therefore in the case, when

$$\frac{T_{\Sigma б}\,\Delta f_1}{P_{п.}\beta_1} \ll \frac{T_{\Sigma з}\,\Delta f_2}{P_{пб}\beta_2}, \tag{5.5.8}$$

it is possible to be bounded to calculation of noises only in section 2, i.e. to carry out calculation instead of expression (5.5.6) according to ф-le

$$\left(\frac{P_ш}{P_c}\right)_{вых} \cong \frac{\kappa T_{\Sigma з}\,\Delta f_2}{P_{пб}\beta_2}, \tag{5.5.9}$$

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Chapter 6.

Principles of multiplexing of lines of communications.

6.1. Characteristics of signals.

In examination of principles of multiplexing lines of communications, which make it possible to carry out simultaneous transmission along one channel of communication of several telephone or TV signals, it is necessary to consider fundamental characteristics of signals. In accordance with this should be in the most general form examined the following characteristics of telephone signals, TV signal and signal of the sonic tracking:

- signal aspect,
- the frequency band, occupied by signal,
- dynamic range.

Telephone signal is random process with zero average/mean value. For the faultless reproduction of Russian speech the frequency band of 65-10000 Hz [6.7] is necessary, during the use of a frequency band within the limits from 100 to 6000 Hz the reproduction of speech is recognized by good. One should note that the contraction of the frequency band to the values of 300-3400 Hz, accepted by MKKTT

[International Telegraph and Telephone Consultative Committee] in 1946, barely decreases the intelligibility of speech, although somewhat decreases its naturalness.

Phase distortions little affect quality of telephone communication. Therefore in the band of frequencies of the channel an admissibly considerable change of the group propagation time, which is determined derived phase-frequency characteristic. A difference in the group propagation time at the highest transmitted frequency and at the medium frequencies must not exceed 30 ms.

Signal of sonic tracking so, as telephone signal presents random process with zero average/mean value. Numerous investigations showed that for a good reproduction of different musical products the frequency band from 50 to 10000 Hz (I class) was sufficient, satisfactory reproduction is obtained with the band 100-6000 Hz (II class) [6.8].

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Phase distortions virtually do not affect quality of reproduction of signal of sonic tracking.

TV signal (video signal) consists of picture signal and signals of synchronization, i.e., is defined both by random and determined functions. Fig. 6.1.1 shows the possible form of the oscillogram of one row of video signal. From the figure it follows that the video

signal contains constant component, whose value changes in the dependence on the illumination or the brightness of the transmitted image. One sh uld note that according to the standard accepted picture signal occupies 70% of spread/scope of video signal; 30% of spread/scope of video signal occupy timing pulses.

In accordance with Fig. 6.1.1 repetition period of row of video signal on standard, accepted in USSR and which corresponds to recommendations of MKKR [International Radio Consultative Committee], is 64 μs. During one frame by the standard are transmitted 625 rows, in one second - 25 frames/personnel (50 fields), which corresponds to the transmission of 15625 rows. Clock line-scanning frequency is equal to 15625 Hz.

For TV signal bandwidth, accepted by MKKR for communication systems through ISZ, is 5 MHz. The analysis of the power distribution along the spectrum shows that in video signal basic part of the power is concentrated in the frequency region of below 200-250 kHz, and approximately 5-15% of entire power - in the frequency band of above 1.2-1.7 MHz. One should note that near the harmonics of horizontal frequency the concentration of power is observed, and in the middle of the gap/interval between the adjacent harmonics the power of signal is close to zero. Especially sharply this phenomenon is developed at the frequency region, which lie below 250-400 kHz. Therefore in one band with the TV signal the transmission also of other signals is possible. This is utilized during the transmission of color television, and in some systems - for the transmission of sonic tracking.

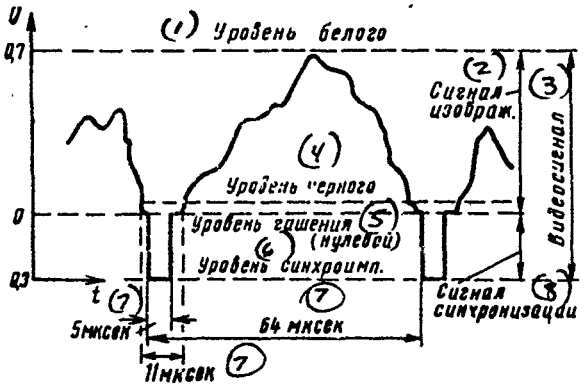

Fig. 6.1.1. Oscillogram of video signal.

Key: (1). White level. (2). Picture signal. (3). Video signal. (4). Black level. (5). Level of extinguishing (zero). (6). Level of synchro pulse. (7). μs. (8). Signal of synchronization.

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In connection with the fact that amplifiers widely used in communication systems do not pass constant component, during transmission of video signal it is necessary to take special measures, which make it possible at point of reception to restore/reduce its level. For this the diagrams of fixation of one of the levels of TV signal (apex/vertex level of timing pulses or extinguishing pulses/momenta) can be used. Let us note that fixation of the level of video signal must be realized directly in control electrode of modulator (for example, in the grid of reactance tube, in the repeller of klystron, etc.), since otherwise the linear section of the characteristics of modulator and demodulator must be expanded, approximately, one and a half times.

If communication channel must be utilized both for transmission of multichannel telephone signal and for transmission of TV signal, application of clamping circuits is undesirable, since this disrupts universality of channel. In this case can be used of the predistorting filters on the side transmissions (to the modulator) and those restoring - on the side of reception (after detector). The introduction of the predistorting filter reduces the amplitude of the low-frequency components of TV signal. Filter can be established/installed at any point of video circuit, i.e., in any cascade/stage, which precedes modulator, while the installation of the schematic of fixation of the level must be realized directly in the modulator. Therefore the predistorting filters, established/installed in the television video amplifiers, will not be utilized during the transmission of multichannel telephone signal. This makes it possible to create the universal communication channel with one and the same modulators and demodulators.

Fig. 6.1.2 gives amplitude-frequency characteristic of predistorting filter, recommended by MKKR for radio relay lines with ChM during transmission of TV signal with standard of 625 rows. The amplitude-frequency characteristic of the restoring filter must be reverse/inverse to the characteristic, shown in Fig. 6.12.

At present for communication systems through ISZ norms to characteristics of predistorting filter are absent.

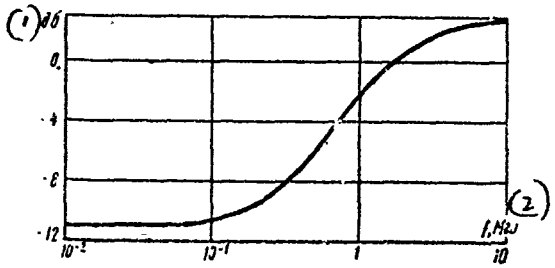

Fig. 6.1.2. Amplitude-frequency characteristic of predistorting filter.

Key: (1). dB. (2). MHz.

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6.2. Principle of the construction of equipment for frequency division multiplex during the transmission of telephone signals.

Multiplexing is necessary in order along one line of communications to carry out transmission of several signals. With the frequency division multiplex the bias/displacement (transposition) of the spectrum of the transmitted signals is utilized. So, for example, if it is necessary to transmit the telephone signals of several subscribers, each of whom occupies one and the same frequency spectrum from $F_{мин}$ to $F_{макс}$, then with the aid of the equipment for frequency division multiplex the transposition of these spectra is accomplished then, as shown in Fig. 6.2.1. From Fig. 6.2.1 it follows that the width of the spectrum of each signal, equal to $\Delta F_к = F_{макс} - F_{мин}$, during the transposition it remains constant/invariable. The spectra of the signals of individual subscribers (1, 2, 3, 4, ..., P, M, ..., N) on the side of transmission are shifted/sheared in the frequency in such

a way that between them are formed some protective belts, necessary for eliminating interferences between the transmitted signals and facilitating the separation of line spectrum by filters during inverse transformation on the side of reception.

In accordance with recommendations of MKKTT each telephone signal occupies frequency band from 300 Hz to 3400 Hz, and width of band $\Delta F_к = 3400 - 300 = 3100$ Hz.

Total band, determined by sum of bands of telephone signals and by protective belts between adjacent signals, is equal to 4 kHz. Therefore during the transmission of telephone signals along the N channels the width of line spectrum will comprise

$$\Delta F = F_2 - F_1 = 4N \text{ кгц.} \quad (1)$$

Key: (1). kHz.

Here F_2 and F_1 respectively upper and lower frequencies of line spectrum. One should note that with a considerable number of channels between the adjacent groups of the channels are introduced wider guard bands of frequencies than between the adjacent individual channels; this somewhat simplifies the execution of filters in the equipment for multiplexing and channel separation. Therefore, beginning with N=300, it is possible to accept

$$\Delta F = F_2 - F_1 \cong 4{,}2N \text{ кгц.} \quad (1) \qquad (6.2.1)$$

Key: (1). kHz.

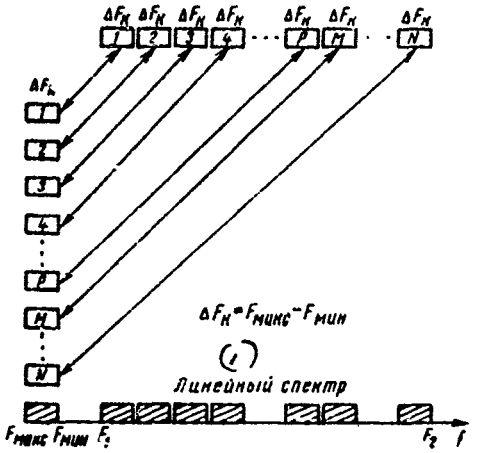

Fig. 6.2.1. Conversion of spectra with frequency division multiplex.
Key: (1). Line spectrum.

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Fig. 6.2.2 gives block diagram, which elucidates principle of construction of equipment with frequency division multiplex and formation of telephone channel with number p. Telephone signal with spectrum $\Delta F_к = F_{макс} - F_{мин}$ from subscriber P passes through differential system ДС and it proceeds to the balanced modulator БМ. The same modulator the oscillation with a frequency of F_p simultaneously is supplied. As a result of modulation of variations of frequency F_p by signal with spectrum $F_{мин} \div F_{макс}$ at the output of БМ are formed the oscillations of upper side bands $F_в = F_p + (F_{мин} \div F_{макс})$ and lower side bands $F_н = F_p - (F_{мин} \div F_{макс})$.

Oscillations either of upper or lower side bands pass through band-pass filter $ПФ_p$ and after amplifier УС connections/communications are supplied into line.

Oscillations, which passed line of communications, amplifier УС and band-pass filter $ПФ_p$, proceed to demodulator Д of equipment for multiplexing. Simultaneously with them to demodulator they are fed/conducted variation of frequency F_p. At the output of demodulator the oscillations of the difference frequencies

$$\Delta F_{вых} = F_p - F_в = F_p - [F_p - (F_{мин} \div F_{макс})] = F_{мин} \div F_{макс}. \quad (6.2.2)$$

are isolated.

These oscillations, after filter of low frequencies, are supplied to differential system ДС, and then proceed to correspondent p'. Transmission from point/item B to point/item A occurs analogously. Let us note that for the undistorted transmission it is necessary that the frequencies F_p which are supplied to the modulator and to the demodulator, established/installed in different points/items, would be identical. Therefore during development and operation of equipment for multiplexing special attention is paid to frequency stability of the generators, from which the oscillations proceed to user modulators and demodulators. For this at each station is utilized the general/common generator with high relative frequency stability (reaching to 10^{-8}), from which through different multipliers and frequency converters the oscillations are supplied to modulators and demodulators of different subscribers.

Fig. 6.2.2 shows connection to common line of communications of individual circuits of other subscribers. Each of such circuits contains ДС, ФНЧ, and also band-pass filters ПФ and individual

modulators БМ and demodulators Д, to which the oscillations of the corresponding frequencies are fed/conducted, in consequence of which the necessary bias/displacement of the signal of subscriber into the specific section of line spectrum is provided. The connection of the individual circuits of different subscribers is realized after the band-pass filters of channels on the side of transmission and before the band-pass filters of channels on the side of reception. The described principle of the construction of equipment for multiplexing is accepted during the creation of standard twelve-channel block with the line spectrum 60-108 kHz. This block forms the primary group of channels.

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One should note, that for transmission of broadcast signal or signal of sonic tracking of television program, with band 300-6000 Hz, in standard module association/unification of two telephone channels is provided that it is achieved by start of auxiliary equipment. For the transmission of broadcast signal or signal of the sonic tracking of the I class (with the band 50-10000 Hz) the association/unification of three telephone channels is possible. For the creation of the multichannel equipment, designed for the transmission of a large number of channels, they utilize several primary groups, whose line spectra with the aid of the group frequency converters Гр. П (analogous to individual balance modulators) and filters are transferred into the appropriate section of the group spectrum. In this way can be formed the secondary group, which consists of five primary groups of

channels.

Block diagram of equipment to 60 telephone channels is shown in Fig. 6.2.3. of this figure it follows that 60-channel equipment has line spectrum of frequencies from 312 kHz to 552 kHz, i.e., width of line spectrum ΔF=240 kHz. There is somewhat different construction of equipment on 60 telephone channels, with which the frequency spectrum from 12 kHz to 252 kHz is occupied. The difference between these versions of the construction of equipment consists in the fact that in the second version the supplementary conversion of the spectrum from region 312-552 kHz to region 12-252 kHz is realized. Let us note that in the systems, which use frequency modulation, is more expedient the application of that equipment for multiplexing, whose line spectrum of frequencies lies/rests at the region of lower frequencies, since in this case with the constant value of the deviation of frequency as a result of an increase of the index of modulation the power of noises in telephone channels at the output of line will prove to be less.

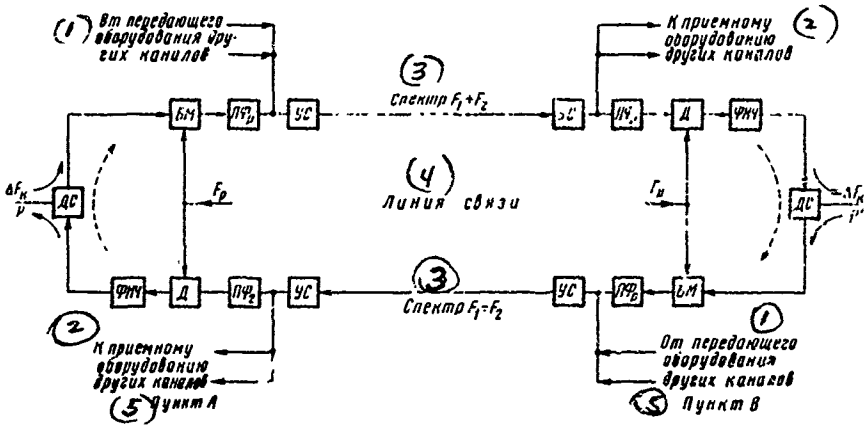

Fig. 6.2.2. Principle of construction of equipment with frequency division multiplex.

Key: (1). From the transmitting equipment of other channels. (2). To the receiving equipment of other channels. (3). Spectrum. (4). Line of communications. (5). Point/item.

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For forming tertiary group, by capacity/capacitance 300 of telephone channels, are utilized five secondary groups, constructed in accordance with block diagram of Fig. 6.2.3, in which then line spectra are transferred into different sections of group spectrum. As a result of this during the transmission of 300 telephone channels will be occupied the frequency spectrum from 60 kHz to 1300 kHz, i.e., with a span of ΔF=1300-60=1240 kHz.

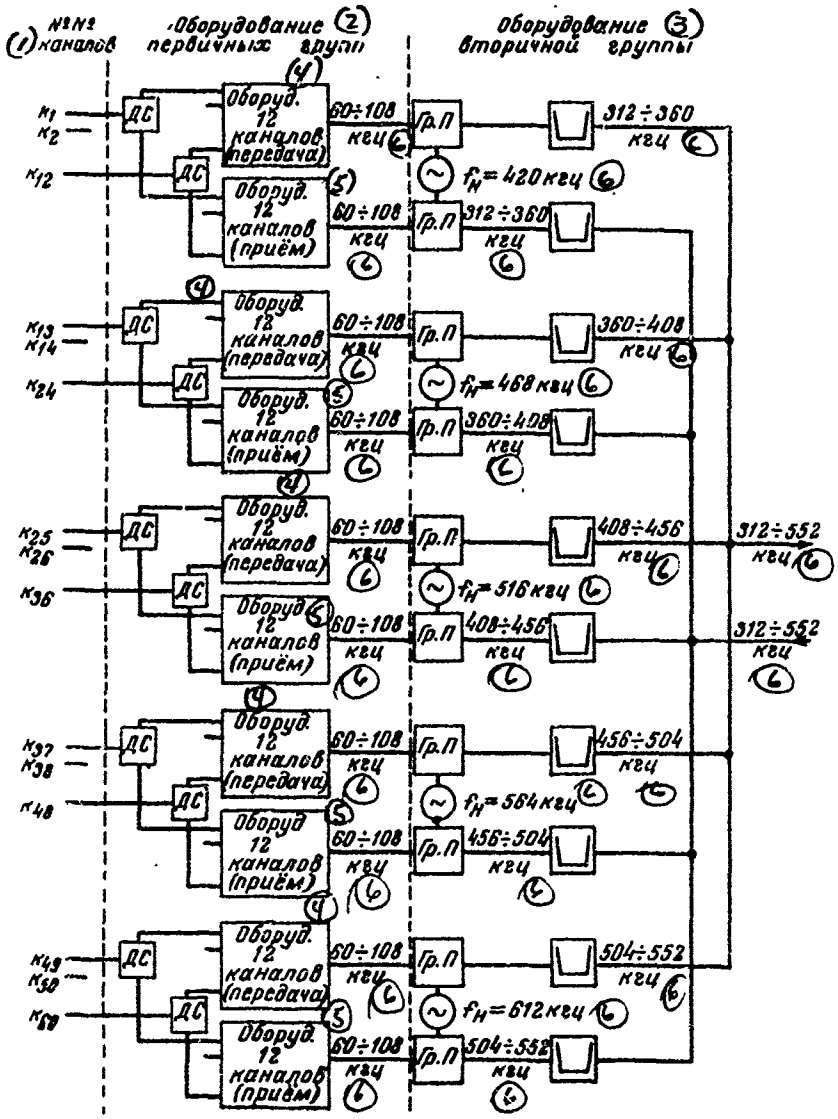

Fig. 6.2.3. Block diagram of equipment to 60 channels.

Key: (1). channels. (2). Equipment of primary groups. (3). Equipment of secondary group. (4). Equip. of 12 channels (transmission). (5). Equip. of 12 channels (reception). (6). kHz.

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For forming the tertiary group, by capacity/capacitance into 300

telephone channels, are utilized 25 primary groups (twelve-channel), from which are five secondary groups (six-ten-channel). Multichannel equipment with an an even larger number of channels is formed from several channel groups. In Table 6.2.1 are brought the data, characterizing group spectra of multichannel equipment.

One should note that construction of multichannel equipment from standard modules, multiple 12, 60 and 300 for channels, makes it possible to unify nodes of station equipment and to improve their interchangeability and redundancy.

Let us pass to examination of characteristics of multichannel telephone signal. From the construction of the equipment with the frequency division multiplex of channels, shown in Fig. 6.2.2, it follows that the linear (or group) spectrum is formed from the signals of different subscribers, which enter the input of equipment for multiplexing. In the multichannel systems telephone channels are utilized not only for the transmission of telephone communications/reports; along some telephone channels is conducted the transmission of telegraph and phototelegraphic signals, or broadcast signals. At the same time along the channels the impulse/transmission of line signals, measuring signals, etc sometimes is realized. Therefore the definition of the characteristics of group signal taking into account all possible combinations of charging is very complicated matter. One should still consider that the value of charging the communication system also is not constant, since a number of utilized

channels depends on the time of days, occurring events, etc. In order to exclude the inconstancy of charging, characteristic of multichannel signal accepted to determine for the hour of the greatest charging of system, moreover the fundamental form of charging is created by telephone signals. It is possible to consider that the telephone signals, which come from subscribers the input of the communication system, are independent from each other. These signal levels are the random functions of time, whose instantaneous values it depends on the sounds, pronounced by subscribers, the timbre of voice and temperament of subscribers, quality of telephone set and line of communications between the subscribers, equipment of multiplexing, etc. Together with these random and independent from each other phenomena during the definition of the characteristics of multichannel communication/report should be considered also that the fact that the simultaneity of the conduct of conversations by absolutely all subscribers is very scarcely probable.

Table 6.2.1.

(1) Число каналов	F_1 (2) кгц	F_2 (2) кгц	$\Delta F = F_2 - F_1$ (2) кгц
60	12	252	240
60	312	552	240
120	60	552	492
300	60	1300	1240
600	60	2540	2480
960	60	4028	3968
1800	312	8204	7892
2700	312	12388	12076

Key: (1). Number of channels. (2). kHz.

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For these reasons a quantity of the "active" channels n, i.e., those, along which is conducted the transmission of telephone signals during 99% of time, is not equal, but it is less than the total number of channels N of the considered/examined communication system. Fig. 6.2.4 gives the dependence of activity coefficient k=n/N on a total number of channels N, constructed for the hour of greatest charging [6.1]. From the examination of this graph/curve it follows that with increase in N the activity coefficient approaches value of 0.25.

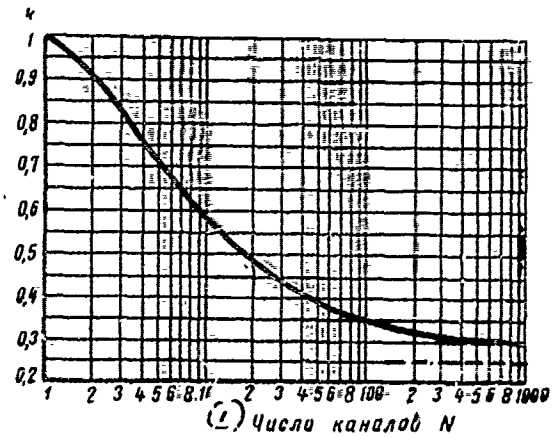

Fig. 6.2.4.

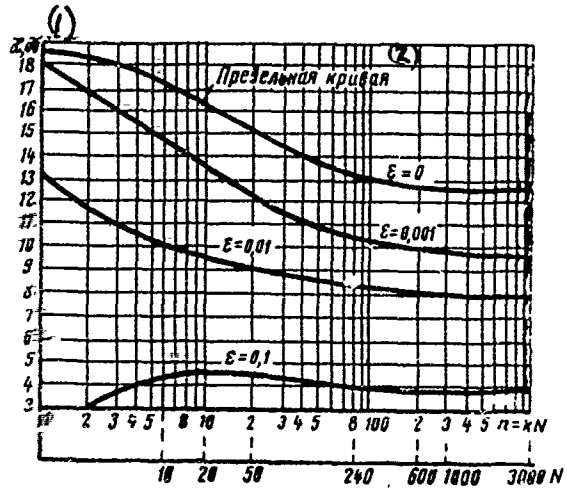

Fig. 6.2.5.

Fig. 6.2.4. Dependence of activity coefficient on number of channels N.

Key: (1). Numbers of channels N.

Fig. 6.2.5. Graph/curve of values of peak factor κ of multichannel signal.

Key: (1). κ, dB. (2). Limiting curve.

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Taking into account chance and mutual independence of levels of telephone signals, which enter channels of multichannel communication system, it is possible with confidence to consider that complete group signal is also random function of time, which has very abrupt changes of voltage/stress and power at separate moments.

Peak factor κ^2 of group signal, equal to relation of peak power (exceeded for time, equal to ϵ, from total time of observations), to

average/mean power of group signal, for different number of active channels is characterized by dependences, given in Fig. 6.2.5. The same figure gives the second scale, on which is indicated the total number of channels, calculated in accordance with *Fig.* 6.2.4. One should note that curves, shown in Fig. 6.2.5, are obtained for smooth conversation. The fact that with a small number of channels the difference between the peak value of power and the average/mean power proves to be considerably larger than with a larger number of channels, is characteristic for these curves.

Different researchers obtained not identical values for average/mean power $p_к$ of active channel, developed during sufficiently large time interval of observations at point with zero relative level (determination of zero relative level see in *A*ppendix 1). Thus, for the telephone exchange of the USA Holbrook and Dixon [6.1] found that this value composes 10.1 dBm, and Brokbank and Woss [6.2] in Great Britain determined its equal to 15.6 dBm. This disagreement in determination $p_к$ is explained by the national temperament of subscribers and by different quality of exchange lines, since with a good intelligibility of speech subscribers attempt to speak less loudly.

One should note that in spite of possibility of sharp difference in levels of separate channels, average/mean power of group signal $p_{гр}$ will be more constant. During determination $p_{гр}$ it is possible to consider that in connection with the random law of a change in the

levels, the signals in different channels prove to be not correlated. Therefore the average/mean power of all channels will be equal to the sum of average/mean power p_n of the active channels n and the certain supplementary power p', defined by the fact that along the line of communications together with the telephone signals both line signals and oscillations incompletely suppressed in the balanced modulators of the carrier frequencies of the individual channels are transmitted. Therefore the average/mean power of all channels

$$p_{cp} = p_к + 10\lg n + p' = p_к + p' + 10\lg kN = p_{к\,ср} + 10\lg N. \tag{6.2.3}$$

Here $p_{к\,ср} = p_к + p' + 10\lg k$ - conditional value of the absolute level of the average/mean power of the signal, transmitted on telephone channel to the hour of the greatest charging; k=n/N - activity coefficient of channels.

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MKKTT and MKKR for point with zero relative level with large number of telephone channels (N≥240) accepted value $p_{к\,ср} = -15$ dBm (-1.73 Np). Therefore the absolute level of the average/mean power of group signal at point with the zero relative level in the hour of the greatest charging can be defined as:

$$\left.\begin{aligned} p_{ср} &= -15 + 10\lg N,\ дБм\ (1)\ при\ (2)\ N \geqslant 240 \\ p_{ср} &= -1{,}73 + \frac{1}{2}\ln N,\ нп\ (3)\ при\ (2)\ N \geqslant 240 \end{aligned}\right\}. \tag{6.2.4}$$

Key: (1). dBm. (2). with. (3). Np.

One should note that in the case of secondary multiplexing of telephone channels (for transmission of telegraph and phototelegraphic signals) $p_{к\,ср}$ proves to be somewhat more. Taking this into account in

the Soviet systems of multichannel communication it is customary to assume $p_{\text{ш ср}} = -12$ dBm (-1.38 Np).

With number of channels N<240 MKKTT and MKKR recommend to use expressions:

$$\left.\begin{array}{l} p_{\text{ср}} = -1 + 4\lg N, \text{ дбм (1)} \quad \text{при (2)} \quad 12 \leqslant N \leqslant 240 \\ p_{\text{ср}} = 0{,}115 + 0{,}2\ln N, \text{ нп (3)} \quad \text{при} \quad 12 \leqslant N \leqslant 240 \end{array}\right\}. \quad (6.2.5)$$

Key: (1). dBm. (2). with. (3). Np.

Fig. 6.2.6 gives diagram, characterizing at point with zero relative level levels of multichannel communication/report with N>240 and sine wave with measuring level. Determination of measuring level is given in Appendix 4.

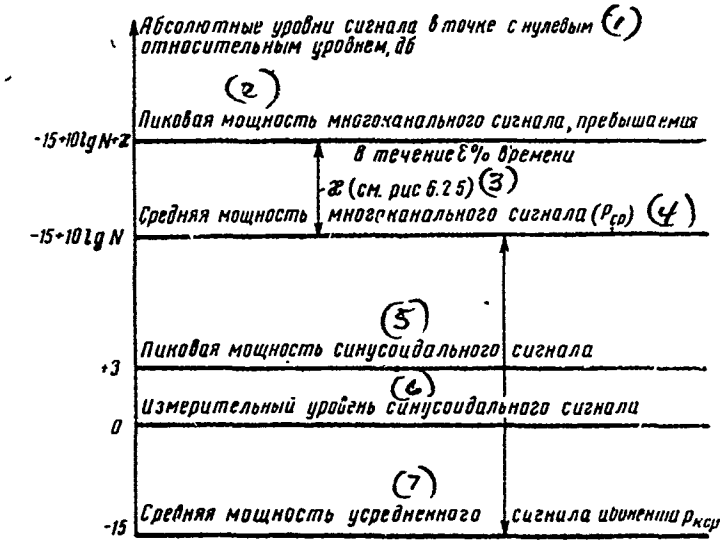

Fig. 6.2.6. Diagram of multichannel signal level with N>240. Key: (1). Absolute levels of signal in point with the zero relative level, dB. (2). Peak power of multichannel signal, exceeded by ε% in course of time. (3). ... (see Fig. 6.2.5). (4). Average/mean power of multichannel signal. (5). Peak power of sine wave. (6). Measuring level of sine wave. (7). Average/mean power of averaged signal of subscriber.

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From expression (6.2.4) and (П.4.8), which is given in Appendix 4, it follows that at point with zero relative level average/mean power of multichannel telephone signal:

$$\left.\begin{aligned} P_{ср} &= e^{2p_{ср}\cdot\,нп},\ мвт\ (1) \\ P_{ср} &= 10^{0{,}1p_{ср}\cdot\,дбм},\ мвт\ (1) \end{aligned}\right\}. \qquad (6.2.6)$$

Key: (1). mW.

Absolute levels of average/mean power of group signal at point

with zero relative level, designed according to expressions 6.2.4 and 6.2.5, and also amount of average/mean power of this signal at the same point, calculated according to expression 6.2.6 for different number of telephone channels, are given in Table 6.2.2.

Theoretical and experimental studies [6.3], [6.4] multichannel telephone signal showed that in this random process of U(t) probability distribution for instantaneous values of signal is subordinated to normal law. At the same time it was proved [6.5] that the multichannel signal is stationary random process.

In group signal is absent constant component; therefore average/mean value (designated in further formulas by feature collation) of multichannel signal $a=\overline{U(t)}=0$.

Exemplary/approximate form of energy spectrum of multichannel signal is shown in Fig. 6.2.7. Is here along the axis of ordinates spectral power density $G(\Omega)$ plotted.

From Fig. 6.2.7 it follows that spectral power density of real multichannel signal has different values on edges and in middle of band of each telephone channel and is turned into zero among separate channels.

Table 6.2.2.

(1) Число каналов	$P_{ср}$ (2) нп	(3) дБ	$P_{ср}$ мвт (4)
60	0,70	6,1	4,05
120	0,84	7,3	5,36
240	1,01	8,8	7,53
300	1,12	9,8	9,39
600	1,47	12,8	18,92
960	1,70	14,8	29,96
1200	1,81	15,8	37,33
1800	2,01	17,6	55,65
1920	2,05	17,8	60,37

Key: (1). Number of channels. (2). Np. (3). dB. (4). mW.

Fig. 6.2.7. Energy spectrum of multichannel signal.

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The approximately energy spectrum of real multichannel signal in the limits of cut-off frequencies can be replaced by the uniform energy spectrum (Fig. 6.2.8). In this case:

$$\left.\begin{aligned} G(\Omega) &= G_0 \text{ (1) при } F_1 \leqslant F \leqslant F_2 \\ G(\Omega) &= 0 \text{ (1) при } F_1 > F \text{ (2) } F > F_2 \end{aligned}\right\}. \qquad (6.2.7)$$

Key: (1). with. (2). or.

Here $F = \frac{\Omega}{2\pi}$; $F_2 = \frac{\Omega_2}{2\pi}$; $F_1 = \frac{\Omega_1}{2\pi}$.

Assuming that average/mean power of multichannel signal on resistor/resistance into 1 ohm is equal to $\overline{U^2(t)}$, taking into account

expression (П.5.17), given in Appendix 5 and (6.2.7), we obtain

$$G(\Omega) = G_0 = \frac{\overline{U^2(t)}}{(F_2 - F_1)}. \qquad (6.2.8)$$

In accordance with (П.5.15) it is possible to determine correlation function of multichannel signal

$$B(\tau) = \overline{U(t)U(t+\tau)} = \frac{1}{2\pi}\int_{\Omega_1}^{\Omega_2} G_0 \cos \Omega\tau \, d\Omega = \frac{G_0}{2\pi\tau}(\sin \Omega_2\tau - \sin \Omega_1\tau) \qquad (6.2.9)$$

In accordance with (П.5.14) with $\tau=0$ average/mean power on resistor/resistance of 1 ohm

$$B(0) = \overline{U^2(t)} = \frac{1}{2\pi}\int_{\Omega_1}^{\Omega_2} G_0 d\Omega = \frac{G_0}{2\pi}(\Omega_2 - \Omega_1). \qquad (6.2.10)$$

After substitution (6.2.8) in (6.2.9) we will obtain

$$B(\tau) = \frac{\overline{U^2(\tau)}}{(F_2 - F_1)2\pi\tau}(\sin \Omega_2\tau - \sin \Omega_1\tau) = \frac{\overline{U^2(\tau)}}{\Delta\Omega'\tau}\sin(\Delta\Omega'\tau)\cos\Omega_0\tau.$$

Here $\Omega_0 = \frac{\Omega_1 + \Omega_2}{2}$, $\Delta\Omega' = \frac{\Omega_2 - \Omega_1}{2}$.

Equalizing (6.2.10) and (П.5.14), let us find

$$\sigma^2 = \frac{G_0}{2\pi}(\Omega_2 - \Omega_1)$$

or

$$G_0 = \frac{\sigma^2}{F_2 - F_1} = \frac{\sigma^2}{\Delta F}. \qquad (6.2.11)$$

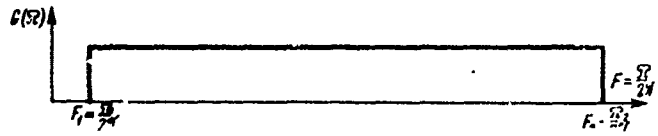

Fig. 6.2.8. Idealized energy spectrum.

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From explanations to (П.5.14) it follows that for multichannel telephone signal average/mean power with resistance/resistor of load $R_н$ is equal to $P_{ср}=\frac{B(0)}{R_н}$.

Taking into account (П.5.12) and (6.2.6) for the point with the zero relative level, we will obtain

$$B(0)=\sigma^2=R_н e^{2p_{ср}} \qquad (6.2.12)$$

Coefficient of correlation of multichannel signal in accordance with (П.5.7), (6.2.9) and (6.2.10)

$$R(\tau)=\frac{B(\tau)}{B(0)}=\frac{\sin\Omega_2\tau-\sin\Omega_1\tau}{\Omega_2\tau-\Omega_1\tau}. \qquad (6.2.13)$$

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6.4. Principles of the construction of equipment for multiplexing with KIM.

During use of KIM for transmission of telephone signals are possible several methods constructions of equipment for multiplexing. With one of them KIM is realized on the signals of subscribers, with other method - the signals from the subscribers first are supplied to equipment for frequency division multiplex, and KIM is realized along the group spectrum. Thus, in the second case is fulfilled conversion in KIM not of individual user telephone signals, but multichannel signal.

Let us examine first method of multiplexing telephone channels. In this case the line of communications is given to each subscriber to a certain interval of time τ_0, and then line is thrown to the following subscriber, etc.

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As already mentioned in Section 5.1, in accordance with Kotelnikov's theorem clock period T_t, the period of time, through which each subscriber must be connected to the line, is determined by width of band of the transmitted communication/report, i.e.

$$T_t \leqslant \frac{1}{2\Delta F_к}, \qquad (6.4.1)$$

the clock frequency

$$F_i = \frac{1}{T_i} \geqslant 2\Delta F_к. \qquad (6.4.2)$$

Fig. 6.4.1 shows time intervals, during which each of three subscribers, who transmit communications/reports $U_1(t)$, $U_2(t)$ and $U_3(t)$ must be connected to line.

Number of communications/reports N, which can be transmitted along one line of communications, is determined by relationship/ratio between T_i and τ_0 and necessary shielding gaps/intervals.

Fig. 6.4.2 shows simplified block diagram of equipment for multiplexing telephone channels. Pulse generator ГИ creates the sequence of the pulses/momenta with a clock frequency of F_i, which enter the distributor of pulses/momenta РИ. The latter is delay line with the removals/outlets, the distances between which correspond to the time intervals between the pulses/momenta of adjacent channels. Pulses/momenta from the appropriate removals/outlets of delay line РИ are fed/conducted to the channel by modulators М1, М2, М3 and so forth, where the numerals designate the number of channel. To the modulators, through the differential systems ДС, the signals from the subscribers are fed/conducted, as a result of which it is realized by АИМ. From the outputs of all modulators the pulses/momenta supply to the system of conversion of АИМ in КИМ, and then are supplied to the transmitter, where modulation of the oscillations of transmitter on the amplitude, the frequency or the phase is realized. Fig. 6.4.2 shows also introduction from РИ through the amplifier У of

synchronizing pulse. From the output of receiver all pulses/momenta enter the converter KIM in AIM. After this, the modulated pulses enter the selector of the timing pulses ССИ.

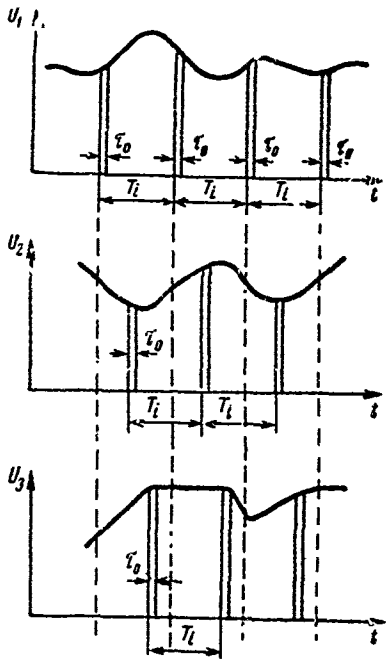

Fig. 6.4.1. Principle of time-division multiplex.

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Timing pulses from the output of ССИ are supplied to the distributor of pulses/momenta РИ and to the selectors of channels СК. The selectors of channels realize isolation/liberation of channel pulses, which then are supplied to the filter of the lower frequencies of ФНЧ With the output of ФНЧ of low-frequency oscillation they proceed to ДС and then to the subscribers.

With КИМ during that abstracted/removed to subscriber of interval of time τ_0 is realized impulse/transmission n pulses/momenta by duration τ, that form word or character. Let us note that for purposes of the decrease of pulse band it is desirable to make as

widely as possible; in connection with this between the pulses/momenta in a number of cases it is not done gaps/intervals.

Subsequently code combinations are supplied to modulator of transmitter, where modulation of oscillations of transmitter in amplitude, phase or frequency is realized by pulses/momenta, entering code combinations.

In receiver after detection are obtained initial code combinations, which subsequently with the aid of selector are supplied to individual demodulators of channels. as a result of this on characters it is conducted restoration/reduction of initial communication/report.

One should emphasize that with KIM all pulses/momenta, entering code combinations, not only are identical with respect to amplitude and duration, but they have also fixed/recorded position. The presence or the absence of pulse/momentum in code combination is the result of modulation during the binary coding only.

Fixed/recorded position of pulses/momenta in code combinations makes it possible to carry out pulse regeneration. The sense of regeneration is of the replacement of the "noised" pulses/momenta new, not having noises, i.e., in the regeneration instead of the pulses/momenta with the noises superimposed on them to the same fixed/recorded position in a precise conformity with code combination the new pulses/momenta, which do not have noises, are introduced.

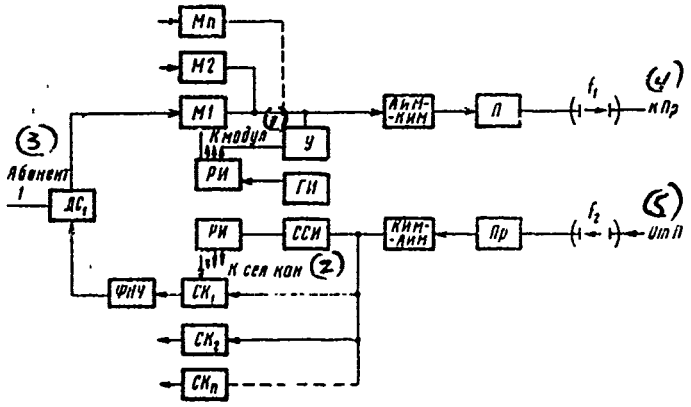

Fig. 6.4.2. Principle of construction of equipment with KIM.
Key: (1). To modul. (2). To channel sel. (3). Subscriber. (4). to Пр. (5). From П.

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Pulse regeneration makes it possible to considerably increase the interference shielding of the entire communication system, since the possibility of the accumulation of noises with pulse advancing on the sections of connection/communication from one station to another eliminates.

However, equipment for multiplexing with KIM and equipment for pulse regeneration proves to be fairly complicated. Therefore in the communication systems through ISZ the introduction to pulse regeneration on board the satellite thus far can be examined only over the long term.

Let us conduct evaluation/estimate of bandwidth, necessary in the

case of KIM. Taking into account that width of band of telephone signal is equal to 3100 Hz, in accordance with (6.4.2) we obtain, that the clock repetition frequency of code combinations in each channel can be $F_1=8\times10^3$ Hz (2.35 times of more than the higher frequency of the transmitted signal). If we consider that word can consist of n pulses/momenta, then in the case of N telephone channels the necessary bandwidth will comprise

$$\Delta F_1 = 1{,}05\times8\times10^3 nN = 8{,}4\times10^3 nN, \text{ кГц (1)} \qquad (6.4.3)$$

Key: (1). kHz.

coefficient of 1.05, which entered this expression, determines certain increase in the frequency band due to the transmission of timing pulses. Let us note that expression (6.4.3) confirms the statement, made during the analysis of formula (5.4.8).

Let us pass to examination of second method of transmission of telephone channels in the case of KIM. In this case during the use of standard equipment for the frequency division multiplex of telephone channels the transition/junction from the cable or radio relay lines to the communication systems with KIM can be simplified. The frequency band, which will be necessary, can be determined as follows. Considering that for the transmission of group signal it is necessary to have n of pulses/momenta in word, and increasing by 5% bandwidth due to the transmission of timing pulses, let us find

$$\Delta F_2 = F_{1rp}\,1{,}05n. \qquad (6.4.4)$$

Here F_{1rp} - clock frequency during transmission of group spectrum. Since $F_{1rp}=\frac{1}{T_{1rp}}=2\Delta F_1$ taking into account (5.4.1) and

(6.2.1), we will obtain

$$\Delta F_3 \geqslant 1{,}05 \cdot 2\Delta Fn = 1{,}05 \cdot 2 \cdot 4{,}2 \cdot 10^3 Nn = 8{,}8 \cdot 10^3 Nn, \text{ (1) гц.} \quad (6.4.5)$$

Key: (1). Hz.

Comparing (6.4.5) and (6.4.3), it is possible to see that the second method of the transmission of telephone channels with KIM requires somewhat larger frequency band, which is determined by the presence in equipment of the frequency division multiplex of protective belts, frequencies between telephone channels and by the groups of these channels.

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Chapter 7.

Transmission of telephone signals along the line of communications.

7.1. Electrical characteristics of channels.

During establishment of norms for communication channels, organized in different communication systems, MKKR proceeds from completely specific hypothetical standard circuits, which serve as management/manual during design and creation of real communication systems.

For communication system with the help of active satellites during transmission of multichannel telephone signals or signals of black and white television as hypothetical standard circuit for intercontinental systems is accepted line of communications Earth-satellite-Earth, shown in Fig. 7.1.1 [7.1; 7.2]. In the recommendation of MKKR is noted that the hypothetical circuit includes one pair of modulators and demodulators.

This hypothetical circuit corresponds to majority of intercontinental airlines of communications, organized through active ISZ, since with distances of satellite of more than 20000 km in this case transmission of signals up to distances, measured along surface of Earth, is ensured, not less than 7500 km. If is required the organization of the line of communications up to the distances, which

cannot overlap with the help of one satellite, will have to connect two, and in certain cases three hypothetical lines, shown in Fig. 7.1.1.

In Fig. 6.2.2 it is shown that for transition/junction from twin-lead circuit to four-wire and vice versa at input and exit of telephone channel is connected differential system, which must create high attenuation between circuit of reception and transmission.

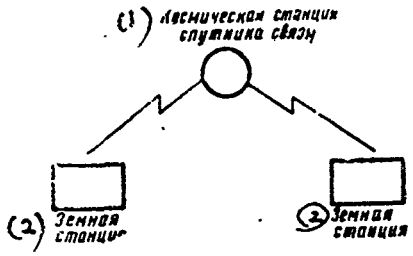

Fig. 7.1.1. Hypothetical standard circuit.

Key: (1). Space station of communication satellite. (2). Terrestrial station.

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Since the absolute balance of differential system in the circuit is impossible to provide, part of the energy passes from the circuit of reception into the circuit of transmission. As a result of this, as shown in Fig. 6.2.2, the loop of feedback appears. For preventing the self-excitation it is necessary that the level in the circuit of transmission would be higher than in the circuit of reception, i.e., in other words so that the product of all amplifications and weakenings in the communication channel from the input to the output would be less than one. Let us note that the product of all amplifications and weakenings, introduced into the communication channel, is called overall line attenuation.

According to recommendations of MKKTT [International Telegraph and Telephone Consultative Committee] value of overall line attenuation in telephone channel is equal to +0.8±0.2 Np, i.e., in channel there is always weakening signal level from 0.6 to 1 Np.

Since telephone channels are frequently used not only for the transmission of telephone communications/reports, but also for the transmission of other information, this universality of telephone channel leads to the complication of requirements. In particular, the permissible change in the signal frequency during the transmission along telephone channel must not exceed 2 Hz [7.4]. This condition leads to the fact that in the communication systems through ISZ must be accepted the measures, eliminating the change in the frequency, caused by the Doppler effect.

According to recommendations of MKKTT telephone channel must have passband from 300 to 3400 Hz, and frequency characteristic, which determines change of overall line attenuation in this frequency band, must lie/rest within limits of boundaries of stencil, shown in Fig. 7.1.2 [7.3].

Requirements for phase response of telephone channel are set to nonuniformity of group time of emission of the signal, which is defined as derivative of phase response in frequency. According to the recommendation of MKKTT the group propagation time from its minimum value can differ at the frequency of 300 Hz not more than on 10 μs, and at the frequency of 3400 Hz not more than on 50 μs.

Value of noises at output of telephone channel of communication system through ISZ is determined at point with zero relative level and is measured by psophometer, whose frequency characteristic considers physiological special feature of human ear/eye, which has different sensitivity at different frequencies of sonic spectrum.

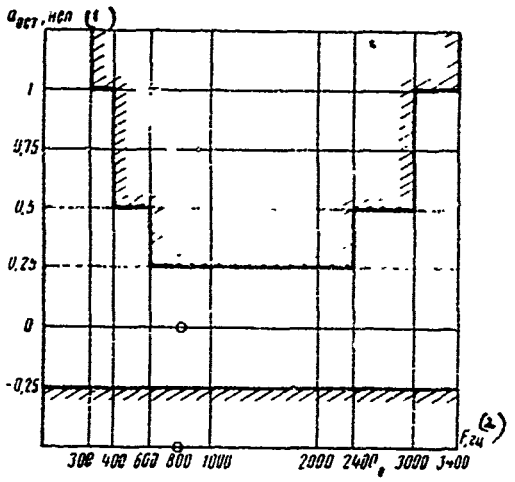

Fig. 7.1.2. Norms to frequency characteristic of telephone channel.
Key: (1). Np. (2). Hz.

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The ratio of the psophometric power of noises in the band of telephone channel to the unweighted power, measured in the same frequency band, is determined by the square of psophometric coefficient $\kappa_ш$ which is equal to 0.56 (-2.5 db).

According to recommendation of MKKR [7.1, 7.2] at output of telephone channel of communication system through ISZ at point with zero relative level psophometrically weighed power of noise in any telephone channel of hypothetical circuit must not exceed following values:

averaged in any hour ... 10000 pW

average-minute power of noise, which can be exceeded not more than 20% time of any month ... 10000 pW

average-minute power of noise, which can be exceeded not more

than 0.3% time of any month ... 50000 pW

average/mean in 5 ms unweighted power of noise, which can be exceeded not more than 0.03% time in any month ... 10^4 pW.

In this recommendation of MKKR are not considered noises of equipment for multiplexing, or peaks of noises and clicks, created by equipment of feed and commutation and by switchings from one satellite to another. Let us note that if at point with the zero power level of the psophometrically weighed noises is equal to $P_ш = 10000$ pW (10^{-5} mW), then the ratio of the power of test signal to the power of noise comprises

$$10\lg\frac{P_c}{P_ш} = 10\lg\frac{1}{10^{-5}} = 50\ дб. \qquad (7.1.1)$$

Since psophometric coefficient $\kappa_п = 0{,}75$, which is 20 lg 0.75=-2.5 db, at the same point ratio of power of test signal to power of unweighted noise will comprise

$$50 - 2{,}5 = 47{,}5\ дб. \qquad (7.1.2)$$

Noise at output of telephone channel can be subdivided into two groups: thermal (fluctuation) noises and transient noises, caused by distortions in different elements of communication system. The overall value of all noises must not exceed the values, determined by the given norms. Consequently, during the design of the communication system, on the basis of the possibilities and the technical-economic advisability of the construction of the diverse variants, developer must justify and select such values of the separate groups of the

noises, with which the overall value of noises will not exceed the norms of MKKR.

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7.2. Transmission of one sideband (OBP).

Principle of transmission of multichannel signal with frequency division multiplex is clarified by block diagram, given in Fig. 7.2.1. Here the telephone signals of subscribers with the frequency spectrum within limits $F_{мин} \div F_{макс}$ enter the equipment for multiplexing AU, at output of which are obtained the oscillations with the group spectrum F_1-F_2. After amplification in the group amplifier GU1 these oscillations fall on the converter of the transmitter Pr1, with the help of which the transposition (bias/displacement) of the group spectrum into the frequency region $f+(F_1-F_2)$ and $f-(F_1 \div F_2)$, occurs, moreover $f \gg F_1$; $f \gg F_2$. Through the filter F1 is passed one of these transposed group bands, cut-off frequencies of which let us designate through f_1-f_2, and width of band $\Delta f = f_2 = f_1 = F_2 - f_1$. It is obvious that during the transposition of the spectrum in accordance with the given inequalities will without fail be fulfilled the following condition, which determines the relative bandwidth:

$$\beta_F \gg \beta_f, \qquad (7.2.1)$$

here

$$\beta_F = \frac{F_2}{F_1}, \quad \beta_f = \frac{f_2}{f_1}. \qquad (7.2.2)$$

After filter F1 of oscillation with frequencies of f_1-f_2 they

fall on amplifiers of transmitter УП and on feeder proceed to antenna A'. Then oscillations are received by antenna of repeater, are fed/conducted to the receiver, are amplified, are displaced in the frequency in the converter, which forms part of auxiliary step/stage BC, and proceed to transmitter. Let us note that the bias/displacement of oscillations in the frequency in the repeater is necessary to avoid self-excitation, which can appear with the reception of the oscillations of its own transmitter.

Oscillations, which passed repeater, are accepted terrestrial antenna A", and after passage through receiver $П_3$ they proceed to converter of receiver $Пр_3$, in which is realized inverse transformation from oscillations with cut-off frequencies of f_1-f_2 to initial group spectrum with cut-off frequencies of F_1-F_2.

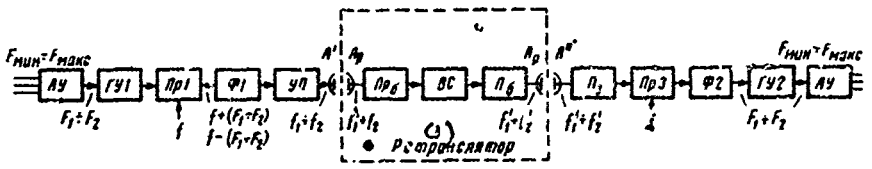

Fig. 7.2.1. Block diagram of transmission with OBP.

Key: (1). Retransmitter.

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Oscillations with these frequencies after the passage through the filter Ф2 and the amplifier ГУ2 enter the receiving part of the equipment for multiplexing АУ, at output of which they are converted into the telephone signals with spectrum $F_{мин} \div F_{макс}$, supplied to the subscribers.

From examination of this block diagram it follows that in transit through receivers oscillations will undergo effect of thermal noises and will be distorted due to imperfection of characteristics of equipment.

Thermal noises. Let us determine the power of thermal noises at the output of telephone channel at point with the zero relative level. In this case for simplification in the recording index "VYKh" let us drop/omit.

In accordance with expression (5.5.6) upon consideration (5.2.3) and (5.2.1) let us find

$$\frac{P_{ш}}{P_{c}} = \frac{kT_{эб}\Delta F B_{э1}}{P_{пэ}} + \frac{kT_{ээ}\Delta F B_{э2}}{P_{пб}}. \qquad (7.2.3)$$

With OBP average/mean power of multichannel signal at point with zero relative level can be determined according to expression (П.4.9) of appendix 4, i.e.

$$P_{c} = P_{ср} = 10^{0,1 p_{ср}}, \ мвт. \qquad (7.2.4)$$

Here level $p_{ср}$ is determined according to (6.2.4) or (6.2.5).

With substitution (7.2.4) in (7.2.3) let us find power of unweighted noises in band of multichannel signal $\Delta F = F_2 - F_1$ at point with zero relative level

$$P_{ш} = k\Delta F\left[\frac{T_{эб}B_{э1}}{P_{пэ}} + \frac{T_{ээ}B_{э2}}{P_{пб}}\right]P_{ср}, \ mW. \qquad (7.2.5)$$

Passing to calculation of weighed noises in one telephone channel, should be replaced ΔF by $\Delta F_{к}$ and introduced psophometric coefficient $\kappa_{п}^{2}$

$$P_{ш} = k\Delta F_{к}\left[\frac{T_{эб}B_{э1}}{P_{пэ}} + \frac{T_{ээ}B_{э2}}{P_{пб}}\right]P_{ср}\kappa_{п}^{2}, \ mW. \qquad (7.2.6)$$

From this expression it follows that value of noise will not depend on $F_{к}$, in all channels noise will be identical. With satisfaction of conditions (5.5.7) and (5.5.8) the calculation of the psophometric power of noises for one channel at point with the zero relative level in accordance with (5.5.9) can be conducted according to the simplified formula

$$P_{ш} \cong k\Delta F_{к}\kappa_{п}^{2}\frac{T_{ээ}B_{э2}}{P_{пб}}P_{ср}, \ mW, \qquad (7.2.7)$$

or, since $k=1{,}38\cdot10^{-23}\frac{дж}{град}\Delta F_{к}=3{,}1\cdot10^{3}$ Hz; $\varkappa_{п}^{2}=0{,}56$,

$$P_{ш}=2{,}4\cdot10^{-11}\frac{T_{вз}B_{т2}}{P_{пб}}P_{ср},\ \text{pW}. \qquad (7.2.8)$$

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Let us recall that in all given formulas value T is taken in degrees Kelvin, power of transmitters - in W, and $P_{ср}$ - in mW.

In the case, when value of noises at output of channel for section of connection/communication in question is assigned, and it is necessary to determine power of ground-based or onboard transmitter, on the basis of latter/last formula it is possible to record

$$P_{п}=2{,}4\cdot10^{-11}\frac{T_{т}B_{т}}{P_{ш\,зад}}P_{ср},\ \text{W}. \qquad (7.2.9$$

Here $P_{ш\,зад}$ - in pW, $P_{ср}$ - in mW, $T_{т}$ - in degrees, °K. Values $B_{т}$ and $T_{т}$ correspond to the section of connection/communication, for which is assigned the value of noises. The obtained from expression (7.2.9) power determines the average/mean power coefficient of transmitter. Its peak value can be calculated upon consideration of the peak factor of the multichannel signal $\varkappa$, which is located according to the graphs, given in Fig. 6.2.5. It is obvious that

$$\left.\begin{aligned}P_{п\,пик}&=P_{п}\varkappa'\\ \varkappa'&=10^{0{,}1\varkappa}\end{aligned}\right\}. \qquad (7.2.10)$$

Transient noises, caused by nonlinearity of amplifiers. Multichannel telephone signal in transit through the amplifiers or

other elements with the nonlinear amplitude characteristics is distorted.

We will consider that to input of quadrupole with nonlinear amplitude characteristic multichannel telephone signal $U_1(t)$ is given.

Assuming that amplitude characteristic of nonlinear quadrupole can be represented in the form of polynomial, we will find the expression for output signal $U_{вых}(t)$ in the form

$$U_{вых}(t) = a_1 U_1(t) + a_2 U_1^2(t) + a_3 U_1^3(t) + \ldots a_n U_1^n(t). \quad (7.2.11)$$

Here all coefficients $a_1, a_2, \ldots, a_n$ do not depend on time. The first term of this expression is signal at the output of quadrupole in the absence of nonlinear distortions. Introducing the designation

$$a_1 U_1(t) = U(t), \quad (7.2.12)$$

instead of (7.2.11) we will obtain

$$U_{вых}(t) = U(t) + a_2 U^2(t) + a_3 U^3(t) + \ldots \quad (7.2.13)$$

From (7.2.12) it follows that signal at output of quadrupole can be represented in the form of sum, comprised of undistorted part of U(t) of output signal and components, into which enter coefficients $a_2 = \frac{a_2}{a_1^2}$, $a_3 = \frac{a_3}{a_1^3}$; $\ldots a_n = \frac{a_n}{a_1^n}$ of polynomial.

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We will designate these components, called the products of nonlinear distortions,

$$\varepsilon(t) = a_2U^2(t) + a_3U^3(t) + \ldots \qquad (7.2.14)$$

Thus, output signal can be represented in the form

$$U_{вых}(t) = U(t) + \varepsilon(t). \qquad (7.2.15)$$

During analysis of distortions of cutput signal it is necessary to consider questions about spectral composition of functions $U(t)$ and $\varepsilon(t)$ and about relationship of their average/mean power in frequency band, occupied by transmitted signal. If the products of nonlinear distortions fall into the frequency spectrum, which occupies U(t), this will lead to a change in the value of the undistorted part of the output signal U(t), i.e., to the nonlinear distortions. For the evaluation of the distortions of group signal with its passage through the quadrupole with the nonlinear amplitude characteristic it is necessary to know:

- the spectral density of the undistorted part of the output signal;

-the spectral density of the products of the nonlinear distortions $\varepsilon(t)$ in the frequency band, which the undistorted part of the output signal U(t) occupies;

- relation of the spectral densities of functions ep(t) and U(t) in the individual sections of the frequency band, by the occupied undistorted part of signal U(t).

Knowing enumerated values, it is possible to solve question very important for practice about requirements for degree of linearity of

amplitude characteristic of amplifiers (quadrupoles), with execution of which distortions of output signal will not exceed those permitted.

Let us consider question about spectral density of function U(t) and $U_{вых}(t)$. From expressions (7.2.11) and (7.2.12) it follows that these functions are connected with signal $U_1(t)$, supplied to the input of quadrupole, with the coefficients of the polynomial, which do not depend on time. In the sections. 6.2 were noted, one hundred multichannel telephone signal are random stationary process, consequently, and functions $U_{вых}(t)$, $U(t)$ and $\varepsilon(t)$, connected with the input signal $U_1(t)$ with the coefficients of the polynomial, will also be random stationary processes. Therefore for the determination at the output of the quadrupole of the spectral density of the enumerated functions can be used the relationships/ratios between the correlation function and the spectral density, determined by Fourier transform (П.4.15) and (П.4.16). In this case it is implied that the corresponding characteristics of input signal are known and are determined by expressions (6.2.8)-(6.2.13).

Taking into account aforesaid, let us find fundamental characteristics for undistorted part of output signal during transmission of multichannel telephone signal.

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Spectral density $G(\Omega)$. In accordance with (6.2.7) and (7.2.12), we have

$$\left.\begin{aligned} G(\Omega) &= \alpha_1 G_0 \text{ when } F_1 \leqslant F \leqslant F_2 \\ G(\Omega) &= 0 \quad \text{with } F < F_1,\ F > F_2 \end{aligned}\right\}. \qquad (7.2.16)$$

Correlation function B(τ). According to (6.2.9) for region $F_1 \leq F \leq F_2$ we have

$$B(\tau) = \frac{1}{2\pi}\int_{\Omega_1}^{\Omega_2} \alpha_1^2 G_0 \cos \Omega\tau \, d\Omega = \frac{\alpha_1^2 G_0}{2\pi\tau}(\sin \Omega_2\tau - \sin \Omega_1\tau). \quad (7.2.17)$$

Average/mean power on resistor/resistance of 1 ohms. For this let us compute correlation function with $\tau=0$

$$B(0) = \frac{1}{2\pi}\int_{\Omega_1}^{\Omega_2} \alpha_1^2 G_0 d\Omega = \frac{\alpha_1^2 G_0}{2\pi}(\Omega_2 - \Omega_1). \qquad (7.2.18)$$

Let us now move on to determination of spectral function for $\epsilon(t)$. For this let us first find correlation function, and then, using Fourier transform, let us determine $G_\epsilon(\Omega)$.

Correlation function $B_\epsilon(\tau)$ in accordance with (П.5.9) and (П.5.6) can be recorded in the form

$$B_\epsilon(\tau) = \frac{1}{2\pi\sigma^2\sqrt{1-R^2(\tau)}}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \varepsilon(t)\,\varepsilon(t+\tau)\, e^{-\frac{U^2(t)+U^2(t+\tau)-2R(\tau)U(t)U(t+\tau)}{2(1-R^2(\tau))\sigma^2}} \times$$
$$\times dU(t)\, dU(t+\tau),$$

substituting (7.2.14), being limited to polynomial not higher than fifth degree, let us find

$$B_\epsilon(\tau) = \frac{1}{2\pi\sigma^2\sqrt{1-R_\tau^2}}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty}(a_2U^2+a_3U^3+\ldots a_5U^5)(a_2U_\tau^2+\ldots a_5U_\tau^5)\times$$
$$\times e^{-\frac{U^2+U_\tau^2-2R_\tau UU_\tau}{2(1-R_\tau^2)\sigma^2}} dU dU_\tau.$$

Here for the brevity of recording it is accepted

$$R_\tau = R(\tau); \qquad U = U(t); \qquad U_\tau = U(t+\tau).$$

As a result of integration, which can be carried out by different procedures (see for example [7.5] and by [7.6]), let us find

$$B_\varepsilon(\tau) = C_0^2 + R(\tau)C_1^2 + 2R^2(\tau)C_2^2 + 6R^3(\tau)C_3^2 + 24R^4(\tau)a_4^2 + 120R^5(\tau)a_5^2.$$

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Here:

$$\begin{aligned} C_0 &= a_2 + 3a_4\sigma^2, \\ C_1 &= 3a_3\sigma^2 + 15a_5\sigma^4, \\ C_2 &= a_2 + 6a_4\sigma^2, \\ C_3 &= a_3 + 10a_5\sigma^2. \end{aligned} \qquad (7.2.19)$$

Examining expression $B_\varepsilon(\tau)$, it is possible to do following conclusions:

- first term does not contain correlation function and, consequently, it does not depend on time. Therefore first term determines the constant component of function $\varepsilon(t)$, does not lead to the distortions in the transmitted signal and therefore subsequently it can be disregarded;

- second term contains correlation function to the first degree and, consequently, is determined that part of the products of nonlinear distortions, which will be coherent with the transmitted signal. Therefore this component/term will not lead to the nonlinear distortions, but it will only change the value of different components of signal. These distortions will be similar to those, which are obtained with the nonuniform amplitude-frequency characteristic of

amplifier or quadrupole. In the case, when percentage distortions are small, these distortions it is possible not to consider.

The terms, beginning from the third, will lead to nonlinear signal distortions. Thus, that part of the correlation function, which determines nonlinear signal distortions, it is described by the expression

$$B_{\Sigma}(\tau) = 2R^2(\tau)C_2^2 + 6R^3(\tau)C_3^2 + 24R^4(\tau)a_4^2 + 120R^5(\tau)a_5^2.$$

For determining spectral density of products of nonlinear distortions latter/last expression should be substituted into formula (П.5.15); in this case, since coefficients C and a do not depend on time, they can be carried out as integral sign. As a result we will obtain:

$$G_{\Sigma}(\Omega) = G_2(\Omega) + G_3(\Omega) + G_4(\Omega) + G_5(\Omega), \qquad (7.2.20)$$

$$\left.\begin{aligned} G_2(\Omega) &= 8C_2^2\int_0^{\infty} R^2(\tau)\cos\Omega\tau\,d\tau \\ G_3(\Omega) &= 24C_3^2\int_0^{\infty} R^3(\tau)\cos\Omega\tau\,d\tau \\ G_4(\Omega) &= 96a_4^2\int_0^{\infty} R^4(\tau)\cos\Omega\tau\,d\tau \\ G_5(\Omega) &= 480a_5\int_0^{\infty} R^5(\tau)\cos\Omega\tau\,d\tau \end{aligned}\right\} \qquad (7.2.21)$$

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For computing integrals it is possible to use theorem, known from theory of Fourier transforms [7.7]: if two functions $f_1(t)$ and $f_2(t)$ have spectral densities $\psi_1(\omega)$ and $\psi_2(\omega)$, then for the product of the functions, i.e., for function $f(t) = f_1(t)f_2(t)$, spectral density $\psi(\omega)$ can be found from expression

$$\psi(\omega) = \int_{-\infty}^{\infty} \psi_1(\omega_1)\,\psi_2(\omega-\omega_1)\,d\omega_1.$$

Using this theorem for computation $G_2(\Omega)$, let us find

$$G_2(\Omega) = 8C_2^2\int_0^{\infty} R^2(\tau)\cos\Omega\tau\,d\tau = 8C_2^2\int_0^{\infty} R(\tau)\,R(\tau)\cos\Omega\tau\,d\tau =$$
$$= 2C_2^2\int_{-\infty}^{\infty} G(\Omega_1)\,G(\Omega-\Omega_1)\,d\Omega_1. \qquad (7.2.22)$$

In the latter/last expression a change of integration limits indicates the transition/junction from one frequency region $F_1<F<F_2$, which was determined by conditions (6.2.7), to two regions with the conjugated/combined frequencies, shown in Fig. 7.2.2. Integral in expression (7.2.22) depends on the product of two functions, which determine spectral density, similar to that represented in Fig. 7.2.2, moreover one of these functions is shifted relative to another value Ω. The value of this shift/shear will determine the domains of existence of integral and integration limits. It is obvious that the integral, determined (7.2.22), will exist only in such cases, when $G(\Omega_1)$ and $G(\Omega-\Omega_1)$ is simultaneously different from zero. Fig. 7.2.3 shows functions $G(\Omega_1)$ and $G(\Omega-\Omega_1)$ with different value of shift/shear $\Omega=2\pi F$. Shading determines the domain of existence of integral in expression (7.2.22), and the boundaries of these regions - integration limits. Therefore instead of (7.2.22) it is possible to record:

$$G_2(\Omega) = 2C_2^2G_0^2\left\{\int_{2\pi(F-F_1)}^{-2\pi F_1} d\Omega_1 + \int_{2\pi(F+F_1)}^{2\pi F_2} d\Omega_1\right\} = 2C_2^2G_0^2\,4\pi(\Delta F - F)$$
(1) при $0 \leqslant F \leqslant F_2 - F_1$,

$$G_2(\Omega) = 2C_2^2G_0^2\int_{2\pi F_1}^{2\pi(F-F_1)} d\Omega_1 = 2C_2^2G_0^2\,2\pi(F-2F_1)$$
при $2F_1 \leqslant F \leqslant F_1 + F_2$,

$$G_2(\Omega) = 2C_2^2G_0^2\int_{2\pi(F-F_1)}^{2\pi F_2} d\Omega_1 = 2C_2^2G_0^2\,2\pi(2F_2-F)$$
при $F_1 + F_2 \leqslant F \leqslant 2F_2$.

Key: (1). with.

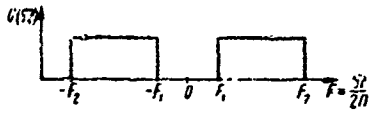

Fig. 7.2.2. Domains of existence G(Ω).

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By taking into account (6.2.11) and by passing from the anguiar frequencies Ω=2πF to the linear, we will obtain:

$$G_2(F) = 2C_2^2 \frac{\sigma^4}{\Delta F}\left(1 - \frac{F}{\Delta F}\right) \overset{(1)}{\text{при}}\ 0 \leqslant F \leqslant F_2 - F_1,$$

$$G_2(F) = 2C_2^2 \frac{\sigma^4}{\Delta F}\frac{F - 2F_1}{2\Delta F} \quad \overset{(2)}{\text{при}}\ 2F_1 \leqslant F \leqslant F_1 + F_2,$$

$$G_2(F) = 2C_2^2 \frac{\sigma^4}{\Delta F}\frac{2F_2 - F}{2\Delta F} \quad \overset{(3)}{\text{при}}\ F_1 + F_2 \leqslant F \leqslant 2F_2.$$

Key: (1). with.

It is here marked $\Delta F = F_2 - F_1$ - band of multichannel signal. From the analysis of resulting expressions it follows that into the band of multichannel telephone signal ΔF only the components, determined by the first two expressions, can fall. Moreover in the second expression it is necessary, in order to $F \leq F_2$. Taking into account the aforesaid, subsequently the third expression can be rejected/thrown.

Introducing dimensionless frequency

$$\xi = \frac{F - F_1}{F_2 - F_1} = \frac{F - F_1}{\Delta F} \qquad (7.2.23)$$

and designating

$$\beta = \frac{F_2}{F_1}, \qquad (7.2.24)$$

after conversion of expressions for $G_2(F)$ we will obtain

$$G_2(F) = 2C_2^2 \frac{\sigma^4}{\Delta F} y_2(\xi).$$

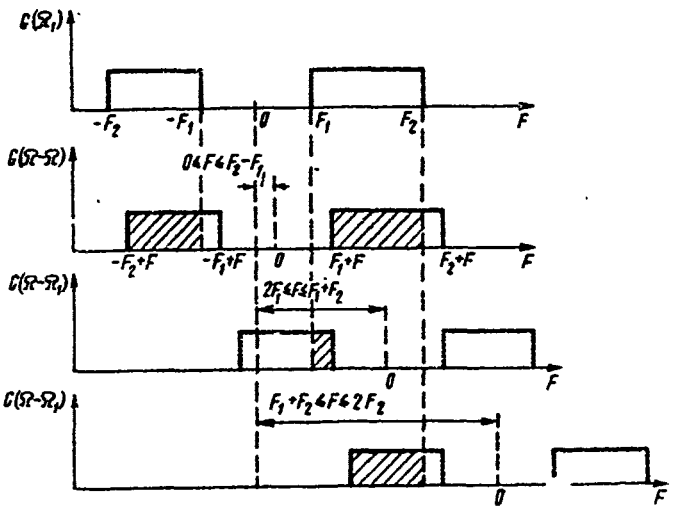

Fig. 7.2.3. Domains of existence of integral.

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Here

$$\left.\begin{aligned} y_2(\xi) &= \frac{\beta-2}{\beta-1}-\xi \quad \text{при (1)} \ 0\leqslant\xi\leqslant\frac{\beta-2}{\beta-1} \\ y_2(\xi) &= \frac{1}{2}\left(\xi-\frac{1}{\beta-1}\right) \quad \text{при (2)} \ \frac{1}{\beta-1}\leqslant\xi<1 \end{aligned}\right\}. \qquad (7.2.25)$$

Key: (1). with.

From expression (7.2.25) it follows that spectral distribution of products of nonlinear distortions, determined by quadratic member of polynomial, depends on value $y_2(\xi)$.

Fig. 7.2.4 shows dependence $y_2(\xi)$. Let us note that in accordance with (7.2.23) with a change of the frequency within the limits of band of multichannel telephone signal, value ξ vary within the range of zero to one. Thus, value $\xi=0$ corresponds to telephone

channel, the band of frequencies of which is located near the lower (smaller) cut-off frequency F_1 of the group spectrum, and to value $\xi=1$ corresponds telephone channel, the band of frequencies of which is located near the highest (greatest) cut-off frequency F_2 of the group spectrum.

Converting, analogous to those, which were carried out for determining of $G_2(\Omega)$, it is possible to find all values, determined (7.2.21):

$$G_3(F) = 6C_3^2 \frac{\sigma^6}{\Delta F} y_3(\xi);$$

$$G_4(F) = 24a_4^2 \frac{\sigma^8}{\Delta F} y_4(\xi);$$

$$G_5(F) = 120a_5^2 \frac{\sigma^{10}}{\Delta F} y_5(\xi).$$

All these values, which determine spectral density of products of nonlinear distortions, formed respectively by members of polynomial of third, fourth and fifth degree, contain factors $y_n(\xi)$. Dependences $y_3(\xi)$, $y_4(\xi)$ and $y_5(\xi)$ are represented respectively in Fig. 7.2.5, 7.2.6, 7.2.7.

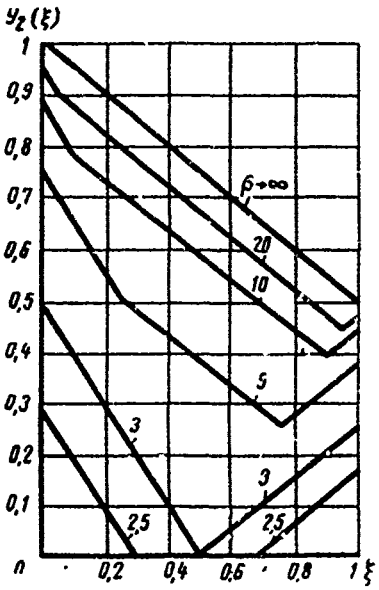

Fig. 7.2.4. Dependence $y_2=f(\xi)$.

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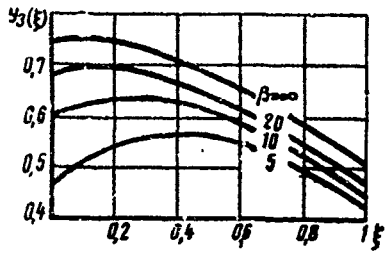
The total quantity of the spectral density of the products of nonlinear distortions can be found according to (7.2.20), if we take into account (7.2.19), (7.2.25) and the last expressions:

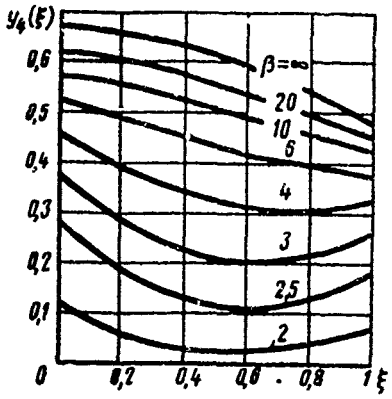
$$G_z(\Omega) = 2(a_2 + 6a_4\sigma^2)^2 \frac{\sigma^4}{\Delta F} y_2(\xi) + 6(a_3 + 10a_5\sigma^2)^2 \frac{\sigma^6}{\Delta F} y_3(\xi) + \\ + 24a_4^2 \frac{\sigma^8}{\Delta F} y_4(\xi) + 120a_5^2 \frac{\sigma^{10}}{\Delta F} y_5(\xi). \qquad (7.2.26)$$

We use this expression for calculation of transient noises, which are obtained in telephone channels due to products of nonlinear distortions. Taking into account (6.2.10), it is possible to record that in the band of telephone channel $\Delta F_к = F_{2к} - F_{1к}$ loaded to resistor/resistance of R, the average/mean power of interferences $P'_н$, nonlinear distortions caused by products, is determined as follows:

$$P'_{\text{ш}} = \frac{\overline{U^2(t)}_{\Delta F_{\text{к}}}}{R} = \frac{1}{R} \int_{F_{1\text{к}}}^{F_{2\text{к}}} G_{\text{т}}(\xi)\, dF. \qquad (7.2.27)$$

It is obvious that in this expression change of frequency F for value $G_{\text{т}}(\xi)$ corresponds to change ξ. From the examination Fig. 7.2.4-7.2.7 follow that with the relatively small changes ξ, which correspond to a change of the frequency within the limits of the band of any telephone channel, values $y_n(\xi)$ are changed insignificantly. Therefore in the frequency range, equal to the band of telephone channel $\Delta F_{\text{к}}$, values $y_n(\xi)$, and consequently, and $G_{\text{т}}(\xi)$ can be considered constant/invariable and equal to those values, which are determined by medium frequency $F_{\text{к}}$ of telephone channel in question.





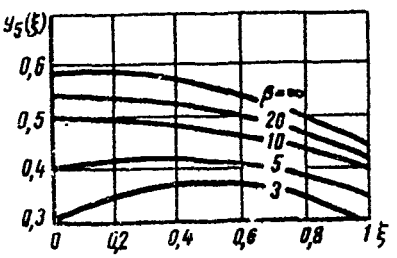


Fig. 7.2.5. Fig. 7.2.6. Fig. 7.2.7.

Fig. 7.2.5. Dependence $y_3=f(\xi)$.

Fig. 7.2.6. Dependence $y_4=f(\xi)$.

Fig. 7.2.7. Dependence $y_5=f(\xi)$.

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On this foundation expression (7.2.27) can be simplified

$$P'_{\text{п}} = \frac{G_{\text{т}}(\xi_{\text{к}})}{R} \int_{F_{1\text{к}}}^{F_{2\text{к}}} dF = \frac{G_{\text{т}}(\xi_{\text{к}})}{R} \Delta F_{\text{к}}.$$

Taking into account that measurement of interferences is made by psophometer, let us introduce psophometric coefficient $\kappa_{\text{п}}^2$. Therefore

$$P_{\text{п}} = \kappa_{\text{п}}^2 P'_{\text{п}} = \frac{G_{\text{т}}(\xi_{\text{к}})\,\Delta F_{\text{к}}}{R} \kappa_{\text{п}}^2. \qquad (7.2.28)$$

Taking into account that according to (6.2.12) $\sigma^2 = k e^{2\rho_{\text{ср}}}$ after substitution in (7.2.28) expressions (7.2.2[illegible], let us find

$$P_{\text{п}} = \kappa_{\text{п}}^2 \frac{\Delta F_{\text{к}}}{\Delta F}\Big[2\left(a_2 + 6a_4 R e^{2\rho_{\text{ср}}}\right)^2 R e^{4\rho_{\text{ср}}} y_2(\xi) + 6\left(a_3 + 10a_5 R e^{2\rho_{\text{ср}}}\right)^2 \times$$
$$\times R^2 e^{6\rho_{\text{ср}}} y_3(\xi) + 24a_4^2 R^3 y_4(\xi) e^{8\rho_{\text{ср}}} + 120a_5^2 R^4 e^{10\rho_{\text{ср}}} y_5(\xi)\Big]. \qquad (7.2.29)$$

Let us consider coefficients, entering first two terms of latter/last expression. taking into account the values of coefficients of a_2 and a_4, determined by formulas (П.6.22), let us find

$$a_2 + 6a_4\sigma^2 = C_2 = \frac{2}{\sqrt{2R}}\left[e^{-(b_2+p_1)} - 4e^{-(b_4+p_1)} + \frac{24}{2R}e^{-(b_4+3p_1)}Re^{2p_{cp}}\right] =$$
$$= \frac{2}{\sqrt{2R}}e^{-(b_2+p_1)}\left[1 - 4e^{-(b_4-b_2)} + 12e^{-(b_4-b_2+2p_1-2p_{cp})}\right].$$

In the case, when characteristics of amplifiers or quadrupoles are close to linear, it is possible to consider expression in bracket close to one. Therefore

$$a_2 + 6a_4\sigma^2 \cong a_2 = \frac{2}{\sqrt{2R}}e^{-(b_2+p_1)}. \qquad (7.2.30)$$

Analogous examination of coefficient $a_3+10a_5\sigma^2=C_3$ with small nonlinearity of characteristics leads to expression

$$a_3 + 10a_5\sigma^2 \cong a_3 = \frac{2}{R}e^{-(b_3+2p_1)} \qquad (7.2.31)$$

Substituting (7.2.30), (7.2.31) and value of a_4 and a_5 from (П.6.19) in expression (7.2.29), let us find

$$P_{ш} = \frac{\Delta F_K}{\Delta F}\kappa_n^2\left|4e^{-2(b_2+p_1)}e^{4p_{cp}}y_2(\xi) + 24e^{-2(b_3+2p_1)}e^{6p_{cp}}y_3(\xi) + \right.$$
$$\left. + 192e^{-2(b_4+3p_1)}e^{8p_{cp}}y_4(\xi) + 1920e^{-2(b_5+4p_1)}e^{10p_{cp}}y_5(\xi)\right|. \qquad (7.2.32)$$

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In the case, when calculation of power of products of nonlinear distortions is performed at point with relative level $p_{ш}$ and values of indices of nonlinearity are determined on zero level of fundamental tone, we will obtain following expression for determining transient

noises in band of one telephone channel

$$P_{но} = e^{+2p_к} \frac{\Delta F_к \kappa_п^2}{\Delta F} \left[4e^{-2b_2} e^{4p_{ср}} y_2(\xi) + 24e^{-2b_3} e^{6p_{ср}} y_3(\xi) + \right.$$
$$\left. + 192e^{-2b_4} e^{8p_{ср}} y_4(\xi) + 1920e^{-2b_5} e^{10p_{ср}} y_5(\xi) \right], \text{ мвт.}^{(1)} \quad (7.2.33)$$

Key: (1). mW.

Under analogous conditions, when percentage distortions are determined by supply of two sinusoidal oscillations on zero level of fundamental tone, in accordance with expressions (П.6.35) and (7.2.29), we will obtain

$$P_{но} = e^{2p_к} \frac{\Delta F_к \kappa_п^2}{\Delta F} \left[e^{-2b_1-1} e^{4p_{ср}} y_2(\xi) + \frac{8}{3} e^{-2b_2-1} e^{6p_{ср}} y_3(\xi) + \right.$$
$$\left. + \frac{16}{3} e^{-2b_2-2} e^{8p_{ср}} y_4(\xi) + \frac{96}{5} e^{-b_3-2} e^{10p_{ср}} y_5(\xi) \right], \text{ мвт.}^{(1)} \quad (7.2.34)$$

Key: (1). mW.

Expression (7.2.33) can be converted so that instead of the indices of nonlinearity b_n into it would enter the coefficients of nonlinearity K_n. Upon consideration of expression (П.6.15) we will obtain

$$P_{но} = 4e^{2p_к} e^{4p_{ср}} \frac{\Delta F_к \kappa_п^2}{\Delta F} \left[K_2^2 y_2(\xi) + 6e^{2p_{ср}} K_3^2 y_3(\xi) + \right.$$
$$\left. + 48e^{4p_{ср}} K_4^2 y_4(\xi) + 480e^{6p_{ср}} K_5^2 y_5(\xi) \right], \text{ мвт.}^{(1)} \quad (7.2.35)$$

Key: (1). mW.

One should note that according to expressions (7.2.33), (7.2.34) and (7.2.35) it is possible to calculate power of transient noises, which are obtained due to nonlinearity of characteristics of any of amplifiers, shown in Fig. 7.2.1, i.e., amplifiers ГУ and УП, and also amplifiers of repeater. However, during calculations should be

focused attention on the fact that the coefficient β, in accordance with which on the curves in Fig. 7.2.4-7.2.7 are determined values $y_{\text{к}}(\xi)$, will be different for the amplifiers ГУ and УП or ГУ and the amplifiers of repeater, which follows from expression (7.2.2).

From expressions (7.2.33), (7.2.34) and (7.2.35) it follows that power of products they are nonlinear distortions (transient noises) it grows with increase in number of channels, since in this case increases $p_{\text{ср}}$ (see Table 6.2.2). Transient noise especially sharply depends on the nonlinear products of higher orders.

For estimate of the magnitude, which are obtained during calculations according to expressions (7.2.33) and (7.2.35), let us give calculation of indices of nonlinearity and coefficients of nonlinear distortions of KNI with following initial data:

- number of channels N=60 and 600;
- power of products of nonlinear distortions (transient noises) at point with zero relative level $P_{\text{но}}=500$ pW$=5\cdot10^{-7}$ mW, $p_{\text{к}}=0$.

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All components/terms, entering brackets of expression (7.2.33), are taken as equal to each other. This means that is accepted the condition, according to which the powers of noises, created by the products of the nonlinear distortions of different order, are equal to each other and, consequently, are equal to (500/4)=125 pW;

- value $y_2(\xi)=y_3(\xi)=y_4(\xi)=y_5(\xi)=0.5=y(\xi)$.

Expression (7.2.33) for this special case can be recorded as

$$P_{\text{но}} = \frac{\Delta F_{\text{к}} \kappa_{\text{п}}^2}{\Delta F} y(\xi) 4M. \qquad (7.2.36)$$

Here coefficient

$$M = 4e^{-2b_2} e^{4p_{\text{ср}}} = 24e^{-2b_3} e^{6p_{\text{ср}}} = 192e^{-2b_4} e^{8p_{\text{ср}}} = 1920e^{-2b_5} e^{10p_{\text{ср}}}. \qquad (7.2.37)$$

From (7.2.37) it follows:

$$e^{-2b_2} = \frac{M}{4} e^{-4p_{\text{ср}}},$$

$$b_2 = 2p_{\text{ср}} - \frac{1}{2} \ln \frac{M}{4},$$

it is analogous:

$$b_3 = 3p_{\text{ср}} - \frac{1}{2} \ln \frac{M}{24},$$

$$b_4 = 4p_{\text{ср}} - \frac{1}{2} \ln \frac{M}{192},$$

$$b_5 = 5p_{\text{ср}} - \frac{1}{2} \ln \frac{M}{1920}.$$

If b_2-b_5 are found, then according to (П.6.15) it is easy to determine permissible coefficients of nonlinear distortions

$$K_n\% = \frac{100U_n}{U_1} = e^{-b_n} \cdot 100.$$

Coefficient M can be determined from (7.2.36):

$$M = 0{,}25P_{\text{н}} \frac{\Delta F}{\Delta F_{\text{к}} \kappa_{\text{п}}^2 y(\xi)}.$$

Substituting the value $P_{\text{н}} = 5 \cdot 10^{-7}$ mW; $\Delta f = 240$ kHz with N=60; $\Delta f = 2480$ kHz with N=600 (see Table 6.2.1); $\Delta F_{\text{к}} = 3{,}1$ кгц; $\kappa_{\text{п}}^2 = 0{,}56$; $y(\xi) = 0{,}5$, it is easy to determine all values. Values $p_{\text{ср}}$ are given in table 6.2.2 or are counted according to formulas (6.2.4) and (6.2.5). The results

of calculating the indices of nonlinearity b_n and percentage distortions K_n are given in table 7.2.1.

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From Table 7.2.1 it follows that requirements for linearity of amplitude characteristics of amplifiers, modulators and demodulators are very high: indices of nonlinearity for second and third harmonics during the supplying into channel of measuring level must be 7-9 Np (coefficient of nonlinear distortions - several hundredths of a percent), and for fourth and fifth harmonics - 10-15 Np in dependence on number of channels (coefficient of nonlinear distortions - thousandths and hundred-thousandths of a percent).

If measurement of nonlinearity of characteristics will be realized by level, in Z of times which exceeds measuring (on voltage/stress), then according to application/appendix 6 value of percentage distortions on n harmonic can be Z^{n-1} once that larger given in Table 7.2.1. The value of the indices of nonlinearity in this case in comparison with the tabular value can be reduced on Z^{n-1}.

For explanation of aforesaid let us find permissible value of percentage distortions in measurements with the help of sine voltage with level, equal to p_{cp}. Initial conditions are the same as during the calculation of Table 7.2.1, and a number of channels N=600.

From Table 7.2.1, and it follows that $p_{cp}=1,47$ Np and therefore in $e^{2\cdot1,47}=19$ times according to power exceeds measuring level. Thus, value $Z=\sqrt{19}=4.35$. So that in telephone channel at point with the zero relative level the total power of transient noises would not exceed 500 pW, group amplifier (modulator, demodulator) was necessary, in which in the measurements with sine voltage with level p_{cp} of value K_n in Z^{n-1} times differed from those given in Table 7.2.1. Therefore value $K_2 \le 0.055 \cdot 4.35 = 0.24\%$; $K_3 \le 0.006 \cdot 4.35^2 = 0.11\%$; $K_4 \le 3.6 \cdot 10^{-4} \cdot 4.35^3 = 0.03\%$; $K_5 \le 1.1 \cdot 10^{-5} \cdot 4.34^4 = 0.004\%$.

Table 7.2.1.

(1) Число каналов	P_{Hu} мат (2)	M	$P_{ср}$	(3) Показатель нелинейности, нп				(4) Коэффициент нелинейности			
				b_2	b_3	b_4	b_5	K_2	K_3	K_4	K_5
60	$5\cdot10^{-7}$	$34{,}5\cdot10^{-6}$	0,7	7,23	8,83	10,63	12,5	0,07%	0,014%	$2{,}5\cdot10^{-3}$ %	$4\cdot10^{-4}$ %
600	$5\cdot10^{-7}$	$34{,}5\cdot10^{-5}$	1,47	7,62	9,88	12,56	15,18	0,055%	0,006%	$3{,}6\cdot10^{-4}$ %	$1{,}1\cdot10^{-5}$ %

Key: (1). Number of channels. (2). mW. (3). Index of nonlinearity, Np. (4). Percentage distortion.

Hence follows that identical power of transient noises is obtained in ratio $(K_2/K_4) \approx 8$ and $(K_2/K_5) \approx 60$. Thus, the unjustified neglect in the calculations by values K_4 and K_5 (or b_4 and b_5) can lead to the inadmissibly high value of transient noises.

Let us note that given data are not comprehensive and carry illustrative character. Obvious also that the accepted during calculations total value of transient noises, equal to 500 pW, during the identical distribution of this value between the products of the nonlinearity of different orders (on 125 pW) it is a special case.

From expressions (7.2.33), (7.2.35) it follows that in general case of power of transient noises, created by products of nonlinearity of different orders, they can be not identical, but different. With other words, during the creation of group amplifiers developer can arbitrarily change the relationships/ratios between the values of percentage distortions or the indices of nonlinearity; however, these arbitrary changes must be implemented so that the total power of noises, determined by expressions (7.2.33)-(7.2.35), would not be above given one. For the illustration of the possible limits of a change in coefficients and indices of nonlinearity Fig. 7.2.8 gives the graphic dependence, which determines the relationships/ratios between these coefficients for the following case: the total power of transient noises at point with the zero absolute level is equal to 10, 100, 200 and 1000 pW (it is indicated as the parameter to curves),

number of channels N_1=600 (unbroken curves) and N_2=60 (dotted curves); $y_2(\xi)=y_3(\xi)=0.5$; the products of the nonlinearity of the fourth and fifth order they are absent (i.e. $b_4\to\infty$; $b_5\to\infty$; $K_4=0$, $K_5=0$).

From examination of graphs Fig. 7.2.8 and Table 7.2.1 it follows that during transmission of OBP all amplifiers must have amplitude characteristic with very high degree of linearity. Especially large difficulties this condition it will cause during the development of powerful steps/stages of transmitter, since in this case of their efficiency will be small.

Transient noises, which appear in frequency converters. Let us consider the block diagram of the frequency converter of oscillations/vibrations with the filter, shown in Fig. 7.2.9. Here to the frequency converter is supplied the oscillation/vibration with a frequency of shift/shear of f_0 and the multichannel telephone signal, which occupies frequency spectrum F'-F". Let us note that this spectrum can lie/rest both above, and lower than frequency kf_0, where k=1, 2, 3. If f_0>(F'-F"), then at the output of converter will appear the oscillations/vibrations of upper and lower sidebands or as they are sometimes called, the total k sidebands kf_0+(F'-F") and the inverse differential k sidebands κf_0—(F'-F").

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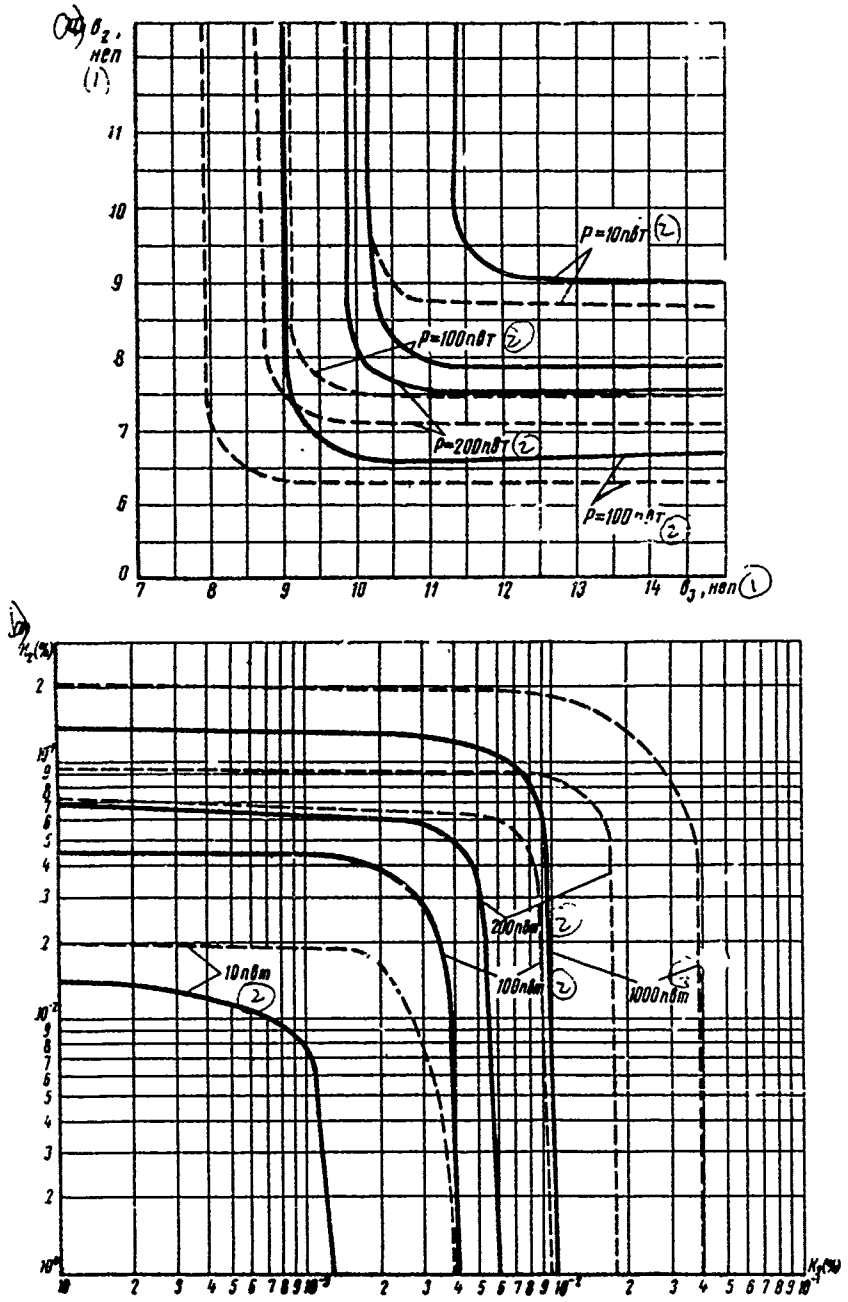

Fig. 7.2.8. Relationships between indices of nonlinearity (graph A) and percentage distortions (graph B) with different value of power of noises for 60 and 600 channels: - - - - N=60, —— N=600.

Key: (1). Np. (2). pW.

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The filter, which stands after converter, is tuned to one them them, and thus is selected the useful signal, the frequency spectrum of which lies/rests above the frequency spectrum of the input signal. If $f_0<(F'-F")$, then at the output c. converter they appear, furthermore, and the sidebands of form $(F'-F") - kf_0$, called straight/direct differential k^{th} sideband. Due to the nonlinearity of amplitude conversion diagram at its output, besides the enumerated sidebands, will arise other bands, determined by different combination frequencies. They all can be decomposed into three groups. The first group of combination frequencies will be the products of the nonlinearity of different order, formed by input signal. The second group of combination frequencies is determined by the products of the nonlinearity of different order, formed by oscillations/vibrations with the frequencies, to the multiple frequency of shift/shear. Finally, the third group of combination frequencies will contain the products, which are determined both by input signal and oscillations/vibrations with the frequencies, to the multiple frequency of shift/shear. In it will be contained the useful signal, which will pass through the filter, shown in Fig. 7.2.9. It is obvious that all products of conversion, which fall into the band of useful signal, will cause transient noises in telephone channels. The power of these noises at point with the zero relative level at the output of telephone channel can be determined employing the same procedure, by which was obtained expression (7.2.32). As a result of examination, with which conversion diagram is approximated by the

polynomial of fifth degree, analogous to expression (7.2.11), it is possible to obtain following expression [7.6]:

$$P_{\text{ш}} = \frac{\Delta F_{\kappa} \kappa_{\text{п}}^2}{\Delta F} \left\{ e^{-2b_0} \left[\left(\frac{A_{01}}{A_1} \right)^2 e^{2p_{\text{ср}}} + \left(\frac{A_{02}}{A_1^2} \right)^2 e^{4p_{\text{ср}}} e^{-2[\, b_{1-1}(p_1)+p_1]} y_2 (\xi_{02}) + \right.\right.$$
$$+ \frac{8}{3} \left(\frac{A_{03}}{A_1^3} \right)^2 e^{6p_{\text{ср}}} e^{-2[\, b_{2-1}(p_1)+2p_1]} y_3 (\xi_{03}) +$$
$$+ \frac{16}{3} \left(\frac{A_{04}}{A_1^4} \right)^2 e^{8p_{\text{ср}}} e^{-2[\, b_{2-2}(p_1)+3p_1]} y_4 (\xi_{04}) +$$
$$\left.\left. + \frac{96}{5} \left(\frac{A_{05}}{A_1^5} \right)^2 e^{10p_{\text{ср}}} e^{-2[\, b_{3-2}(p_1)+4p_1]} y_5 (\xi_{05}) + \ldots \right] + \right.$$
$$+ \sum_{m=1}^{4} e^{-b_m} \left[\left(\frac{A_{m1}}{A_1} \right)^2 e^{2p_{\text{ср}}} + \left(\frac{A_{m2}}{A_1^2} \right)^2 e^{4p_{\text{ср}}} e^{-2[\, b_{m,\,1-1}(p_1)+p_1]} y_2 \left(\xi_{m2}^{(\kappa)} \right) + \right.$$
$$+ \frac{8}{3} \left(\frac{A_{m3}}{A_1^3} \right)^2 e^{6p_{\text{ср}}} e^{-2[\, b_{m,\,2-1}(p)+2p_1]} y_3 \left(\xi_{p3}^{n} \right) +$$
$$\left.\left. + \frac{16}{3} \left(\frac{A_{m4}}{A_1^4} \right)^2 e^{8p_{\text{ср}}} e^{-2[\, b_{m,\,2-2}(p_1)+3p_1]} y_4 \left(\xi_{m4}^{(\kappa)} \right) + \ldots \right] \right\}. \quad (7.2.39)$$

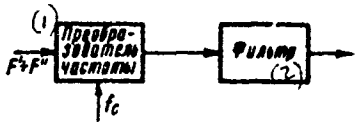

Fig. 7.2.9. Block diagram of frequency converter.

Key: (1). Frequency converter. (2). Filter.

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Here b_0 and b_m - indices of nonlinearity, determined at output of converter during the supplying to input of converter of two identical with respect to amplitude oscillations/vibrations with frequencies F_a and F_c from expressions

$$e^{-b_0} = \frac{U_{1,1}}{U_{1\pm1}}; \quad e^{-b_m} = \frac{U_{m\pm1}}{U_{1\pm1}},$$

where $U_{1,1}$ - amplitude of voltage component on output of converter, that relates to first group of combination frequencies and determined by product of first-order nonlinearity (i.e. voltage/stress U_{F_a} or U_{F_c}); $U_{m\pm1}$ - amplitude of voltage component on the output of converter, relating to the $m^{\pm th}$ sideband of the products of first-order nonlinearity (i.e. voltage/stress $U_{mf_0+F_c}$ or $U_{mf_0+F_a}$);

$U_{1\pm1}$ - amplitude of component of first sideband, determined by the products of first-order nonlinearity (i.e. voltage/stress $U_{f\pm F_a}$ or $U_{f\pm F_c}$);

$b_{r-s(p_1)}$ and $b_{m,r-s(p_1)}$ - indices of nonlinearity, determined from the expressions

$$e^{-b_{r-s(p_1)}} = \frac{U_{r-s}}{U_{1,1}}; \quad e^{-b_{m,r-s(p_1)}} = \frac{U_{m\pm(r-s)}}{U_{m\pm1}};$$

where U_{r-s} - amplitude of component of output potential of converter, that relates to the first group of the combination frequencies of the products of nonlinearity r+s of order;

$U_{m\pm(r-s)}$ - amplitude of voltage component, which relates to the m-th sideband of the products of nonlinearity r+s of order.

Index p_1 stresses that during determination of enumerated indices of nonlinearity of amplitude U_{i+1} at output of converter has absolute level, equal to p_1, no.

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A_1 - is determined according to expression (P.6.18):

$$A_{01} = A_0 + \frac{1}{2A_1^2} P_{cp} e^{-[b_{2-1(p_1)}+2p_1]} \times$$

$$\times \left\{1 - 5\left(1 - \frac{3}{20A_1^2} P_{cp} e^{-2p_1}\right) e^{-[b_{3-2(p_1)}-b_{2-1(p_1)}]}\right\},$$

$$A_0 = 1 - 3e^{-b_{2-1(p_1)}} + 5e^{-b_{3-2(p_1)}},$$

$$A_{02} = 1 - 4e^{-[b_{2-2(p_1)}+b_{1-1(p_1)}]}\left(1 - \frac{1}{4A_1^2} P_{cp} e^{-2p_1}\right),$$

$$A_{03} = 1 - 5e^{-[b_{3-2(p_1)}-b_{2-1(p_1)}]}\left(1 - \frac{3}{10A_1^2} P_{cp} e^{-2p_1}\right),$$

$$A_{04} = A_{05} = 1,$$

$$A_{m1} = A_m + \frac{1}{2A_1^2} P_{cp} e^{-[b_{m,\,2-1(p_1)}+2p_1]} \times$$

$$\times \left\{1 - 5\left(1 - \frac{3}{20A_1^2} P_{cp} e^{-2p_1}\right) e^{-[b_{m,\,3-2(p_1)}-b_{m,\,2-1(p_1)}]}\right\},$$

$$A_m = 1 - 3e^{-b_{m,\,2-1(p_1)}} + 5e^{-b_{m,\,3-2(p_1)}},$$

$$A_{m2} = 1 - 4e^{-[b_{m,\,2-2(p_1)}-b_{m,\,1-1(p_1)}]}\left(1 - \frac{1}{4A_1^2} P_{cp} e^{-2p_1}\right).$$

$$A_{m3} = 1 - 5e^{-[b_{m,3-2(p_1)} - b_{m,2-1(p_1)}]}\left(1 - \frac{3}{10A_1^2} P_{cp} e^{-2p_1}\right),$$

$$A_{m4} = A_{m5} = 1; \qquad \xi_{0n} = \frac{f - F'}{F'' - F'}; \qquad \xi_{mn}^{(1)} = \frac{f - mf_0 - F'}{F'' - F'};$$

$$\xi_{mn}^{(2)} = \frac{mf_0 - f - F'}{F'' - F'}; \qquad \xi_{mn}^{(3)} = \frac{f + mf_0 - F'}{F'' - F'}.$$

Here f_0 - frequency of shift/shear (see Fig. 7.2.9), f - value of frequency in group channel. Index in the brackets in values ξ indicates, which of the sidebands is selected by the filter, which stands after converter. Index 1 relates to the case, when filter is tuned to the total sideband, index 2 - corresponds to use to inverse differential sideband, while index 3 - to straight/direct differential sideband.

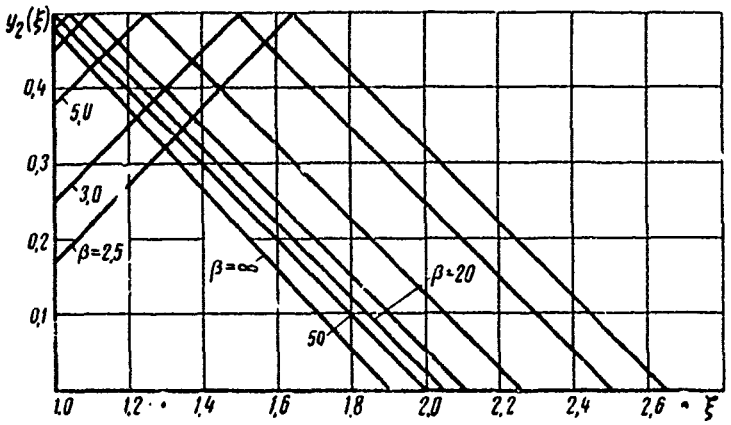

Fig. 7.2.10. Dependence $y_2(\xi>1)$.

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Functions $y_n(\xi_{0n})$ are determined according to graphs, given in Fig. 7.2.4-7.2.7. Function $y_n(\xi_{mn}^{(\kappa)})$ with $\xi\leq 1$ is determined according to the graphs Fig. 7.2.4-7.2.6, and with $\xi>1$ - according to the graphs Fig. 7.2.10-7.2.13. On these graphs value $\beta=(F''/F')$.

Depending on selection of value of frequency f in expressions for ξ, using formula (7.2.39), it is possible to determine power of transient noises, introduced during frequency conversion, for any channel of system.

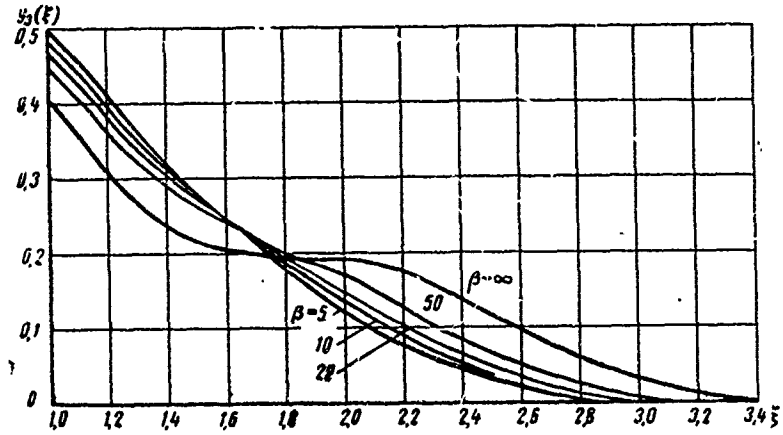

Fig. 7.2.11. Dependence $y_3(\xi>1)$.

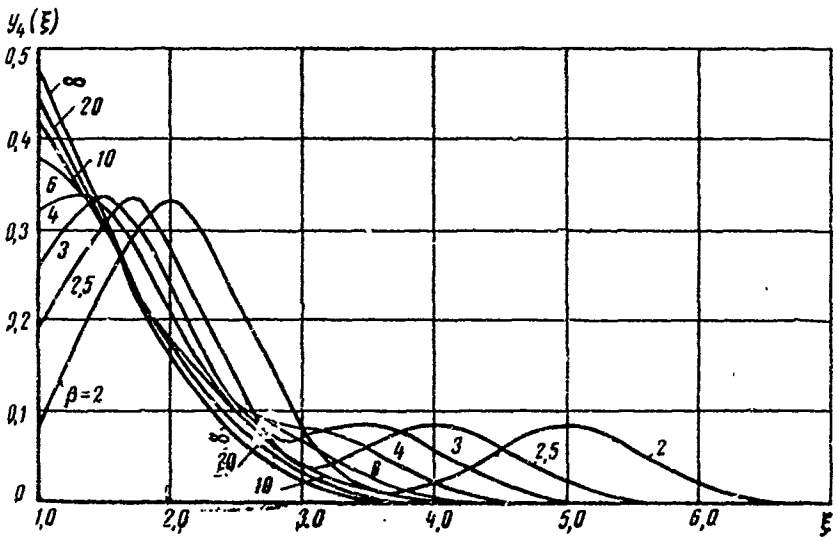

Fig. 7.2.12. Dependence $y_4(\xi>1)$.

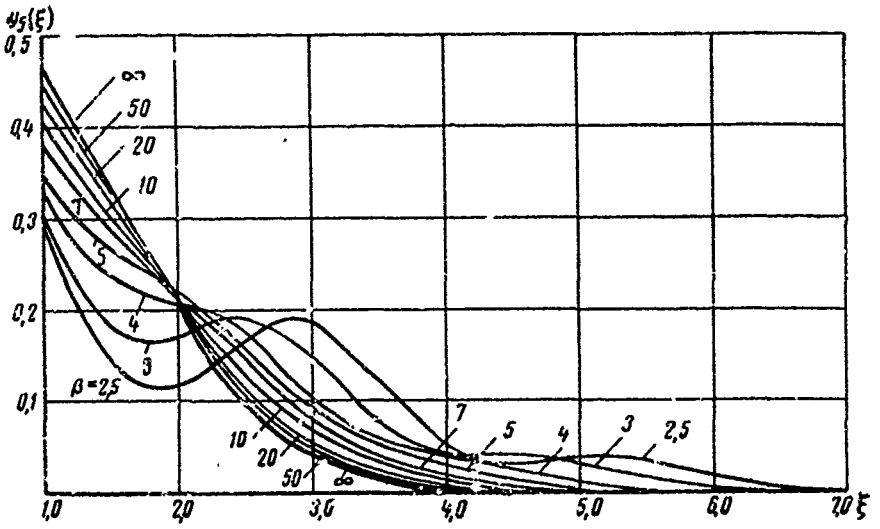

Fig. 7.2.13. Dependence $y_5(\xi>1)$.

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7.3. Transmission with the frequency modulation.

Transmission of multichannel telephone signal with frequency division multiplex in the case of ЧМ can be clarified on block diagram Fig. 7.3.1.

Telephone signals with frequency spectrum $F_{мин}$-$F_{макс}$, which come from N subscribers, after passage of equipment of multiplexing АУ and group amplifier ГУ proceed to frequency shift key of ЧМ transmitter П in the form of multichannel telephone signal with spectrum F_1-F_2. In the absence of modulating voltage the transmitter emits oscillations/vibrations with a carrier frequency of f_0; with modulation the instantaneous value of frequency will be changed.

Oscillations/vibrations, after passing the feeder of transmitter, will act into antenna A' and they will be emitted to the side of repeater. On the repeater the oscillations/vibrations with a carrier frequency of f_0 are accepted, are amplified, they pass through the amplitude limiter Orp, they are displaced in the frequency, for preventing the self-excitation of repeater, and they are again emitted by antenna.

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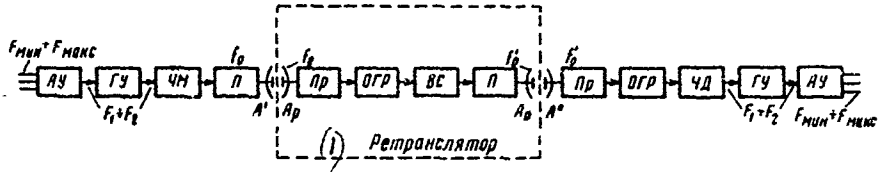

Fig. 7.3.1.

Fig. 7.3.1. Block diagram of transmission with ChM.

Key: (1). Repeater.

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Thus, frequency, emitted by repeater, is equal to f'_0. The frequency switch of the oscillations/vibrations, passing through the repeater, is realized in the auxiliary step/stage of repeater BC. Let us note that in the same step/stage it can be realized and an increase in the deviation of frequency, what can be useful for increasing the gain in the section satellite - Earth. The oscillations/vibrations, emitted by repeater, are received as the low-noise terrestrial receiver Пр. At the output of receiver stand the amplitude limiter of oscillations Огр and frequency discriminator ЧД. After ЧД are obtained the oscillations with the group spectrum, limited by frequencies F_1-F_2. These oscillations after the series/row of conversions in the receiving part of the equipment for multiplexing АУ are supplied to the subscribers.

Let us conduct estimate of the magnitude of noises, which appear in contemporary systems with ChM, on the basis of results, obtained during development of radio relay lines (RRL).

For hypothetical RRL with length of 2500 km average/mean in any hour psophometrically weighed power of noises at point with zero relative level in any telephone channel must not exceed 7500 pW. The relationships/ratios between the separate components of the

psophometrically weighed power of noises at point with the zero relative level at the output of telephone channel RRL in accordance with the considerations, given in [7.8], are approximately similar:

the thermal noises ... 3400 pW.

transient noises, caused by reflections in feeders ... 1800 pW.

transient noises, caused by nonlinearity of phase responses of circuit of shf/SVCh and PCh ... 2300 pW.

If we consider that hypothetical RRL, in the case of transmission of telephone conversations on 60 and more to channels, consists approximately of 54 stations and, consequently, signal in this case passes through 108 feeder lines, we obtain, which to one feeder falls not less (1800/108) ≈17 pW.

Let us assume that power of noises, created by nonlinearity of phase responses of each station RRL is equal to power of noises, which are obtained from nonlinearity of amplitude characteristics of group amplifiers and characteristics of frequency shift keys and detectors. Then, taking into account that accepted by MKKR hypothetical RRL contains 54 stations, to 9 ChM modulators and demodulators and 18 group amplifiers, let us find that for each of such elements, and also to one complete circuit of the high and intermediate frequency of each station falls the power of noises, equal to (2300/54+9+9+18)=23 pW.

Passing now to Fig. 7.3.1, it is possible to say that if technical indices of nodes of communication system through ISZ will be approximately the same as indices of analogous elements RRL, then power of transient noises, which appear in communication system through ISZ, as a result of reflections in feeders will be about 4×17=68 pW, and caused by nonlinearity characteristics of different elements - about 150 pW.

In real communication systems through ISZ corrected values of noises will be somewhat different. This is determined by the fact that the power of the noises, which appear with the reflections in the feeder, sharply is reduced with the decrease of its length. If we take into account that the length of the feeders of repeater is very small due to compactness of equipment, and the length of the feeder of ground-based receiver is done as small as possible for purposes of reduction in the thermal noises, then we will obtain that the value of the noises, which appear as a result of the reflections in the feeders, in essence will be determined only by the feeder, which connects terrestrial transmitter with the antenna. Consequently, the total quantity of noises due to the reflections in the feeders with the fulfillment of the transmitting feeder of terrestrial station with the same thoroughness as feeders RRL, will not exceed 20-30 pW and in further calculations can be disregarded.

Another position will be with noises, caused by nonlinearity of characteristics of repeater. Taking into account the need for maximum

reduction in the weight of the elements of equipment of repeater, compactness of the nodes and other requirements, cannot be considered that the nonlinearity of the characteristics of equipment of repeater will correspond to the characteristics of the analogous elements RRL station. Apparently, are more correct to present to the linearity of response of repeater less stringent requirements, than for the characteristics of equipment RRL.

Taking into account aforesaid, subsequently for communication systems through ISZ we will examine only thermal noises, noises, which appear as a result of nonlinearity of amplitude characteristics of group amplifiers, characteristics of ChM modulators and demodulators, and also noises, caused by nonlinearity of phase responses of elements of circuit, through which are passed by ChM oscillation.

Thermal noises. Keeping in mind (5.5.6) and (5.3.25), in the general case, when in the auxiliary step/stage BC of the repeater (see Fig. 7.3.1) a change in the parameters of modulation (for example, an increase in the deviation of frequency) is realized, we will obtain

$$\frac{P_{ш}}{P_c} = k\Delta F_к F_к^2 \left[\frac{T_{э6} B_{э1}}{P_{п3} \Delta f_{к1}^2} + \frac{T_{э3} B_{э2}}{P_{п6} \Delta f_{к2}^2} \right]. \qquad (7.3.1)$$

Let us recall that indices 1 and 2 relate respectively to section of connection/communication Earth - satellite and satellite - Earth, and $P_{п3}$ and $P_{п6}$ are determined power of transmitters (terrestrial and onboard) in W.

Power of psophometrically weighed noises at output of telephone channel at point with zero relative level will be

$$P_{ш} = k\Delta F_{к} F_{к}^2 \varkappa_{п}^2 \left[\frac{T_{вх6} B_{в1}}{P_{пв} \Delta f_{к1}^2} + \frac{T_{вх} B_{в2}}{P_{п6} \Delta f_{к2}^2} \right], \text{ mW.} \qquad (7.3.2)$$

With satisfaction of conditions (5.5.7) and (5.5.8) calculation of thermal noises it is possible to carry out only for section satellite - Earth. In this case we obtain the simplified formula

$$P_{ш} = k\Delta F_{к} F_{к}^2 \varkappa_{п}^2 \frac{T_{вх} B_{в2}}{P_{п6} \Delta f_{к2}^2}, \text{ mW.} \qquad (7.3.3)$$

Taking into account that $k=1.38 \cdot 10^{-23}$ J/deg; $\Delta F_{к}=3,1$ kHz; $\varkappa_{п}^2 =0,56$, we will obtain

$$P_{ш} = 2,4 \cdot 10^{-11} \frac{T_{вх} B_{в2}}{P_{п6} \Delta f_{к2}^2} F_{к}^2, \text{ pW.} \qquad (7.3.4)$$

In the case, when assigned magnitude of fluctuation noises at output of telephone channel for specific section of connection/communication and it is necessary to determine power of ground-based or onboard transmitter, on the basis of latter/last expression and given numerical values we will obtain

$$P_{п} = 2,4 \cdot 10^{-11} \frac{T_{з} B_{з}}{\Delta f_{к}^2 P_{ш\,зад}} F_{к}^2, \text{ W.} \qquad (7.3.5)$$

Here $P_{ш\,зад}$ - psophometrically weighed assigned power of noises in pW, $T_{з}$ - in degrees of °K. Values $T_{з}$ and $B_{з}$ must correspond to the section of connection/communication, for which is assigned the value

of noises.

Let us note that calculation according to expressions (7.3.1)-(7.3.5) is feasible only on signal level, which exceeds threshold value.

From expressions (7.3.2) and (7.3.5) it follows that power of noises depends substantially on value of effective value of deviation of frequency in telephone channel and on medium frequency of telephone channel $F_к$. Consequently, with the identical value of the deviation of frequency in all telephone channels the value of noises will increase in proportion to the square of the medium frequency of telephone channel, what will lead to a qualitative difference in the channels - upper channels will have smaller signal-to-noise ratio, than lower. At the same time during the selection of the identical values of the effective deviation of frequency $\Delta f_к$ for all telephone channels the threshold value of signal will prove to be different: with a gradual change in the signal level with reception (for example, as a result of a deterioration in the orientation of antenna) the signal will achieve threshold value first in the lower channels, and then - in the upper. The latter is explained by a difference in the indices of modulation in the lower and upper channels.

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For equalization of quality of telephone channels it is necessary on transmitter before frequency shift key to establish/install

predistorting filter, with the help of which deviation of frequency is changed is proportion to medium frequency of each telephone channel. On the receiving side after frequency discriminator for equalizing of the levels in different telephone channels is established/installed compensating filter, whose frequency characteristic is opposite (it is reverse/inverse) to the frequency characteristic of the predistorting filter.

During determination of frequency characteristic of predistorting filters they proceed of two conditions. One provides for the uniform distribution of the total power of noises in different telephone channels as far as possible. Another determines the invariability of charging the communication system with the switched on and switched-off predistorting filter, i.e., the invariability of value $p_{ср}$.

At present for communication systems through ISZ frequency characteristic of predistorting filter of MKKR is not yet determined, since spectral distribution of total power of noises, which depends on relationship of levels of received signal and threshold signal, is insufficiently known. Nevertheless MKKR examined the schematic of predistorting filter [7.2], whose attenuation $b_{0ф}$ at frequency F of the group spectrum is determined by the expression

$$b_{0\Phi} = 10\lg\left[1 + \frac{3{,}5}{1 + \frac{6{,}1}{\left(\frac{F_r}{F} - \frac{F}{F_r}\right)^2}}\right], \text{ dB.} \qquad (7.3.6)$$

Here $F_r=1{,}25\ F_2$ - resonance frequency of ducts/contours of predistorting filter; F_2 - maximum frequency of group spectrum.

For such a filter the transfer coefficient, equal to 1 ($b_{0ф}=0$), is obtained at frequency $F+0.55F_2$. The frequency response of the predistorting filter, defined by the dependence of $b_{0ф}$ on the ratio F/F_2, is shown in Fig. 7.3.2.

To determine $P_ш$ and $P_п$ with installation of a predistorting filter in the circuit, into the denominator of expressions (7.3.2)-(7.3.5) should be introduced factor $W=10^{0{,}1b_{0ф}}$, which characterizes change of deviation of frequency due to predistortions in telephone channel with central frequency of $F_к$. Value $b_{0ф}$ is found from the graph of Fig. 7.3.2 or from an analogous one taken for the filter of predistortions.

Introduction into the group circuit of a predistorting filter with the response in Fig. 7.3.2 will improve the signal-to-noise ratio in the upper telephone channel to a value equal to 3 dB (W=2 times with respect to power).

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The schematic diagram of the predistorting filter with the response in Fig. 7.3.2 is shown in Fig. 7.3.3 and has the following parameters with input and output resistance R_0:

$$\sqrt{\frac{L_1}{C_1}} = 0{,}455R_0; \qquad \frac{1}{2\pi\sqrt{L_1C_1}} = F_r = 1{,}25F_2,$$

$$\sqrt{\frac{L_2}{C_2}} = \frac{R_0}{0{,}455}; \qquad \frac{1}{2\pi\sqrt{L_2C_2}} = F_r = 1{,}25F_2.$$

At frequency F_r must be observed conditions $R_1 << 0.01R_0$; $R_2 < 0.01R_0$. Fig. 7.3.4 shows the schematic diagram of compensating filter, which must be established/installed after ЧЩ of receiver, if before the frequency shift key the filter, made according to diagram 7.3.3, is established/installed. Parameters of the filter shown in Fig. 7.3.4, are as follows:

$$\sqrt{\frac{L_1}{C_1}} = 1{,}3R_0; \qquad \frac{1}{2\pi\sqrt{L_1C_1}} = F_r = 1{,}25F_2,$$

$$\sqrt{\frac{L_2}{C_2}} = \frac{R_0}{1{,}3}; \qquad \frac{1}{2\pi\sqrt{L_2C_2}} = F_r = 1{,}25F_2.$$

At frequency F_r satisfaction of condition $R_1 < 0.01R_0$ is necessary.

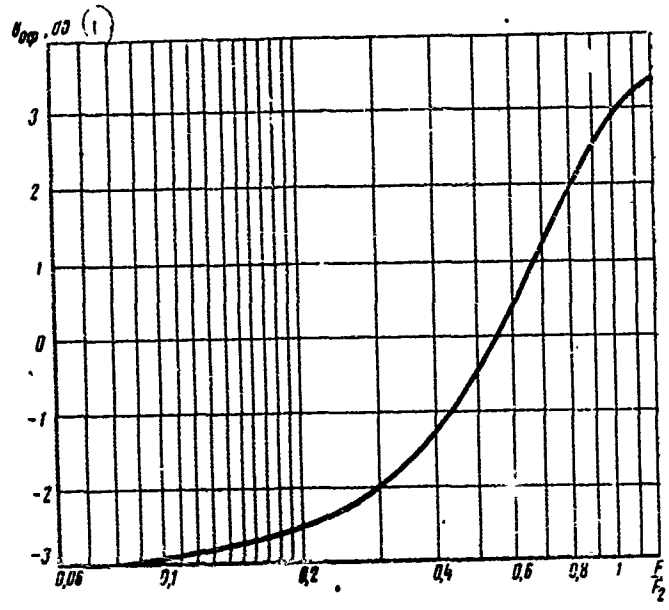

Fig. 7.3.2. Attenuation of predistorting filter.

Key: (1). dB.

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Optimum value of deviation of frequency in communication systems through ISZ at present is not yet determined. However, taking into account the special features of these systems in comparison with radio relay lines, it is possible to say that in the communication systems through ISZ it is expedient to choose the value of the deviation of frequency considerably larger than in RRL. This is connected with the fact that basic part of the power of noises at the output of telephone channel in the communication systems through ISZ is determined by thermal noises, whereas in RRL more than half of the power of all noises is determined by transient noises.

One should note that increase of deviation of frequency in communication systems through ISZ in combination with negative feedback in frequency (or with other methods, which make it possible to reduce threshold value of signal) they make it possible to considerably decrease power of onboard transmitter. However, the excessive value of deviation will cause a considerable increase in the total frequency band, which will hinder/hamper the creation of onboard and ground-based receiver.

For some specific conditions can be taken/removed characteristics, which determine optimum value of root-mean-square deviation of frequency, at which at output of line of communications will be obtained greatest value $q_{max} = \left(\frac{P_c}{P_ш}\right)_{max}$. Such characteristics, which make it possible to select the optimum value of the deviation of frequency, they are shown in Fig. 7.3.5 [7.10]. They were taken/removed during the tests of the communication system through ISZ "Telstar" during charging of system by noise signal with the band from 60 to 1052 kHz, equivalent to 240 telephone channels. Along the axis of abscissas it is plotted to the ratio (in dB) of the effective deviation of frequency $\Delta f_{ср}$ to 10 MHz. Measurements were conducted at the frequency of 534 kHz. In Fig. 7.3.5 solid lines showed the results of the measurements, carried out in the presence of negative feedback in the frequency, and dotted line - with the usual frequency discriminator. From these curves it follows that the optimum value of the deviation of frequency for one channel proves to be about 1 MHz.

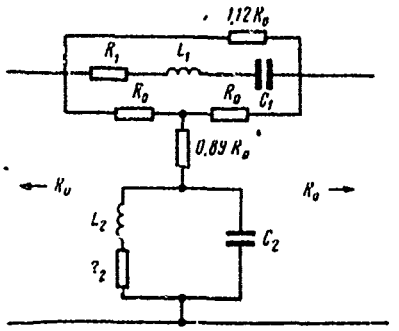

Fig. 7.3.3.

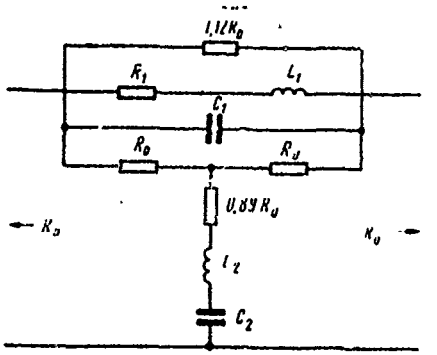

Fig. 7.3.4.

Fig. 7.3.3. Schematic of predistorting filter.

Fig. 7.3.4. Schematic of restoring filter.

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At present still there are no recommendations of MKKR for selection of deviation of frequency for communication systems through ISZ. Nevertheless in MKKR was conducted the discussion of this question [7.9, 7.11], in Fig. 7.3.6 solid lines gave the curves, characterizing the effective value of deviation to channel $\Delta f_к$ without the introduction of predistortions and width of band Δf in the dependence on a number of channels N in the communication system through ISZ. In Fig. 7.3.6 points showed the values of the effective deviation of frequency $\Delta f_к$, accepted by MKKR for RRL, which correspond to supply to the input of telephone channel RRL of measuring sine wave with the power of 1 mW.

Calculation of width of band Δf, occupied during transmission of multichannel telephone signal with ChM, and effective value of

deviation of frequency during transmission N of channels Δf_{cp} can be made according to expressions:

$$\left.\begin{aligned} \Delta f &= 2(\Delta f_{пик} + F_д) \\ \Delta f_{пик} &= \Delta f_{ср} \varkappa'' \\ \Delta f_{ср} &= \Delta f_к 10^{\frac{p_{ср}(дБ)}{20}} = \Delta f_к e^{p_{ср}(N_{ср})} \\ \varkappa'' &= 10^{\frac{\varkappa}{20}} \end{aligned}\right\}. \qquad (7.3.7)$$

Entering these expressions values are determined according to given earlier formulas and graphs. Thus, value $p_{ср}$ is located through formulas (6.2.4) or (6.2.5), values $F_д$ are given in Table 6.2.1, and the value of peak factor $\varkappa$ is located according to the graphs Fig. 6.2.5. During the determination of band it suffices to take the values $\varkappa$, which correspond to ϵ=0.01.

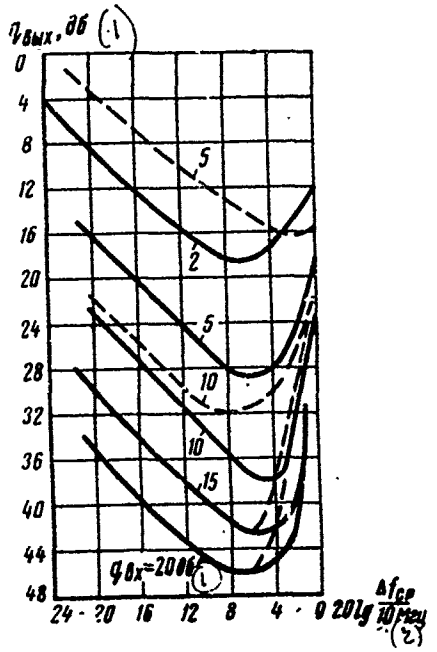

Fig. 7.3.5. Results of tests during charging by noise signal, equivalent N=240: —— demodulator with OSCh, - - - usual demodulator, frequency of experimental tone 534 kHz.

Key: (1). dB. (2). MHz.

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Let us note that curves Δf, given in Fig. 7.3.6, are designed upon consideration (6.2.4) - solid line and (6.2.5) - dotted line with ϵ=0.001.

Transient noises, which appear as a result of nonlinearity of amplitude characteristic of group amplifiers and characteristics of ChM modulator and frequency discriminator. The onset of transient noises with the passage of the multichannel signal through the nonlinear amplitude characteristics of group amplifiers, and also due

to the nonlinearity of the characteristics of ChM modulator and frequency discriminator, is analogous with that examined in Section 7.2. Therefore the calculation of the level of this form of noises can be implemented according to expressions (7.2.33)-(7.2.35).

Transient noises, which appear as a result of nonlinearity of phase responses of circuits of shf/SVCh and PCh. During the analysis of these noises we will use quasi-stationary method. This means that the multichannel telephone signal is assumed this slowly changing function of time that the processes of establishment do not play the significant role, in connection with which the static, and not dynamic, frequency and phase responses of the elements of circuit can be considered.

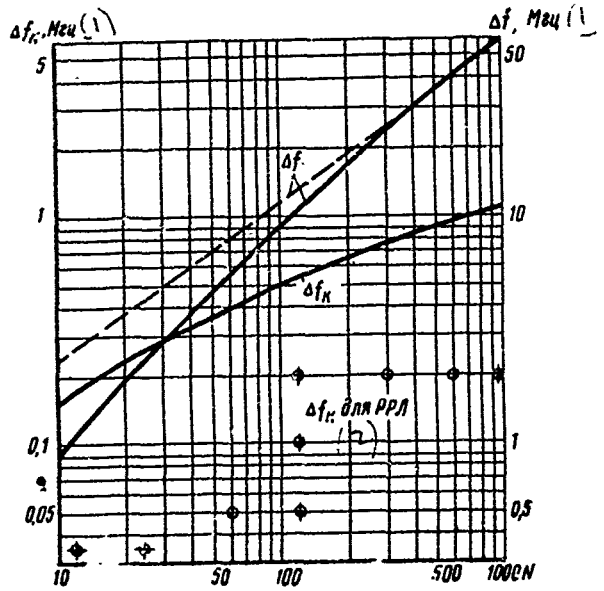

Fig. 7.3.6. Dependence $\Delta f_к$ and Δf on number of channels.
Key: (1). MHz. (2). for RRL.

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We will consider that frequency and phase responses of elements, according to which passes ChM signal, can be respectively approximated by equations:

$$\left.\begin{aligned} K(\delta\omega) &= e^{-\gamma_2\omega^2} \\ \varphi(\delta\omega) &= \psi_1\delta\omega + \psi_2\delta\omega^2 + \psi_3\delta\omega^3 + \psi_4\delta\omega^4 + \psi_5\delta\omega^5 \end{aligned}\right\}. \quad (7.3.8)$$

Here detuning in frequency $\delta\omega=\omega-\omega_0$, moreover ω_0 - average angular frequency in passband, and ω average angular frequency in passband, and ω - instantaneous value.

Let us note that expression (7.3.8) it corresponds to frequency characteristic of tuned amplifier with infinite number of

single-circuit steps/stages of amplification.

As V. A. Smirnov [7.12] showed, analysis of transient noises, caused by nonlinearity of characteristics (7.3.8), can be substantially simplified, if between higher frequency of spectrum of modulating oscillations $F_в$ and passband of elements of circuit Δf condition

$$F_в \ll \Delta f, \tag{7.3.9}$$

is satisfied, but coefficients, entering resulting expressions, are so low that value γ is of the order $1/(2\pi\Delta f)^2$, and coefficients ψ_n - order $\frac{1}{(2\pi\Delta f)^n}$. When these conditions and in circuit amplitude limiter is present, it is possible to disregard the effect of the nonlinearity of frequency characteristic manifested the stronger, the less relationship/ratio (7.3.9) accurately is fulfilled. Taking into account the aforesaid, in accordance with [7.12], we will consider that if to the input are given ChM oscillations, described by expression (5.3.1), then at the output, after the passage through the ideal limiter and the frequency discriminator, oscillations can be recorded in the form

$$U_{вых}(t) = AV(t) + \varepsilon_\varphi(t). \tag{7.3.10}$$

Here A - proportionality factor, not depending on frequency and time, ε_φ - determines products of nonlinear distortions due to phase response, described by expression (7.3.8), moreover

$$\varepsilon_\varphi(t) = \frac{d}{dt}\left[\psi_2\Delta\omega_m V^2(t) + \psi_3\Delta\omega_m^2 V^3(t) + \psi_4\Delta\omega_m^3 V^4(t) + \psi_5\Delta\omega_m^5 V^5(t)\right]. \tag{7.3.11}$$

According to (5.3.2), entering this expression value V(t) is

initial communication/report, transmitted by communication system, i.e., in this case multichannel telephone signal.

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From comparison of expressions (7.3.11) and (7.2.14) follows their identity, if we assume:

$$\begin{aligned} a_3 &= \psi_3 \Delta\omega_m, \\ a_3 &= \psi_3 \Delta\omega_m^2, \\ a_4 &= \psi_4 \Delta\omega_m^3, \\ a_5 &= \psi_5 \Delta\omega_m^4. \end{aligned} \qquad (7.3.12)$$

Therefore during replacement of coefficients according to (7.3.12) we have

$$\varepsilon_\varphi(t) = \frac{d}{dt}\varepsilon(t). \qquad (7.3.13)$$

Thus, the distortions, which appear with passage of ChM oscillations/vibrations through the elements with the nonlinear phase response, will be determined by derivative of value $\varepsilon(t)$, examined in Section 7.2. In connection with this the compensation for the nonlinearity of phase response by the nonlinearity of the characteristics of modulators, frequency discriminator or by the nonlinearity of the amplitude characteristics of group and video amplifier is impossible.

For calculating transient noises, caused by nonlinearity of phase response, let us take into account expression (П.5.19), which connects spectral densities of two stationary functions, one of which is equal to time derivative of another. This will make it possible

during calculations of noises to use the results, obtained in the sections. 7.2. Thus, taking into account (7.3.13), on the basis (7.5.19) we obtain

$$G_\varphi(\Omega) = \Omega^2 G_\varepsilon(\Omega). \qquad (7.3.14)$$

Here $G_\varphi(\Omega)$ - energy spectrum of distortions, caused by passage through element with nonlinear phase response, and $G_\varepsilon(\Omega)$ is determined by expression (7.2.26), in which coefficients a should be replaced in accordance with (7.3.12).

As a result of substitution (7.2.26) in (7.3.14) and conversions, analogous carried out during conclusion/output of expressions (7.2.29)-(7.2.32), we will obtain expression for determination psopnometrically weighted power of transient noises at output of telephone channel, caused by nonlinearity of phase response:

$$P_\varphi = \frac{\Delta F_\kappa \kappa_n^2}{\Delta F}(2\pi F_\kappa)^2 \left|2\psi_2^2 \Delta\omega_m^2 R_\kappa e^{4p_{cp}} y_2(\xi) + 6\psi_3^2 \Delta\omega_m^4 R_\kappa^2 e^{6p_{cp}} y_3(\xi) + \right.$$
$$\left. + 24\psi_4^2 \Delta\omega_m^6 R_\kappa^3 y_4(\xi) e^{8p_{cp}} + 120\psi_5^2 \Delta\omega_m^8 R_\kappa^4 e^{10p_{cp}} y_5(\xi)\right]. \qquad (7.3.15)$$

This expression is correct for point with zero relative level. Let us find the values of coefficients $\psi^2\Delta\omega^2_m$ entering expression (7.3.15).

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For this we will consider that the modulating voltage/stress is harmonic test signal with frequency Ω_m

$$V(t) = U_{\text{к}} \sin \Omega_{\text{к}} t = \sqrt{2P_{\text{к}}R_{\text{к}}} \sin \Omega_{\text{к}} t, \qquad (7.3.16)$$

moreover in accordance with (P.4.8) $p_{\text{к}} = e^{2p_{\text{к}}} = \frac{U_{\text{к}}^2}{2R_{\text{к}}}$.

With passage of oscillation/vibration, frequency-modulated by harmonic signal with frequency of $\Omega_{\text{к}}$, through elements with nonlinear phase terminal characteristics of frequency discriminator to be obtained oscillation/vibration with frequencies of $2\Omega_{\text{к}}$, $3\Omega_{\text{к}}$ and so forth. Designating the appropriate attenuation indices through b_2, b_3, b_4, b_5, analogous how this was done in Section 7.2, equalizing (7.3.12) and (П.6.22), upon consideration (7.2.30), (7.2.31), after conversions let us find:

$$\psi_2^2 \Delta\omega_m^2 = \frac{2}{R_{\text{к}}(2\Omega_{\text{к}})^2} e^{-2b_2}; \quad \psi_4^2 \Delta\omega_m^6 = \frac{8}{R_{\text{к}}^3(4\Omega_{\text{к}})^2} e^{-2b_4},$$

$$\psi_3^2 \Delta\omega_m^4 = \frac{4}{R_{\text{к}}^2(3\Omega_{\text{к}})^2} e^{-2b_3}; \quad \psi_5^2 \Delta\omega_m^8 = \frac{16}{R_{\text{к}}^4(5\Omega_{\text{к}})^2} e^{-2b_5}. \qquad (7.3.17)$$

Therefore instead of (7.3.15) we will obtain

$$P_{\text{ф}} = \frac{\Delta F_{\text{к}} \kappa_{\text{п}}^2}{\Delta F} (2\pi F_{\text{к}})^2 \frac{1}{(2\pi F_{\text{м}})^2} \left[e^{-2b_2} e^{4p_{\text{ср}}} y_2(\xi) + \right.$$
$$\left. + \frac{8}{3} e^{-2b_3} e^{6p_{\text{ср}}} y_3(\xi) + 12 e^{-2b_4} e^{8p_{\text{ср}}} y_4(\xi) + \frac{384}{5} e^{-2b_5} e^{10p_{\text{ср}}} y_5(\xi) \right], \text{ мВт}. \qquad (7.3.18)$$

Taking into account (П.6.15), we will obtain expressions for determining transient noises through percentage distortions

$$P_{\text{ф}} = e^{2p_{\text{ср}}} \frac{\Delta F_{\text{к}} \kappa_{\text{п}}^2}{\Delta F} \left(\frac{\Omega_{\text{к}}}{\Omega_{\text{м}}} \right)^2 [K_2^2 y_2(\xi) + 3e^{2p_{\text{ср}}} K_3^2 y_3(\xi) +$$
$$+ 12 e^{4p_{\text{ср}}} K_4^2 y_4(\xi) + 77 e^{6p_{\text{ср}}} K_5^2 y_5(\xi)], \text{ мВт}. \qquad (7.3.19)$$

Let us note that expression (7.3.18) is analogous with expression (7.2.33), and (7.3.19) it is analogous with expression (7.2.35). However, in expressions (7.3.18) and (7.3.19) essential is the fact that the value of the noises, caused by the nonlinearity of phase response, is determined by the ratio of the frequency of telephone channel $\Omega_к$ to the frequency of the measuring tone, with the help of which is produced the determination of all indices of nonlinearity in expression (7.2.18) or all percentage distortions in expression (7.3.19). In this case the level of experimental tone must be equal to the normal measuring level of channel.

In practice of expression (7.3.18) and (7.3.19) sometimes more conveniently to represent in another form.

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This is connected with the fact that the coefficients ψ, entering in (7.3.15), can be determined, if the characteristic of group time lag is known. Actually, if phase response is determined by expression (7.3.8), then the group time lag of signal will be

$$\tau = \frac{d\varphi}{dt} = \psi_1 + 2\psi_2\delta\omega + 3\psi_3\delta\omega^2 + 4\psi_4\delta\omega^3 + 5\psi_5\delta\omega^4. \qquad (7.3.20)$$

Here value ψ_1 determines time lag of signal τ_0 at frequency ω_0, which is midband frequency of transmission of element of circuit in question. The graphic dependences of phase response and group time lag, determined by expressions (7.3.8) and (7.3.20), they are shown in Fig. 7.3.7. From this figure it is possible to determine values

$(\psi_n \Delta\omega_m^{n-1})$ entering in (7.3.20), if are known (or they are assigned) the values of a change in the group time with the symmetrical frequency deviation from the average/mean value for values $\pm\Delta\omega_к$ and $\pm a\Delta\omega_к$. It is obvious that for obtaining the deviation of frequency, equal to $\pm\Delta\omega_к$, to the channel must be given the measuring level, and for obtaining the deviation of frequency $\pm a\Delta\omega_к$ this level must be changed in a of times. With the substitution of these values of the deviation of frequency in (7.3.20) in accordance with the designations in Fig. 7.3.7 we will obtain:

$$\left.\begin{aligned}
\tau(\Delta\omega_к) &= 2\psi_2\Delta\omega_к + 3\psi_3\Delta\omega_к^2 + 4\psi_4\Delta\omega_к^3 + 5\Delta\omega_к^4\psi_5\\
\tau(-\Delta\omega_к) &= -2\psi_2\Delta\omega_к + 3\psi_3\Delta\omega_к^2 - 4\psi_4\Delta\omega_к^3 + 5\Delta\omega_к^4\psi_5\\
\tau(a\,\Delta\omega_к) &= 2\psi_2 a\,\Delta\omega_к + 3\psi_3 a^2\Delta\omega_к^2 + 4\psi_4 a^3\Delta\omega_к^3 + 5\psi_5 a^4\Delta\omega_к^4\\
\tau(-a\,\Delta\omega_к) &= -2\psi_2 a\,\Delta\omega_к + 3\psi_3 a^2\Delta\omega_к^2 - 4\psi_4 a^3\Delta\omega_к^3 + 5\psi_5 a^4\Delta\omega_к^4
\end{aligned}\right\}. \quad (7.3.21)$$

Solving jointly these expressions, we obtain formulas given below, which for reduction of recording we will designate through τ with index, which shows order of nonlinearity:

$$\left.\begin{aligned}
\psi_2\Delta\omega_к &= \frac{a^3[\tau(\Delta\omega_к) - \tau(-\Delta\omega_к)] - [\tau(a\,\Delta\omega_к) - \tau(-a\,\Delta\omega_к)]}{4a(a^3-1)} = \tau_2,\\
\psi_3\Delta\omega_к^2 &= \frac{a^4[\tau(\Delta\omega_к) + \tau(-\Delta\omega_к)] - [\tau(a\,\Delta\omega_к) + \tau(-a\,\Delta\omega_к)]}{6a^3(a^2-1)} = \tau_3,
\end{aligned}\right\} \quad (7.3.22)$$

$$\left.\begin{aligned}
\psi_4\Delta\omega_к^3 &= \frac{a[\tau(\Delta\omega_к) - \tau(-\Delta\omega_к)] - [\tau(a\,\Delta\omega_к) - \tau(-a\,\Delta\omega_к)]}{8a(1-a^2)} = \tau_4,\\
\psi_5\Delta\omega_к^4 &= \frac{a^2[\tau(\Delta\omega_к) + \tau(-\Delta\omega_к)] - [\tau(a\,\Delta\omega_к) + \tau(-a\,\Delta\omega_к)]}{10a^3(1-a^2)} = \tau_5.
\end{aligned}\right\} \quad (7.3.22)$$

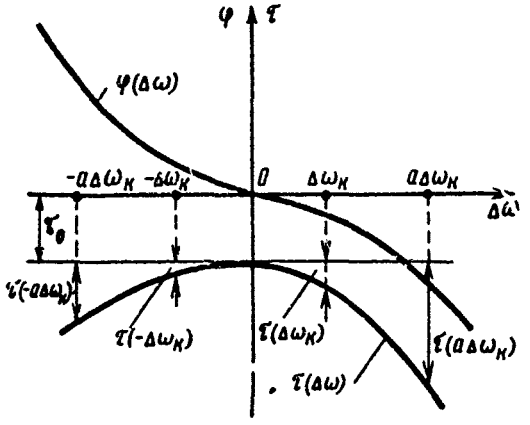

Fig. 7.3.7. Phase response and characteristic of group time lag.

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Taking into account that maximum (peak) level of multichannel signal, which we will consider equal to one, creates peak deviation of frequency $\Delta\omega_m$, and maximum deviation of frequency, created by sine wave with amplitude of $U_к$ is equal to $\sqrt{2}\,\Delta\omega_к$ (here $\Delta\omega_к=2\,\pi\Delta f_к$ - effective value of deviation of frequency, created by sine wave signal), it is possible to compose following equality:

$$\frac{\Delta\omega_m}{\sqrt{2}\Delta\omega_к}=\frac{1}{U_к}.$$

Since $\frac{U_к^2}{2R_к}=P_к=e^{2p_к}$, where $p_к$ - level of sine wave, it is possible to record

$$\Delta\omega_m=\frac{\Delta\omega_к}{\sqrt{R_к}}e^{-p_к}. \qquad (7.3.23)$$

Substituting (7.3.23), (7.3.22) and formula (7.3.15) with value $p_к=0$, let us find power of noises, caused by nonlinearity of phase response, for point with zero relative level

$$P_{\text{ш}} = \frac{\Delta F_{\text{ш}} \kappa_{\text{п}}^2}{\Delta F} \Omega_{\text{к}}^2 [2\tau_2^2 e^{4p_{\text{ср}}} y_2(\xi) + 6\tau_3^2 e^{6p_{\text{ср}}} y_3(\xi) + 24 e^{8p_{\text{ср}}} y_4(\xi) \tau_4^2 + 120 e^{10p_{\text{ср}}} \tau_5^2 y_5(\xi)], \text{ mW.} \quad (7.3.25).$$

If power of noises is determined at point with level $p_{\text{ш}}$ before expression (7.3.25) it must stand factor $e^{2p_{\text{к}}}$. Entering this expression values τ are determined according to the specific characteristic of group time lag UPCh in accordance with Fig. 7.3.7 on formulas (7.3.22).

According to expression (7.3.25) it can be solved and inverse problem - on assigned value $P_{\text{ш}}$ it is possible to determine values τ_2-τ_5, and then from expressions (7.3.22) to find relationship/ratio between parameters $\tau(\pm\Delta\omega_{\text{к}})$, $\tau(\pm a\Delta\omega_{\text{к}})$ and a and thus to determine requirements for values of group time lag and for phase response of circuit.

For calculating values τ_2-τ_5 according to expression (7.3.25) it is necessary to assign exemplary/approximate relationships/ratios between power of noises of different order, for example, to consider that power of noises of different order is identical, i.e.,

$$2\tau_2^2 e^{4p_{\text{ср}}} y_2(\xi) = 6\tau_3^2 e^{6p_{\text{ср}}} y_3(\xi) = 24\tau_4^2 e^{8p_{\text{ср}}} y_4(\xi) = 120\tau_5^2 e^{10p_{\text{ср}}} y_5(\xi). \quad (7.3.26)$$

Can be accepted other relationships/ratios. Let us note that in the case, when phase response can be approximated by the polynomial of the third power, value $\psi_4=\psi_5=0$; therefore according to (7.3.22) $\tau_4=\tau_5=0$. This considerably simplifies expression (7.3.25), which in this case for the point with level $p_к$ can be recorded in the form

$$P_\varphi = e^{2p_к}\frac{\Delta F_к \kappa_п^2}{\Delta F}\Omega_к^2[2\tau_2^2 e^{4p_{ср}} y_2(\xi) + 6\tau_3^2 e^{6p_{ср}} y_3(\xi)], \text{ mW}. \quad (7.3.27).$$

Now for determination τ_2 and τ_3 it suffices to know only parameters $\tau(\Delta\omega_к)$ and $\tau(-\Delta\omega_к)$ or $\tau(a\Delta\omega_к)$ and $\tau(-a\Delta\varphi_к)$ (see Fig. 7.3.7), changes in group time characterizing dependence on frequency, moreover from solution of equ. (7.3.21) with $\psi_4=\psi_5=0$, let us find:

$$\left.\begin{aligned}\tau_2 &= \frac{\tau(\Delta\omega_к)-\tau(-\Delta\omega_к)}{4} = \frac{\tau(a\Delta\omega_к)-\tau(-a\Delta\omega_к)}{4a}\\ \tau_3 &= \frac{\tau(\Delta\omega_к)+\tau(-\Delta\omega_к)}{6} = \frac{\tau(a\Delta\omega_к)+\tau(-a\Delta\omega_к)}{6a^2}\end{aligned}\right\}. \quad (7.3.28)$$

Value a can be any number.

For evaluation/estimate of order of magnitudes of parameters τ_2 and τ_3 in the case, when phase response is approximated by polynomial of third power, Fig. 7.3.8 gives dependence between parameters $\tau(\Delta\omega_к)$ and $\tau(-\Delta\omega_к)$ for upper telephone channel at transmission 60 and 600 channels and different amount of power of noises P_φ, expressed in pW. Graph is constructed according to expressions (7.3.27) and (7.3.28) for the point with zero relative level $(p_к=0)$.

From examination of expression, which determines value of transient noises, caused by nonlinearity of phase response, it follows that these noise in upper telephone channels will be considerably more than in lower. Let us note that the transient noises, determined by the nonlinearity of the group circuit (see expressions (7.2.33) (7.2.35) and the graphs, given in Fig. 7.2.4-7.2.7, comparatively little depend on the location of telephone channel in the group spectrum.

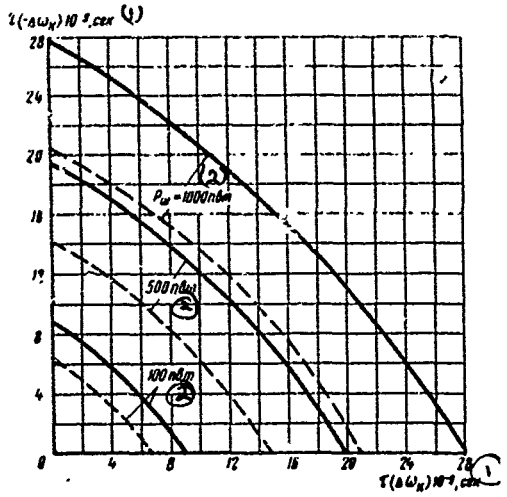

Fig. 7.3.8. Relationships/ratios between $\tau(\Delta\omega_к)$ and $\tau(-\Delta\omega_к)$ - - - - N=600, ---- N=60.

Key: (1). s. (2). pW.

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Reduction in noises in communication systems is necessary to realize by correction of phase response. Equalization of phase response, and consequently, and group time lag, must be realized in the entire passband without a change in the frequency characteristic. Therefore compensating circuits in the passband must have a transmission factor, not depending on the frequency. As compensating circuits different bridge circuits most frequently are used. Usually correction is realized at the intermediate frequency.

7.4. Transmission with KIM.

Transmission of telephone signals in the case KIM is realized according to diagram Fig. 7.4.1. In this diagram signals of subscribers after coding device/equipment KY and equipment for time-division multiplex AY proceed to modulator M. It realizes modulation of amplitude (Kims - AM), frequency (KIM-ChM) or phase (KIM-FM) of the oscillations/vibrations of the transmitter, which works at frequency f_0. The modulated oscillations/vibrations on the feeder proceed to antenna A' and are emitted in the direction of satellite (repeater).

On satellite of oscillation/vibration Ap', are amplified by receiver Пр and after passage of limiter Orp enter auxiliary step/stage BC, where they are displaced in frequency, are accepted as antenna. Then oscillations/vibrations, after the passage of transmitter, are emitted by antenna Ap in the direction of the Earth. Let us note that into the composition of the step/stage VS of repeater can in a number of cases enter the system of the regeneration of the pulses/momenta accepted, as a result of which at the output repeater the pulse impulses/transmissions, free from the fluctuation noises, will be obtained.

Oscillations/vibrations, emitted by antenna of repeater Ap, are

accepted to antenna terrestrial station A" and after passage through receiver Пр and limiter proceed to demodulator, equipment for separation of cnannels АУ and decoder ДУ. From the output ДУ the taken signal is supplied to the subscribers.

Principle of transmission of telephone signals examined with KIM can be realized in the form of two variants: in the absence of pulse regeneration in repeater and during introduction to regeneration.

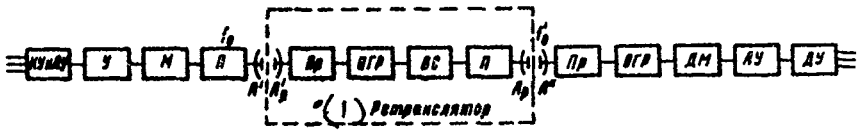

Fig. 7.4.1. Block diagram of transmission from KIM.

Key: (1). Retranslator.

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It is obvious that the quality of reception/procedure in the first version will be determined by general/common effect on the signal of the noises of repeater, and the noises of terrestrial receiver. In this case most appropriate will be such distribution of energy relationships/ratios in the sections of connection/communication, during which the signal-to-noise ratio at the input of onboard receiver will prove to be considerably more than at the input of terrestrial receiver. This is connected with the fact that the creation of large power under the terrestrial conditions proves to be considerably simpler and it is more economical than under the space conditions. This is correct for the second version. The difference between the versions consists in the fact that in the first the more powerful terrestrial transmitter will be required, whereas the secondly - will be required more complicated, installed equipment of satellite, which switches on the system of pulse regeneration,

demodulator of receiver and modulator of transmitter.

Thermal noises. For each section of connection/communication on the basis (5.1.3) it is possible to record the following expression:

$$q_{\text{вых }a} = q_{\text{вх }a}\beta_a,$$

where index a determines the number of the section in question.

In accordance with this psophometrically weighed power of thermal noises, in reference to point, at which measuring output signal level is equal to 1 mW, will be

$$P_{\text{ш}a} = \frac{\kappa_{\text{п}}^2}{q_{\text{вх }a}\beta_a}, \text{ mW}.$$

With substitution into this expression of relationships/ratios (5.4.19) - (5.4.21) we will obtain:

with to Kims

$$P_{\text{ш}a} = \frac{\kappa_{\text{п}}^2 \varkappa_{\text{п}}^2}{2q_{\text{вх }a}} e^{-\frac{q_{\text{вх }a}}{2}}, \qquad (7.4.1)$$

with KIM-ChM

$$P_{\text{ш }a} = \frac{\kappa_{\text{п}}^2 \varkappa_{\text{п}}^2}{2\sqrt{2}\sqrt{q_{\text{вх }a}}} e^{-q_{\text{вх }a}}; \qquad (7.4.2)$$

with KIM-FM

$$P_{\text{ш }a} = \frac{\kappa_{\text{п}}^2 \varkappa_{\text{п}}^2}{4\sqrt{q_{\text{вх }a}}} e^{-2q_{\text{вх }a}}. \qquad (7.4.3)$$

In all these expressions value

$$q_{\text{вх }a} = \frac{kT_{\text{пр}a}\Delta f_a B_{\text{пр}a}}{P_{\text{п}a}}. \qquad (7.4.4)$$

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For definition of power of transmitters, established/installed on board satellite or on Earth with given value of psophometrically weighed power of noises $P_{ш\,зад}$ (mV), expressions (7.4.1)-(7.4.3) for case of measurements by simple tone can be rewritten as:

with to Kims

$$\sqrt{q_{вх\,а}}\,e^{\frac{q_{вх\,а}}{2}} = 0{,}56/P_{ш\,зад}; \qquad (7.4.5)$$

with KIM-ChM

$$\sqrt{q_{вх\,а}}\,e^{q_{вх\,а}} = 0{,}4/P_{ш\,зад}; \qquad (7.4.6)$$

with KIM-FM

$$\sqrt{q_{вх\,а}}\,e^{2q_{вх\,а}} = 0{,}28/P_{ш\,зад}. \qquad (7.4.7)$$

Thus, knowing right side of these relationships/ratios, it is easy to determine values $q_{вх\,а}$ for different forms of modulation. For convenience of the determination of value $q_{вх}$ Fig. 7.4.2 gives graphic relationships/ratios $q_{вх}$ and products $\sqrt{q_{вх}}e^{mq_{вх}}$, which stand on the left side of three latter/last expressions. Thus, after determining according to the graphs Fig. 7.4.2 value q, from expression 7.4.4 it is possible to find the power of transmitter $P_{пд}$.

Let us note that expressions (7.4.1) - (7.4.3) and (7.4.5) - (7.4.7) are obtained for idealized communication system. Therefore during calculations should be considered some other relationships/ratios. For the communication systems with KIM-FM it is

established [7.14], that in the case, when the probability of the erroneous reception/procedure of the elementary impulses/transmissions of code groups id equal $p=3\cdot10^{-5}$, the power of noises at the output of telephone channel are 20000 pW, and value $q_{вх}=8$. With a change of value p, into the same number of times is changed value $q_{вх}$ and power of noises in telephone channel. On the other hand, relationship/ratio between values p and values $q_{вх}$ they are established by graph on Fig. 5.4.1 and by expression on page 132. On the basis of both these conditions was constructed dependence in the form of the dotted curve in Fig. 7.4.2, which makes more precise the connection/communication between product $q_{вх}e^{mq_{вх}}$ and with value $q_{вх}$ for the nonidealized system KIM-FM.

For case of real system to Kims Fig. 7.4.2 gives second dotted curve, which was determined in terms of values Tables 5.4.1.

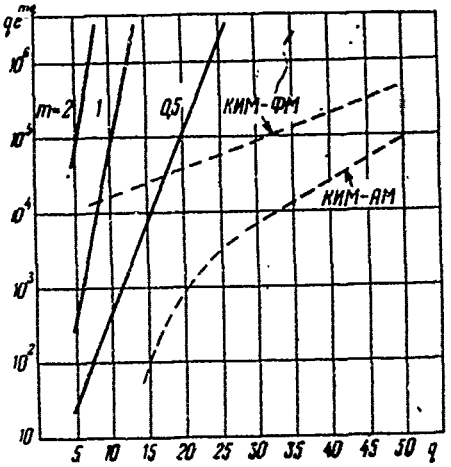

Fig. 7.4.2. Values $\sqrt{q}e^{mq}$ and q.

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Width of band Δf_m, entering expression (7.4.4), determines minimum band, necessary for recpetion of signals.

With transmission of pulses/momenta in the case AM and single FM, with which phase of modulated oscillation/vibration is changed on 180°, width of band is identical and is determined of condition for passage of first side frequencies. Therefore width of band of idealized receiver with account (6.4.3) must be not less

$$\Delta f_a = 2\Delta F = 16{,}8 \cdot 10^3 \, nN, \text{ Hz}. \quad (7.4.8).$$

Upon consideration of instabilities of frequencies of transmitter and heterodyne of receiver bandwidth must be

correspondingly increased.

Expression (7.4.8) is correct for FM in the case, when upon transfer from pulsing to pause phase of modulated oscillation/vibration is changed on 180°. For the reduction of band in a number of cases is used repeated (multiposition) FM, with which the transmitted sequence of pulses/momenta is divided/marked off into several groups. With twofold FM (DFM) with the equal values of a change in the phase angle the sequence of pulses/momenta is divided/marked off into two independent groups, applied to the modulator along two channels.

Depending on they go along both channels impulses/transmissions, pauses or their combination, is changed the phase angle of carrying oscillation/vibration. The latter is clarified by Tables 74.1.

As a result of applying DFM in band, on which in the case AM is transmitted N channels, can be transmitted 2 N channels. Therefore, taking into account expression (7.4.5), for the case DFM we will obtain

$$\Delta f_a = 8{,}4 \cdot 10^3\, nN, \text{ Hz} \quad (7.4.9)$$

Let us note that sometimes DFM call quadrature of phase modulation (KFM).

One should note that with increase in multiplicity of phase modulation value of shift/shear of phase angle between adjacent values is reduced, which decreases noise immunity of communication system.

Table 7.4.1.

Вид передачи по каналу 1 (1)	(2) посылка	(3) пауза	(3) пауза	(2) посылка
По каналу 2 (4)	(2) посылка	(2) посылка	(3) пауза	(3) пауза
(5) Значение фазы несущего колебания	0°	90°	180°	270°

Key (1). Transmission mode along channel 1. (2). impulse/transmission. (3). pause. (4). Along channel 2. (5). Value of carrier frequency oscillation/vibration.

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Bandwidth, occupied by signal in the case ChM with index of modulation m=16, proves to be equal to width of band AM of signal [7.13]. Therefore with the indices of modulation m<16 it is possible to consider that the necessary frequency band in the case ChM will be the same as with AM and single FM, i.e., will be determined by expression (7.4.8).

Noises of quantization. The power of psophometrically weighed quantization noises at point with the zero relative level can be found from expression (5.4.4)

$$P_{\text{ш кв}} = \frac{x^3 \kappa_{\text{п}}^2}{3 \cdot 2^{2n}}, \text{ mW.} \quad (7.4.10).$$

7.5. Mixed transmission modes of multichannel telephone signals.

By mixed transmission modes of multichannel telephone signals we will understand case, when in sections of line of communications Earth-satellite-Earth are used different forms of modulation. A change in the methods of modulation can be stipulated on different reasons. To the bases of them can be attributed the following: an increase in noise immunity in the section of connection/communication a satellite-Earth with the simultaneous decrease of the bandwidth in the section Earth-satellite, the creation of the system, which ensures the repeated use of a repeater for the simultaneous transmission of the signals of several terrestrial stations or as occasionally is referred to as this type of work, for the creation of system with multistation access, etc.

In the first of enumerated cases, taking into account possibility of designing of sufficiently large power in section of connection/communication Earth-satellite, is possible use of following versions of modulation of oscillations:

1) OBP in section Z- S and ChM in section S - Z,

2) OBP in section Z - S and KIM in section S - Z.

3) ChM with narrow band in section Z- S and ChM with broad band in section S - Z,

4) ChM in section Z - S and KIM in section S - Z.

Let us note that first and third versions of modulation can be carried out in such a way that they will not require demodulation of oscillations on repeater. This fact is very important, since ensures not only the simpler construction of equipment of repeater with a simultaneous increase in its reliability, but also excludes the possibility of the appearance of the transient noises, which will unavoidably arise in the process of detecting of the high-frequency oscillations, amplification of the group spectrum and its conversions with modulation of onboard transmitter.

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Versions 2 and 4 lead to more complicated equipment of repeater, since they require presence on board complete complex of equipment, which ensures obtaining KIM and corresponding modulation devices/equipment.

Calculation of value of noises, which are obtained at output of line of communications or calculation of power of transmitters with given values of noises for individual sections of line is realized according to expressions, given in previous sections of this chapter.

7.6. Noises at the output of channel.

As already mentioned, at output of telephone channel is obtained thermal and transient noise, whose overall value must not exceed values, determined by norms MKKR [MKKP - International Radio Consultative Committee]. Therefore during calculations of systems and lines of communications it is necessary to know the laws of the addition of the noises, which appear in the elements of equipment of terrestrial stations and communication satellite.

If processes in question are statistically independent random processes, then their average/mean values are summarized. In this case the total power of processes is equal to the sum of power, i.e.,

$$P_{\Sigma} = \sum_{i=1}^{m} P_i. \qquad (7.6.1)$$

Thermal noises is example of such processes in communication system through ISZ, since phenomena, which cause appearance of thermal noises in onboard and ground-based devices/equipment, are independent from each other.

In the case, when coefficient of correlation of random processes in question is equal to one, i.e., when random processes prove to be in accuracy repeating or differ from each other to constant factor,

it is necessary to summarize instantaneous values of processes - voltage/stress u or current i. Since the values of voltages/stresses or currents during the known resistor/resistance of R can be expressed through the power of each process P_i, the total power of the processes examined/considered by m can be recorded in the form

$$P_{\Sigma} = \frac{1}{R}\left[\sum_{i=1}^{m} U_i\right]^2 = \frac{1}{R}\left[\sum_{i=1}^{m} \sqrt{P_i R}\right]^2 = \left[\sum_{i=1}^{m} \sqrt{P_i}\right]^2. \quad (7.6.2)$$

Transmission with CCC OBP. In this case during the transmission of signal through the circuit at the output of telephone channel the part of the noises will be determined by thermal processes, and the part of the noises - by nonlinearity of circuit.

Since thermal and transient noises are independent, their total power, in accordance with (7.6.1), will be determined by sum of power.

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However, a question about the addition between themselves of the transient noises, which will be obtained with the passage of the signal through the series-connected amplifiers and converters of frequency, spread from each other up to the considerable distance (part on the Earth, part on ISZ), requires more detailed examination. There is no at present comprehensive solution of this question still; therefore further examination is approximate.

We will consider that amplitude characteristics of amplifiers are described by polynomial of fifth degree according to (7.2.11), and phase shifts, introduced by amplifier, are absent. At the same time let us assume that the phase responses of the sections of the coupling links between the amplifiers are identical, linear in entire operational frequencies band and are described by the expression

$$\varphi = \varphi_0 + \varphi_1 \Omega. \tag{7.6.3}$$

Then during the supplying to input of first amplifier of two harmonic oscillations with frequencies of Ω_a and Ω_c, at its output, in accordance with (7.6.25) and (7.6.27), we obtain components with frequencies of Ω_a, Ω_c, $2\Omega_a$, $2\Omega_c$; $3\Omega_a$; $\Omega_a - \Omega_c$; $2\Omega_a - \Omega_c$ and so forth. With the passage of these components through one coupling link to the input of the second amplifier, in accordance with (7.6.3), will act the products of the nonlinear distortions of the following form:

$$\cos(\Omega_a t + \varphi_0 + \varphi_1 \Omega_a), \tag{7.6.4}$$

$$\cos(\Omega_c t + \varphi_0 + \varphi_1 \Omega_c), \tag{7.6.5}$$

$$\left.\begin{aligned}
&\cos(2\Omega_a t + \varphi_0 + \varphi_1 2\Omega_a)\\
&\cos(2\Omega_c t + \varphi_0 + \varphi_1 2\Omega_c)\\
&\cos(3\Omega_a t + \varphi_0 + \varphi_1 3\Omega_a)\\
&\cos(3\Omega_c t + \varphi_0 + \varphi_1 3\Omega_c).\\
&\cos[(\Omega_a \pm \Omega_c) t + \varphi_0 + \varphi_1 (\Omega_a \pm \Omega_c)]\\
&\cos[(2\Omega_a \pm \Omega_c) t + \varphi_0 + \varphi_1 (2\Omega_a \pm \Omega_c)]\\
&\cos[(2\Omega_a \pm 2\Omega_c) t + \varphi_0 + \varphi_1 (2\Omega_a \pm 2\Omega_c)]\\
&\cos[(3\Omega_a \pm 2\Omega_c) t + \varphi_0 + \varphi_1 (3\Omega_a \pm 2\Omega_c)]
\end{aligned}\right\}. \tag{7.6.6}$$

From appendix 6 it follows that components (7.6.4) and (7.6.5) have considerably greater amplitudes than components, described (7.6.6).

Let us consider now passage, enumerated components through second amplifier with nonlinear amplitude characteristic. With the substitution of expressions (7.6.4)-(7.6.6) in the first (linear) term of polynomial (7.2.11), by which is approximated the amplitude characteristic of the second amplifier, at its output will be obtained the components with the same combination frequencies and phase shifts, as at the input.

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Thus, due to the linear term of polynomial at the output of the second amplifier will be obtained the components, described by expressions (7.6.4)-(7.6.6).

With substitution of components (7.6.6) into nonlinear members of polynomial (quadratic, cubic, etc.) at output we will obtain more complicated components with new combination frequencies, which were absent in (7.6.6). However. in the conformity (7.6.27), the amplitudes of these components, with a small nonlinearity of characteristics, will be small and therefore they can be disregarded. With tne substitution into the nonlinear members of the polynomial of components (7.6.4) and (7.6.5) at the output of the second amplifier we will obtain the following products of nonlinear distortions:

$$\left.\begin{array}{l}\cos(2\Omega_a t + 2\varphi_0 + \varphi_1 2\Omega_a)\\ \cos(2\Omega_c t + 2\varphi_0 + \varphi_1 2\Omega_c)\\ \cos(3\Omega_a t + 3\varphi_0 + \varphi_1 3\Omega_a)\\ \cos(3\Omega_c t + 3\varphi_0 + \varphi_1 3\Omega_c)\\ \cos[(\Omega_a \pm \Omega_c)t + (\varphi_0 \pm \varphi_0) + \varphi_1(\Omega_a \pm \Omega_c)]\\ \cos[(2\Omega_a \pm \Omega_c)t + (2\varphi_0 \pm \varphi_0) + \varphi_1(2\Omega_a \pm \Omega_c)]\\ \cos[(2\Omega_a \pm 2\Omega_c)t + (2\varphi_0 \pm 2\varphi_0) + \varphi_1(2\Omega_a \pm 2\Omega_c)]\\ \cos[(3\Omega_a \pm 2\Omega_c)t + (3\varphi_0 \pm 2\varphi_0) + \varphi_1(3\Omega_a \pm 2\Omega_c)]\end{array}\right\}. \quad (7.6.7)$$

Taking into account what has been said and comparing (7.6.7) and (7.6.6), it is possible to draw conclusion that at output of second amplifier will not coincide in phase all components, except components with combination frequencies $2\Omega_a - \Omega_c$ and $3\Omega_a - 2\Omega_c$. More detailed analysis shows that at the output of element with the nonlinear amplitude characteristic will coincide all those to the product of nonlinearity, whose algebraic sum of coefficients at the component frequencies is equal to one. Thus, for example, for the components given above we have respectively following algebraic sums of the coefficients: 2-1=1; 3-2=1.

In accordance with this, with passage of oscillations through elements of circuit with nonlinear amplitude characteristics all products of nonlinearity can be decomposed into two groups: products of nonlinearity, whose phases coincide at outputs of all nonlinear elements, called products first-order nonlinearity, and second-order products, which include all remaining products of nonlinearity.

Let us note that in accordance with conclusion/output, given in examination (7.6.7), products of first-order nonlinearity can be formed only by members of polynomial with odd degrees higher than first. The members of polynomial with even degrees will create the products of the nonlinearity only of second kind.

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When amplifiers do not introduce phase shifts, but phase response of coupling links is linear, addition of products of first-order nonlinearity must be realized on voltage/stress, i.e., in accordance with (7.6.2). However, if the components/links, the connectives of circuit or very elements of circuit in the range of operating frequencies have nonlinear phase responses, the addition of the products of first-order nonlinearity must be realized according to the intermediate law between the addition according to the power and the addition on the voltage/stress. On the basis of this examination it is possible to draw the conclusion that with the divergence of phase response from the linear law the power of the products of first-order nonlinearity will be less than in the presence of the addition of components on the voltage/stress. This power for the case, when all elements possess identical amplitude characteristics, and coupling links - by identical phase and frequency characteristics, it can be approximately determined

according to expression [7, 16]:

$$P_{\text{вл}} = m^2 \nu^2 P_1. \qquad (7.6.8)$$

Here $P_{\text{вл}}$ - total power of products of first-order nonlinearity, P_1 - sum of power of separate components of nonlinearity of first kind, m - number of elements or components/links, and ν - coefficient, which with quadratic or cubic approximation of phase response depending on value of total phase shift ψ_Σ is determined according to graph in Fig. 7.6.1. Value φ_Σ is located through the expression:

$$\varphi_\Sigma = m\varphi_1, \qquad (7.6.9)$$

where m - number of coupling links or nonlinear elements, each of which introduces identical phase shift φ_1.

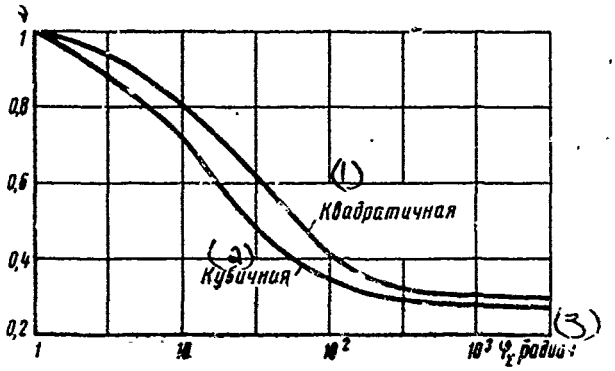

Fig. 7.6.1. Dependence of coefficient ν on total phase shift.

Key: (1). Quadratic. (2). Cubic. (3). radian.

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Value of phase shift φ_1 with known phase response in band of group spectrum with cut-off frequencies F_1-F_2 (or respectively Ω_1-Ω_2) is defined as follows: through point of phase response, which corresponds to middle of passband (to frequency F_0) straight line (z-z') is carried out in such a way, that at cut-off frequencies F_1 and F_2 segments zx=z'x' would be equal, as shown in Fig. 7.6.2.

Let us note that according to (7.6.8) with ν=1 addition of products of nonlinearity is realized on voltage/stress, and when ν^2=1/m - according to power. Therefore on the basis of graphs, Fig. 7.6.1, it is possible to draw the conclusion that in the case, when a number of elements and components/links in the line of communications is sufficiently large, then even with the low values φ_1 the addition

of the products of nonlinearity will approach an addition in the power. With a small number of components/links or elements so that the addition of the products of first-order nonlinearity would approach an addition in the power, in the communication system with OBP, apparently, it is expedient to introduce further elements with the phase shifts. Let us note, that in the communication systems through ISZ with the passage by the electromagnetic vibrations of free space phase shifts it will be variable in the time, that it is possible to lead to a change in the laws of addition and, consequently to a change in the value of transient noises at the output of channel.

On the basis of given examination, counting with approximation/approach, that all products, formed by terms relate to by nonlinear to first-order products, on the basis (7.6.8) we can record

$$P_{23.5} = m^2 v^2 (P_2 + P_5). \qquad (7.6.10)$$

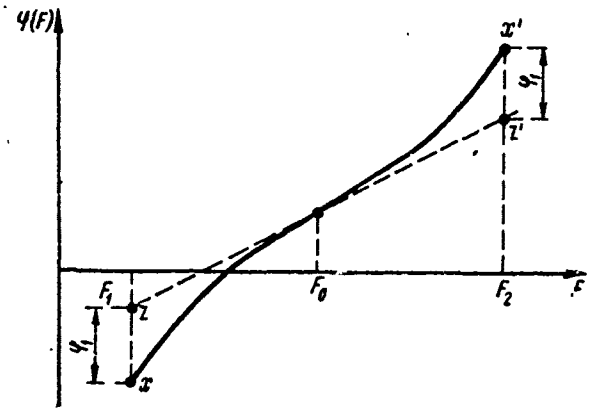

Fig. 7.6.2. determination of φ_1 from phase response.

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Here $P_{\text{шп}}$ - total power of nonlinear products, formed due to members of polynomial with odd degrees. Substituting in this expression of value P_3 and P_5 those found from (7.2.33) or (7.2.35), let us find the value of noises on the output of telephone channel at point with the zero relative level

$$P_{\text{ш3,5}} = m^{2\nu^2}\frac{\Delta F_{\text{к}}\kappa_{\text{п}}^2}{\Delta F}\left[24\,e^{6p_{\text{ср}}}\,y_3(\xi)\sum_{i=1}^{m}e^{-2b_{3i}} + 1920\,e^{10p_{\text{ср}}}\,y_5(\xi)\sum_{i=1}^{m}e^{-2b_{5i}}\right]. \tag{7.6.11}$$

Taking into account given analysis, which relates to products of nonlinearity in presence of phase shifts, it is possible with sufficient foundation to consider that products of second-order nonlinearity, whose phases according to (7.6.6) and (7.6.7) are changed in transit through elements with nonlinear amplitude of characteristics, will store/add up according to power.

Consequently, it is possible to draw conclusion that during transmission OBP power of products of nonlinearity, determined by members of polynomial with even degrees, is located as follows:

$$P_{н2,4} = P_2 + P_4. \qquad (7.6.12)$$

Substituting in this expression of value P_2 and P_4 those found from (7.2.33), we will obtain

$$P_{н2,4} = \frac{\Delta F_к \kappa_n^2}{\Delta F}\left[4e^{4p_{ср}} y_2(\xi) \sum_{i=1}^{m} e^{-2b_{ki}} + 192e^{8p_{ср}} y_4(\xi) \sum_{i=1}^{m} e^{-2b_{ki}}\right]. \qquad (7.6.13)$$

Total power of noises at output of telephone channel during transmission OBP will be determined by sum of thermal noises $P_ш$ and by power of nonlinear products, formed due to members of polynomial with even and odd degrees, i.e.,

$$P_Σ = P_ш + P_{н2,4} + P_{н3,5}. \qquad (7.6.14)$$

Here value $P_ш$ is determined by expression (7.2.6) or (7.2.8), and value $P_{н3,5}$ and $P_{н2,4}$ - respectively by expressions (7.6.11) and (7.6.13).

According to expression (7.6.14) can be solved inverse problem - in the case, if value $P_Σ$ is given one, after taking values $P_ш$, $P_{н2,4}$ and $P_{н2,5}$ equal to some values, according to expressions (7.2.6) or (7.2.8), (7.6.11) and (7.6.13) it is possible to determine approximate values of parameters of equipment. It is obvious that if

between the amplifiers will be included/switched on the frequency converters, which realize a transposition of the spectrum for the frequency, equal to Ω_τ, then can be obtained the expressions, analogous (7.6.6) and (7.6.7), but distinguishing in the arguments to values Ω_τ. This will lead to the fact that the components of nonlinear noises in the sections, divided by converters, will be summarized for the power. During inverse transformation to the initial group spectrum the addition of noises according to the law (7.6.14) will occur.

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Let us note that adopted assumptions can be examined only as first approximation and require supplementary investigations.

Transmission with ЧМ (ChM). With the passage through the circuit ЧМ (ChM) of signal at the output of telephone channel will be obtained the noise, which consist of the following three basic groups of noises: thermal noises, transient noises, determined by the nonlinearity of the amplitude characteristics of group circuit and transient noises, caused by the nonlinearity of the phase responses of those elements of the circuit, through which are passed ЧМ (ChM) of oscillation/vibration.

Taking into account that thermal noises and transient noises of both forms random processes, independent from each other, they will be summarized for power in accordance with (7.6.1).

Let us consider question about addition of transient noises, caused by nonlinearity of elements of group circuit and by nonlinearity of elements of circuit of hf/VCh and PCh. It is obvious that the total quantity of the power of the noises, caused by the nonlinearity of the characteristics of the group amplifiers and other elements of group circuit, can be counted according to the

expressions, given for OBP.

Addition of noises, caused by nonlinearity of phase responses of circuit PCh and shf/SVCh, will be different for even and odd degrees of polynomial.

Nonlinear noise, determined by members of polynomial of even orders, is determined by assymetric nature of characteristics of group time lag. This assymetric nature is not identical for different elements of equipment of shf/SVCh and PCh. Therefore it is possible to consider the addition of such noises according to the power as that permitted.

Passing to examination of noises, caused by members of polynomial of odd order, it is possible to draw conclusion that in different elements of circuit, in such cases, when there is no compensation for nonuniformity of characteristics of group time lag, can appear identical distortions of characteristics.

In this case transient noise, caused by nonlinearity of odd orders, will store/add up on voltage/stress.

In such cases, when in circuit PCh or shf/SVCh nonlinearity-correcting circuits of phase responses are utilized,

both overcompensation of nonlinearity and its undercompensation can occur. It is obvious that under these conditions for distortion they will carry not systematic character and therefore the addition of the nonlinear noises of odd order can be conducted according to the power.

In communication systems through ISZ, unconditionally, elements of correction of phase responses of separate amplifiers PCh and shf/SVCh will be introduced. From the aforesaid it follows that in such systems the noises, caused by the nonlinearity of all orders in the circuits of shf/SVCh and PCh, to admissibly summarize by the power.

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Transmission with КИМ (KIM). In the case KIM total noise at the output of telephone channel is determined by thermal noises and quantization noises. Taking into account the statistical independence of these noises, it is possible to draw the conclusion that the addition of them must be realized according to the power. Consequently, for this case $P_{\Sigma} = P_{\text{ш.а}} + P_{\text{ш.кв}}$.

Here power of thermal noises can be determined according to expressions (7.4.1)-(7.4.4) for two sections of line of communications in the absence of pulse regeneration on board or for one section, if pulse regeneration is realized. The power of

quantization noises is determined according to expression (7.4.10).

7.7. Comparison of different methods of the transmission of telephone signals according to the energy indices.

In order to have representation about fundamental energy relationships/ratios, inherent in different methods of transmission of telephone signals, let us lead calculation of power of onboard and terrestrial transmitters with values of parameters, given in Table 7.7.1.

Let us note that computed values accepted in table of amplification of onboard antennas must be less than their maximum amplification due to inaccuracy in orientation relative to direction to terrestrial stations.

Weakening signal with passage from that transmitting to receiving antenna can be found from expression (3.2.5). During calculations let us take the worst case - sediments/residues with an intensity of 1 mm/h occur on both terrestrial points/items. Then from the graph/curve Fig. 3.2.5 we find losses due to wet protective hood of antenna $\beta_{ox}=2{,}0$ dB. According to Fig. 3.2.3 we have: at the frequency of 4 GHz value $\delta_0 \cong 10^{-3}$ dB/km, and at the frequency of 6 GHz - $\delta_0 \cong 2 \cdot 10^{-3}$ dB/km. During calculations we will count the thickness of

rain clouds l=3 km. on curved A, given in Fig. 3.2.1, at the frequency of 4 GHz we will obtain $B'_{0a}(90)=0{,}03$ dB, and at the frequency of 6 GHz — $B'_{0a}(90)=0{,}035$ dB. From Table 3.2.1 for the angle of elevation of $\beta=10°$ we find $\alpha=5.5$. Keeping in mind the enumerated values after their substitution into expression (3.2.5) and computations, we will obtain:

- weakening signal in the section satellite - Earth

$$B_{02}=199 \text{ dB}; \qquad (7.7.1)$$

- weakening signal in the section the Earth - the satellite

$$B_{01}=202{,}5 \text{ dB}. \qquad (7.7.2)$$

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Table 7.7.1.

№ пп. (a)	Параметр (b)	Величина (c)
1	Высота спутника над поверхностью Земли	$H = 36\,000$ км
2	Угол возвышения	$\beta = 10'$
3	Протяженность участка спутник—Земля или Земля—спутник при $\beta = 10'$ в соответствии с (2.2.4)	$L = 40\,600$ км
4	Частота колебаний (длина волны) на участке Земля—спутник	6000 Мгц (4a) (5,0 см)
5	Частота колебаний (длина волны) на участке спутник—Земля	4000 Мгц (4a) (7,5 см)
6	Диаметр земных антенн	25 м
7	Максимальное усиление земной передающей антенны	61 дб (7a)
8	Максимальное усиление земной приемной антенны	58 дб (7a)
9	Расчетное усиление бортовой передающей антенны	14 дб (7a)
10	Расчетное усиление бортовой приемной антенны	17 дб (7a)
11	Кпд фидеров передатчиков	$\eta_{Фпб} = \eta_{Фп} = 0,7$
12	Кпд фидера земного приемника	$\eta_{Фпр} = 0,9$
13	Кпд фидера бортового приемника	$\eta_{Фпрб} = 0,7$
14	Потери в бортовых фильтрах приема и передачи	$b_{Фпрб} = b_{Фпб} = 0,2$ дб (7a)
15	Потери в земных фильтрах приема и передачи	$b_{Фп} = b_{Фпр} = 0,1$ дб (7a)
16	Потери за счет эллиптичности поляризации	— отсутствуют (16a)
17	Шумовая температура земного приемника	$T = 50°$K
18	Шумовая температура бортового приемника	$T = 700°$K
19	Суммарное значение псофометрически взвешенной мощности шумов на выходе любого телефонного канала в точке с нулевым относительным уровнем при усреднении за час	10 000 пвт (19a)
20	Из них тепловые шумы на участке спутник—Земля	6000 пвт (19a)
21	Из них тепловые шумы на участке Земля—спутник	1000 пвт (19a)
22	Из них тепловые шумы в случае КИМ с регенерацией импульсов на спутнике на участке спутник—Земля или Земля—спутник	9000 пвт (19a)
23	Число каналов	$N = 600$
24	Среднее значение мощности многоканального сигнала в соответствии с (6.2.4)	$p_{ср} = 12,8$ дбм (24a) (18 мвт) (24b)
25	Пикфактор многоканального сигнала (рис. 6.2.5)	$\varkappa = 13$ дб (7a)
26	Частота верхнего канала	$F_в = 2540$ кгц (26a)
27	Выигрыш в верхнем канале за счет предыскажений	$W = 3$ дб (7a)

Key: (a). No pp. (b). Parameter. (c). Value. (1). Height/altitude of satellite above surface of Earth. (2). Angle of elevation. (3). Extent of section satellite - Earth or Earth - satellite at $\beta=10°$ in accordance with (2.2.4). (4). Frequency (wavelength) in section Earth - satellite. (4a). MHz. (5). Frequency (wavelength) in section satellite - Earth. (6). Diameter of terrestrial antennas. (7). Maximum amplification terrestrial of transmitting antenna. (7a). dB. (8). Maximum amplification of terrestrial receiving antenna. (9). Calculated amplification of onboard transmitting antenna. (10). Calculated amplification of onboard receiving antenna. (11). Efficiency of feeders of transmitters. (12). Efficiency of feeder of terrestrial receiver. (13). Efficiency of feeder of onboard receiver. (14). Losses in onboard filters of reception and transmission. (15). Losses in ground filters of reception and transmission. (16). Losses due to ellipticity of polarization. (16a). they are absent. (17). Noise temperature of terrestrial receiver. (18). Noise temperature of onboard receiver. (19). Total value of psophometrically weighted power of noises at output of any telephone channel at point with zero relative level with averaging per hour. (19a). pW. (20). From them thermal noises in section satellite - Earth. (21). From them thermal noises in section Earth - satellite. (22). From them thermal noises in the case KIM with pulse regeneration on satellite in section satellite - Earth or Earth - satellite. (23). Number of channels. (24). Average/mean power coefficient of multichannel signal in

accordance with (6.2.4), (24a). dbm. (24b). mW. (25). Peak factor of multichannel signal (Fig. 6.2.5). (26). Frequency of upper channel. (26a). kHz. (27). Gain in upper channel due to predistortions. Page 216.

Total weakening of signal between transmitter and receiver for each section of connection/communication can be obtained on the basis of expression (3.1.4). Taking into account data of Table 7.7.1. and the preceding/previous expressions, we find the total weakenings of the sections of the connection/communication:

- for the section satellite - Earth

$$B_{0z2} = 129{,}3 \text{ dB}; \quad (7.7.3)$$

- for the section the Earth - the satellite

$$B_{0z1} = 127{,}8 \text{ dB}; \quad (7.7.4)$$

- without protective hood of weakening will comprise:

$$B_{10z2} = 127{,}2 \text{ dB } (5 \cdot 3 \cdot 10^{12} \text{ once}); \quad (7.7.5)$$

$$E_{0z1} = 125{,}7 \text{ dB } (3 \cdot 7 \cdot 10^{12} \text{ once}); \quad (7.7.6)$$

Let us pass to the calculation of the total noise temperature of receivers. The noise temperature of filters can be found after the determination of transmission factor. It is obvious that the transmission factor M and weakening b filters are connected with

relationship/ratio b=1/M. Therefore in accordance with (4.2.23) and the data of Table 7.7.1 we will obtain:

- the noise temperature of the onboard filters

$$T_{\Phi б} = 15°K; \qquad (7.7.7)$$

- the noise temperature of the filters of the terrestrial equipment

$$T_{\Phi} = 9°K. \qquad (7.7.8)$$

Noise temperature of feeders can be designed analogously. The noise temperature, created by protective hood of the antenna of ground-based receiver, can be determined according to the graphs/curves, given in Fig. 4.4.2. Assuming/setting $\varkappa$=0,1 mm, we will obtain

$$T_{эк} = 30°K. \qquad (7.7.9)$$

Temperature, caused by cosmic noises at frequencies of 4 kHz and 6 GHz, in accordance with Fig. 4.3.2, comprises less than 1°K and can be disregarded. The temperature of the atmosphere at these frequencies according to Fig. 4.3.2 at β=10° will comprise

$$T_{а} = 15°K. \qquad (7.7.10)$$

During calculation of temperature of onboard receiver we will consider that temperature of Earth comprises

$$T_{з} = 290°K. \qquad (7.7.11)$$

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Since working amplification of onboard receiving antenna is 17 dB (50 times), we will consider that maximum amplification is 19 dB

(80 times); then according to (3.3.16) width of radiation pattern according to half power will be

$$\theta = 1/\frac{24000}{80} = 18^\circ. \qquad (7.7.12)$$

Taking into account that according to Fig. 2.2.2 angle, at which Earth is visible from satellite

$$2\gamma = 17^\circ, \qquad (7.7.13)$$

in accordance with (4.3.14) after transition/junction to solid angles let us find that noise temperature of Earth, in reference to antenna of satellite, will comprise

$$T_з = 186^\circ K. \qquad (7.7.14)$$

Keeping in mind given values and data of Table 7.7.1, according to (4.1.1) let us find total noise of temperatures, in reference to input of receivers.

For terrestrial receiver with protective hood

$$T_{з1} = 134^\circ K. \qquad (7.7.15)$$

For terrestrial receiver without protective hood

$$T_{з2} \cong 100^\circ K. \qquad (7.7.16)$$

For onboard receiver

$$T_{з6} \cong 1000^\circ K. \qquad (7.7.17)$$

Comparing (7.7.3) with (7.7.5) and (7.7.15) with (7.7.16), it is possible to see that installation of radome will lead to considerable increases in power of onboard transmitter. Therefore subsequently let us take, that protective hood is absent.

Calculation with ОБП (OBP), can be carried out in accordance with expression (7.2.9). Substituting in this expression of the values, determined by Table 7.7.1 and expressions (7.7.5) and (7.7.6), (7.7.16) and (7.7.17) we will obtain:

$$P_{\text{пб}} = 40 \text{ W}, \qquad (7.7.18)$$

$$P_{\text{пз}} = 1700 \text{ W}, \qquad (7.7.19)$$

The peak values of the power of transmitter according to (7.2.10) and Fig. 6.2.5 with ϵ=0.001 must be:

$$P_{\text{пбпик}} \cong 400 \text{ W}, \qquad (7.7.20)$$

$$P_{\text{пзпик}} = 17 \text{ kW}. \qquad (7.7.21)$$

Bandwidth, necessary for transmission of 600 telephone conversations, is determined from Table 6.2.1 and is

$$\Delta F = 2480 \text{ kHz}. \qquad (7.7.22)$$

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Calculation with ЧМ (ChM) can be carried out according to expression (7.3.5). During calculations should be utilized Fig. 7.3.6 and data, given in Table 7.7.1, and expression (7.7.5), (7.7.6),

(7.7.16) and (7.7.17). In the case of predistortions (equal to 3 dB for upper telephone channel) and in the absence of negative feedback, the power of onboard transmitter must be not less

$$P_{\text{пб}} = 15 \text{ W}. \qquad (7.7.23)$$

During application of negative feedback in frequency in accordance with (5.3.29) supplementary gain is obtained. Therefore the power of onboard transmitter can be lowered to the value

$$P_{\text{пб}} = 6 \text{ W}. \qquad (7.7.24)$$

Power of terrestrial transmitter

$$P_{\text{пз}} = 1200 \text{ W}. \qquad (7.7.25)$$

In the case of applying OSCh this power can be lowered to 450 W.

Bandwidth, occupied by transmission of 600 telephone conversations in designed parameters of systems, is determined according to Fig. 7.3.6 and expression (7.3.7):

$$\Delta f = 40 \text{ MHz}. \qquad (7.7.26)$$

Utilizing obtained values, one should ascertain that relation of power of signal and noise exceeds threshold value.

Calculation with КИМ (KIM) is performed accordingly expressions

(7.4.1)-(7.4.4). With N=600 and a number of pulses/momenta in word n=5 width of band

$$\Delta f = 50 \text{ MHz}. \qquad (7.7.27)$$

For case, when pulse regeneration on satellite is not used, in accordance with expressions indicated during use of data of Table 7.7.1 and dotted curves, Fig. 7.4.2, let us find:

with KIM-AM

$$P_{\text{пб}} = 18 \text{ W}, \qquad (7.7.28)$$

$$P_{\text{пз}} = 200 \text{ W}; \qquad (7.7.29)$$

with KIM-FM

$$P_{\text{пб}} = 9{,}5 \text{ W}, \qquad (7.7.30)$$

$$P_{\text{пз}} = 115 \text{ W}; \qquad (7.7.31)$$

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Calculations related to case, when KIM is realized by signals of subscribers, i.e., successive transmission of signals of subscribers is utilized. During the use of equipment for the frequency division multiplex of channels with the subsequent conversion of the group spectrum in KIM, in accordance with expressions (6.4.3) and (6.4.5), occurs an increase of the frequency band 1.05 times. Consequently,

with this method of construction of equipment for multiplexing 1.05 times the power of each transmitter will increase.

It must be noted that application of pulse regeneration on board satellite will lead to reduction in power of transmitters, which will be more considerable for terrestrial transmitter, since according to Table 7.7.1 power capacity of noises in section Earth - satellite in pulse regeneration can be considerably increased.

Calculations, given in accordance with data of Table 7.7.1 show that in pulse regeneration in the case to KIM-AN they will be required power:

$$P_{\text{пб}} = 17 \text{ W},$$

$$P_{\text{пз}} = 85 \text{ W}.$$

Comparison of these values with values, determined (7.7.28) and (7.7.29), shows that introduction of very complicated installed equipment, which ensures pulse regeneration on satellite, does not lead to considerable reduction in power of transmitters.

Conclusions/outputs. On the basis of the comparison of the power of the onboard and terrestrial transmitters, designed for different cases of modulation of oscillations, it is possible to do the

following conclusions:

With ChM and KIM-AM are necessary the approximately identical frequency bands, and the power of onboard transmitters in the case KIM prove to be somewhat greater. During the use KIM and the successive transmission of telephone conversations is required the somewhat smaller power of transmitters, than in the case of the preliminary frequency division multiplex of telephone channels. However, this change in the power of transmitters is so unessential (approximately 1.05 times), that there must not be determining.

Application of OBP with frequency division multiplex of telephone channels leads to essential reduction of required band of frequencies (more than ten times). However, in this case are necessary considerably more powerful transmitters than in the case ChM or KIM. It is necessary to also have in mind that in the case OBP the transmitters must have very large linearity of amplitude characteristic; this will lead to a considerable increase in the rating of tubes and their poor use. Therefore the application of OBP is expedient only in the section the Earth - satellite and only in such cases, when the sharp reduction of the occupied frequency band is the decisive factor.

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7.8. Effect of the Doppler effect.

In Section 3.5 it was noted that in the various forms of modulation Doppler effect will lead to change not only in carrier frequency adopted, but also change in frequencies of transmitted signal (to strain of spectrum of signal) in $1+v/c_0$ times. Consequently, at the output of detector frequency shift F, which is located in the group spectrum with cut-off frequencies of F_1 and F_2 ($F_1 \leq F \leq F_2$), will comprise

$$\delta = F\frac{v}{c_0}. \qquad (7.8.1)$$

Since for normal work of equipment for frequency division multiplexing frequency spread at points of transmission and reception must not exceed 2 Hz, in accordance with (7.8.1), we will obtain condition, which determines maximum frequency of group spectrum

$$F_2 \leqslant 2{,}0 \cdot 10^5 \text{ Hz}. \qquad (7.8.2)$$

Thus, Doppler effect will unavoidably impose specific limitations on number of transmitted channels during construction of equipment of multiplexing accepted. At the same time the Doppler effect will limit the possibilities of the secondary multiplexing of telephone channels by telegraph and phototelegraphic signals.

For correction of change in frequencies, caused by Doppler effect, introduction to automatic frequency correction of generator

equipment, utilized in equipment for frequency division multiplex, is possible. However, this path requires the considerable treatment/processing of the existing equipment.

Another possible method of frequency drift compensation consists in the fact that at terrestrial stations in circuit of signal device/equipment with adjustable time lag switched on. The corresponding adjustment of time lag can be carried out per program, which considers changes in the coordinates of satellite, or automatic on the pilot signal, transmitted from the terrestrial station to the satellite and back.

Let us note that in communication systems with use of stationary ISZ Doppler effect is absent. In connection with this in the systems with stationary ISZ is simplified equipment and the operation.
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7.9. Effect of time lag and echo signals.

Large extent of line of communications between terrestrial stations and repeater, which is located on board ISZ, leads to delay (time lag) of signals, since for passage of distance L(m) to signal is required time

$$t = \frac{L}{c_0} \cong \frac{2H}{c_0}, \qquad (7.9.1)$$

where $c_0=3\cdot10^8$ m/s - speed of light, and H - distance from satellite to surface of Earth. Hence follows that during the motion of satellite along the orbits with a height/altitude of 36000 km, and consequently, and in the case of stationary satellite, the value of time lag will be about 250 ms. With the duplex telephone conversations signal lag leads to the appearance of the forced pauses in the conversation between the subscribers, impedes the supply of replicas and questions, which breaks the naturalness of the conversation of subscribers. In connection with this MKKTT [MKKTT - International Telegraph and Telephone Consultative Committee] he recommends so that the maximum time lag of signal (group time of the emission of the signal) would not exceed 250 ms.

On extended lines of communications at the same time will occur time lag of echo-signals, which can arise upon transfer from four-wire circuits to twin-lead (see Fig. 1.1.5).

Echo-signals will be developed in the form of crosstalk by subscriber of their conversation (signal), delayed to period $t_э$ equal to doubled time of emission of the signal between subscribers. Thus, taking into account (7.9.1)

$$t_э = \frac{2L}{c_0} \cong \frac{4H}{c_0}. \qquad (7.9.2)$$

Consequently, in the case of communication systems, which use satellites, which move along orbits with height/altitude about 36000 km, or stationary satellites value $t_0 \cong 500$ ms. With the long time lag the echo-signals become perceptible and they interfere with the calls between the subscribers.

Results of investigations of the distinctness of echo signals, carried out for Japanese speech by method of subjective-statistical appraisal/review, are given in Fig. 7.9.1 [7.15]. The curve A determines the relationship/ratio between the delay time and this average/mean attenuation of echo-signal with respect to the fundamental signal, with which the echo is noticeable, curve B corresponds to the attenuation, with which echo level is permitted. The shaded zones determine the spread of readings/indications of subjective appraisal/review. Using curve B, taking into account the scatter of readings/indications, it is possible to determine the necessary value of the average/mean attenuation of echo-signals $b_э$ (dB) on the line of communications.

For communication system, which uses stationary ISZ, in accordance with Fig. 7.9.1, it is necessary to ensure attenuation of echo-signals to value, equal to approximately 60 dB.

With work of communication systems, which use moving/driving ISZ, it will have value of inconstancy of time lag.

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For combatting echo-signals in subscribers echo suppressors, which for echo-signals introduce necessary attenuation, must be established/installed. According to the operating principle the echo suppressors can be decomposed into two groups: momentary effect and constant action. The specific attenuation in the very short period introduces the echo suppressor of momentary effect into return circuit of signal. The echo suppressor of constant action introduces into return circuit of signal the attenuation, which gradually increases from zero and which reaches or exceeds the value, sufficient for the suppression of echo-signal.

Schematic of echo suppressor of momentary effect is shown in Fig. 7.9.2. From the diagram it follows that the echo suppressor consists of the amplifier y, connected in parallel to the circuit of reception/procedure, and the flip-flop circuit Опр (Opr), during functioning of which into the circuit of transmission is introduced attenuation b. In the absence of speaking currents the circuit of transmission is connected to the differential system without the supplementary attenuation. Amplifier y is regulated so, in order to

in the absence of speaking currents the trap/barrier did not operate from noises in the circuit of reception/procedure, but it would operate/wear only from the speech input currents of amplifier *y*.

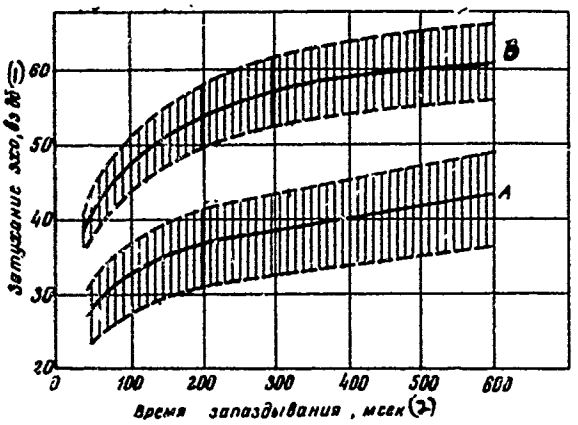

Fig. 7.9.1. Necessary weakening of echo-signals depending on time lag.

Key: (1). Echo attenuation,... dB. (2). Time lag, ms.

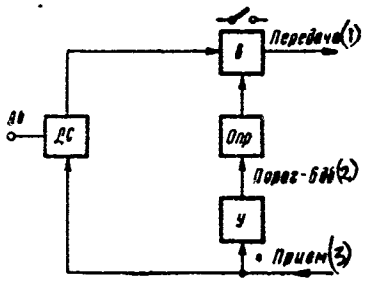

Fig. 7.9.2. Schematic of echo suppressor.

Key: (1). Transmission. (2). Threshold- 6dB. (3). Reception.

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Essential deficiency in this echo suppressor - false response with increase in noise level in receiving circuit. An increase in the protection from bag with reduction in the sensitivity of trap/barrier, leads to decrease in the articulation, since are caused the cutting of separate sounds or "swallowing" of sounds and

syllables.

Somewhat best schematic of echo suppressor is shown in Fig. 7.9.3. In this diagram in the absence of speaking currents the circuit of reception/procedure is included/switched on, and the circuit of transmission is included/switched on. Upon the appearance of speaking currents of this subscriber the flip-flop circuit includes the circuit of transmission and turns off/disconnects the circuit of reception/procedure. In this case, since in the circuit of transmission the noise level is less than in the circuit of reception/procedure, this diagram possesses the best protection from the noise, than preceding/previous.

Deficiency in diagram in Fig. 7.9.3 in the fact that speaking currents from circuit of reception/procedure through differential system can fall into circuit of transmission. This can cause functioning of trap/barrier and lead to a break in the circuit of reception, i.e., to interruption of conversation. This phenomenon due to the large dynamic range of speech can be repeated periodically and is called the phenomenon of "flutter". For eliminating "flutter" are utilized the more compound circuits of echo suppressors, which include the elements of blocking.

Are interesting echo suppressors with use of distinguishing

device/equipment (РУ), which makes it possible to distinguish voice signals and noises. The construction of echo suppressors РУ is shown in Fig. 7.9.4. In this diagram the distinguishing device/equipment operates/wears only from the speaking currents and does not react to the noises. In the presence in circuit of the reception/procedure of speaking currents with the help of РУ and the controlled amplifier УУ the circuit of transmission is cut off. Let us note that in this schematic of echo suppressors the interruption of subscriber (supply of replicas) is feasible only in the pauses between the words of that speaking, if these pauses exceed approximately 100 ms. This diagram does not allow/assume simultaneous contrary conversation.

Let us pause at principle of operation РУ. It is based on what if the spectrum of voice signal is divided into m of equal frequency bands, then the instantaneous power of voice signal, at least, in two of these bands are not equal to each other, since the spectral density of speech is not identical. On the contrary, fluctuation noise has uniform spectral density; therefore the power of fluctuation noise in the equal frequency bands will be identical.

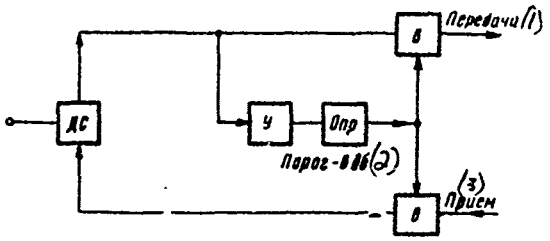

Fig. 7.9.3. Schematic of improved echo suppressor.

Key: (1). Transmission. (2). Threshold- 6 dB. (3). Reception/procedure.

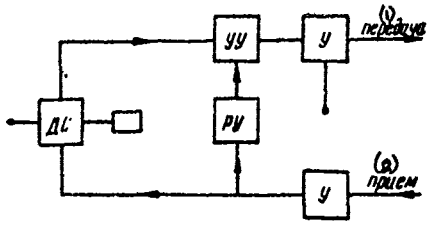

Fig. 7.9.4. Schematic of echo suppressor with distinguishing device/equipment.

Key: (1). Transmission. (2). Reception.

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Thus, after dividing oscillation spectrum by filters on several identical frequency bands and after comparing between themselves output potentials of square law detectors, switched on after these filters, it is possible to distinguish voice signals from fluctuation noises.

Principles of construction of echo suppressors examined make it possible to formulate following requirements for them:

- subscribers must not note work of echo suppressor,

- sensitivity of echo suppressor must be such that would not cut themselves beginnings of syllables and not "were swallowed" separate sounds and syllables,

- echo suppressors must not operate/wear from noises.

7.10. Switching satellites.

In systems, which use moving/driving ISZ for guaranteeing long continuous bond, it is necessary to realize switching antennas of ground stations from one communication satellite to another, which is clarified by Fig. 1.1.3. One should note that at the moment of switching an abrupt change of the value of signal lag in the time can occur, since the extent of routes with the work through different satellites can prove to be different.

Calculations show that disagreement in time lag will hardly exceed 40 ms in limits of sufficiently wide altitude range of orbits.

Disagreement in time lag to 20 ms must not cause disturbances/breakdowns during transmission of telephone conversations, since these switchings are sufficiently rare. However, in the case of the secondary multiplexing of telephone channel by the telegraph signals of switching they will lead to the errors. With the

work of wirephoto the disagreement in the time lag in 20 ms will cause considerable distortions.

Aforesaid pertains to communication systems, which use frequency division multiplex of channels and ChM. During the transmission of television program the disagreement in the time can lead to the disruption/separation of synchronization and, consequently, to the disturbances/breakdowns of transmission. The disturbances/breakdowns of transmission and appearance of errors during the switching it will occur in the case of applying KIM.

Let us note that disagreement in time can be compensated by special devices/equipment, which control time lag in circuit. These devices/equipment must be included in the circuit of the reception/procedure of terrestrial station.

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Chapter 8.

TRANSMISSION OF TV SIGNALS OVER COMMUNICATION LINKS.

8.1. Electrical characteristics of channels.

Electrical characteristics and norms, established/installed for communication channels, by which must be transmitted video signals, are determined by recommendations of MKKR [International Radio Consultative Committee]. As the hypothetical standard circuit during the transmission of the signals of black and white television through ISZ was accepted the communications link Earth-satellite-Earth (see Fig. 7.1.1, [8.1, 8.2]), which includes one pair of modulators and demodulators.

In connection with the fact that expenditures for installation and operation of communication systems through ISZ depend substantially on necessary band of video spectrum, in recommendations of MKKR nominal upper limit of band of video signal is established/installed equal to $F_в$=5 MHz to the same value of 5 MHz it is accepted equal equivalent noise bandwidth.

One should note that corrected value of width of band of video spectrum relates to transmission of signals of black and white image. Width of video spectrum, necessary for transmission of colored of

television of image, in recommendations of MKKR has not yet been determined. In this connection it is possible to note that during the transmission of colored television image along the system SEKAM-3 between Moscow and Paris through the communication satellite Molniya-1 was obtained the completely satisfactory quality of colored transmissions. The band of the video frequencies, transmitted through ISZ Molniya-1 in the regime/conditions of television it was effectively 5 MHz (see Section 10.5 and 11.3 and Table 11.4.1).

Taking into account that norm to signal-to-noise ratio in intercontinental airlines of communications through ISZ must be compared with analogous norm, accepted for switched continental circuits of MKKR, it was recommended:

- Ratio of signal to continuous fluctuation weighed noise at output of standard hypothetical circuit must comprise, at least, 55 dB for 99% of time.

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In this case ratio of signal to continuous fluctuation noise is defined as ratio (in dB) of spread/scope of picture signal without synchronizing pulses to effective noise voltage in band from 10 kHz to upper nominal band edge of video frequencies F_2. The limitation of band from the side of low frequencies makes it possible to exclude the background of power supplies and microphonic noises from the measurements.

In measurement of signal-to-noise ratio is implied application of instrument with "effective time constant" or "time of integration" for power, equal to 1 s, and also filter of lower frequencies and weighing circuit, whose diagrams are given respectively in Fig. 8.1.1 and 8.1.2. Data of elements of filter of low frequencies and its attenuation are cited in Tables 8.1.1 and 8.1.2, respectively [8.3, 8.4] in the form of dependences on higher frequency of video signal.

Let us pause now at other requirements for channel, intended for transmission of video signals.

Overall line attenuation of hypothetical standard circuit during transmission of TV signals according to recommendation of MKKR must be in limits 0±1 of dB.

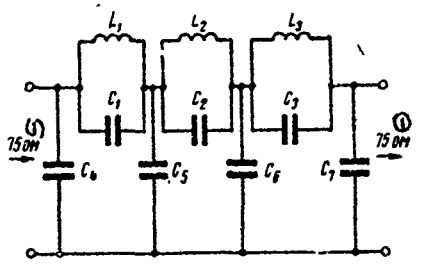

Fig. 8.1.1.

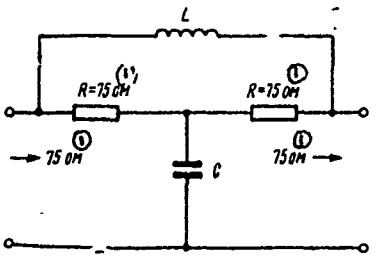

Fig. 8.1.2.

Fig. 8.1.1. Schematic of the filter of low frequencies.

Key: (1). ohm.

Fig. 8.1.2. Diagram of weighing circuit: $L=75\tau$, μH, $C=(\tau/15)10^{4}$, pF; for zone television with resolution into 625 rows, $\tau=0.33$ μs.

Key: (1). ohm.

Table 8.1.1.

(1) Порядковый номер элемента схемы	(2) Индуктивность *мкгн*	(3) Емкость *пф*	(4) Частота резонанса *Мец*
1	$14,38/F_2$	$497,6/F_2$	$1,8816/F_2$
2	$7,673/F_2$	$2723/F_2$	$1,1011/F_2$
3	$8,600/F_2$	$1950/F_2$	$1,2290/F_2$
4		$2139/F_2$	
5		$2815/F_2$	
6		$2315/F_2$	
7		$1297/F_2$	

Value F_2 this table gives in MHz.

Key: (1). The reference number of the element of network. (2). Inductance μH. (3). Capacity/capacitance pF. (4). Frequency of resonance MHz.

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Checking frequency and phase responses of circuit for compliance to demands of MKKR is realized with the aid of test signals Nos. 1 and 2 [8.2, 8.4] (see Fig. 8.1.3 and 8.1.4).

Test signal No 1 is sequence of square pulses with duration of 10 ms (at frame frequency, equal to 50 Hz), of following with porosity, equal to one. these pulses/momenta have spread/scope, equal to spread/scope of video signal, and transmit simultaneously with synchronizing and extinguishing pulses/momenta. With measurement on the oscilloscope face is superimposed the transparent stencil, on which it is plotted/applied to Fig. 8.1.3. In transit through the circuit of test signal No.1 image on the oscilloscope face must be established so that it would not exceed the limits of Fig. 8.1.3,

plotted/applied on the transparent stencil, where the points which correspond to half the spread of the pulse coincide with points M_1 and M_2, while the middles of the horizontal sections coincided with points A and Б, respectively. In this case the shape of the signal at the output should be within the limits of the cross-hatched part of the stencil (Fig. 8.1.3). Thus, the slope of the horizontal part of the test signal should not exceed +10%.

Table 8.1.2.

F/F_s	b_0 (1) (дБ)	F/F_s	b_0 (1) (дБ)
0,98	0,1	1,04	14,8
0,99	0,5	1,05	18,8
1,00	1,8	1,06	23,0
1,01	4,2	1,07	27,7
1,02	7,3	1,08	33,3
1,03	10,9	1,09	41,0

Key: (1). dB.

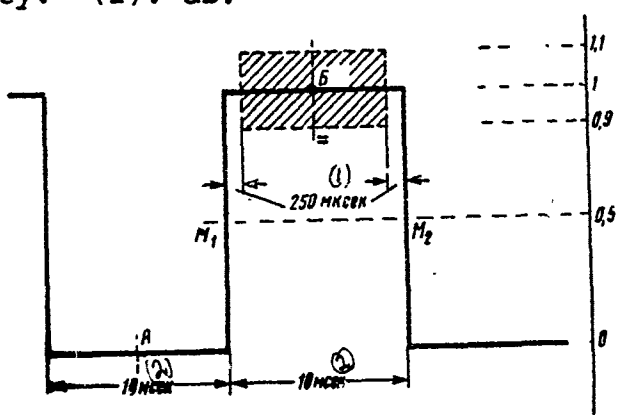

Fig. 8.1.3. Test signal No. 1 and stencil (it is shaded).

Key: (1). 250 μs. (2). 10 ms.

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During transmission through systems with number of rows, equal to 625 test signals No. 2, shown in Fig. 8.1.4a, pulse apex, reproduced on output of line, must be within the limits of shaded part of image on stencil, i.e., inclination/slope of horizontal part must not be more than ±5% of spread/scope of picture signal (see Fig. 8.1.4b). Image on the oscilloscope face must be established analogously of said in the description Fig. 8.1.3. Test signal No. 2 is utilized also for checking the transient response in the region of higher frequencies.

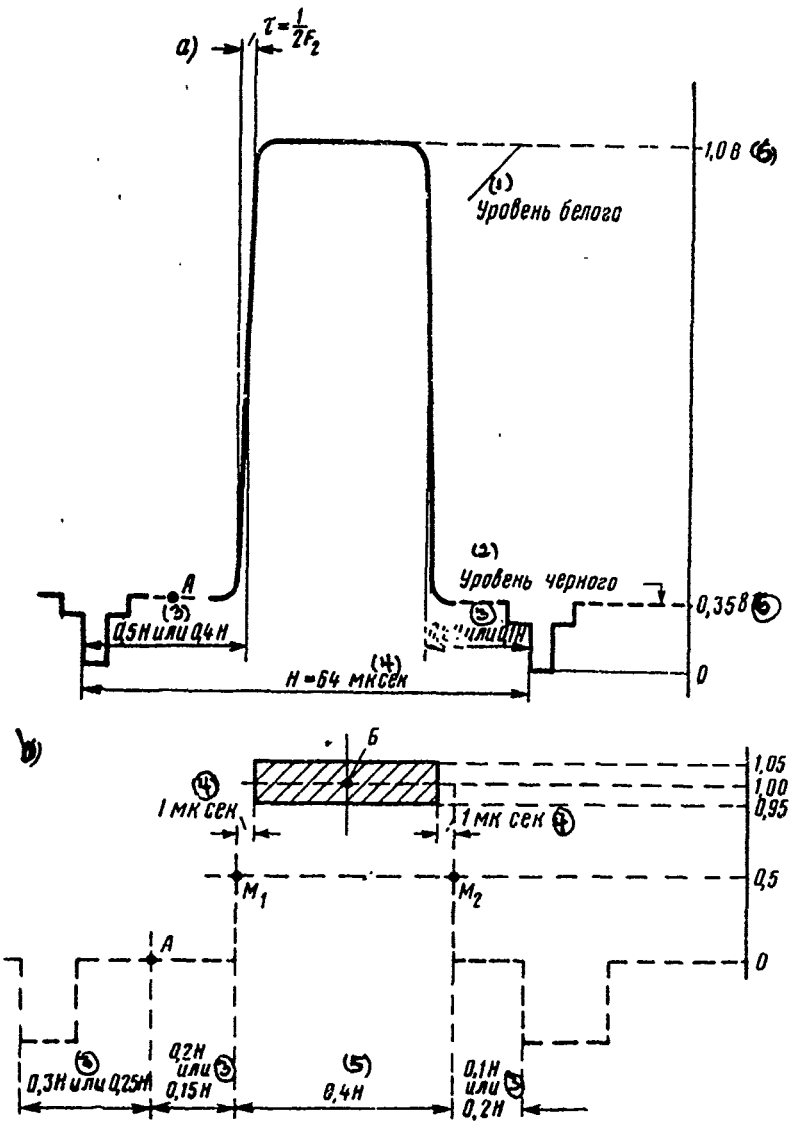

Fig. 8.1.4. a) test signal No. 2, b) stencil (it is shaded). Key: (1). White level. (2). Black level. (3). N or N. (4). μs. (5). 0.4 N. (6). V.

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In this case rise time τ of the signal (see Fig. 8.1.4a) for the systems with a number of rows, equal to 625, must comprise: in the

case of F_2=5 of MHz τ=0.10 μs and with F_2=6 MHz τ=0.08 μs

At the output of the circuit, designed for the transmission of TV signal with the resolution into 625 rows and a higher frequency of 5 MHz, the front of the observed pulse/momentum must not exceed the limits of image on the stencil, which is given in Fig. 8.1.5. In this case the pulse/momentum on the oscillograph is established/installed so that the middle of its front would coincide with point M (Fig. 8.1.5), and the sections of pulse/momentum, which correspond to the levels of black and white, they would coincide with the sections Б and A in Fig. 8.1.5 respectively. If pulse/momentum in the sections Б and A has overshoots, then their peak values must be established/installed symmetrically relative to Б and A. As already mentioned, Fig. 8.1.5 was intended for the systems with a number of rows 625 with the value of F_2=5 MHz. With a number of rows 625 and value F_2=6 MHz to the stencil must be applied the image, given in Fig. 8.1.6.

An number of cases, except given norms for distortion of pulses/momenta, amplitude-frequency characteristic (AChKh) and characteristic, which determines permissible change in time lag $\Delta\tau$, can be useful. Stencils for such characteristics, recommended by MKKR, are given in Fig. 8.1.7.

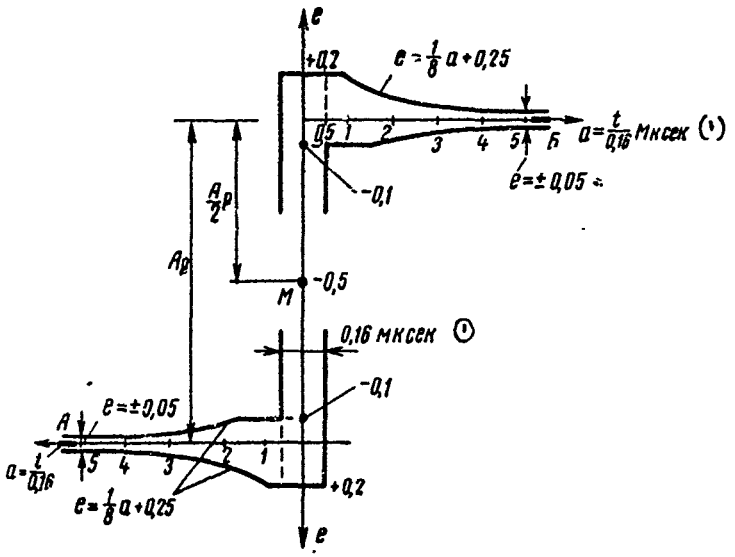

Fig. 8.1.5. Stencil of transient response with F_2=5 MHz.

Key: (1). μs.

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Determination of nonlinear distortions, introduced by circuit during transmission of interval of row, is realized by test signal No. 3, Fig. 8.1.8a. Of this signal the section of the image of each fourth row consists of the sine wave with a frequency of 0.2 F_2 (F_2 - the maximum frequency of video signal) with a spread/scope of 0.1 V, superimposed to the signal of saw-tooth for.m. In the measurements of three intermediate row in the transmitting end/lead they establish/install at the level of black or at the level white.

Nonlinear distortions are determined after isolation/liberation of sine wave at point of reception/procedure with the aid of band-pass filter. In the presence of nonlinear distortions the envelope of sine

wave will take the form, shown in Fig. 8.1.8b. Nonlinear distortions are computed according to expression $U = \left(1 - \frac{m}{M}\right) 100\%$, where m and M - values of the ordinates of signal amplitude envelope (see Fig. 8.1.8b), and according to the recommendation of MKKR must not exceed 20%.

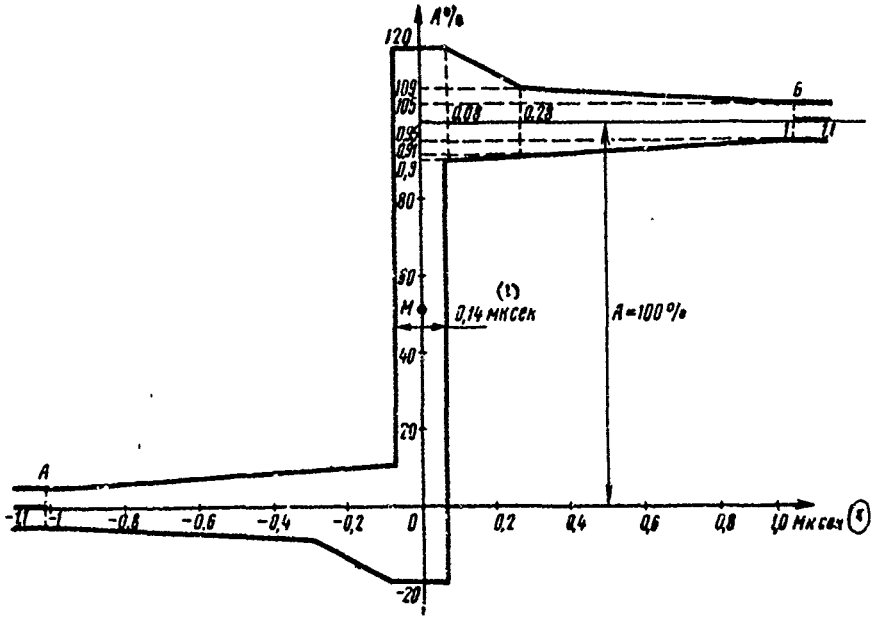

Fig. 8.1.6. Stencil of transient response with F_2=6 MHz.

Key: (1). μs.

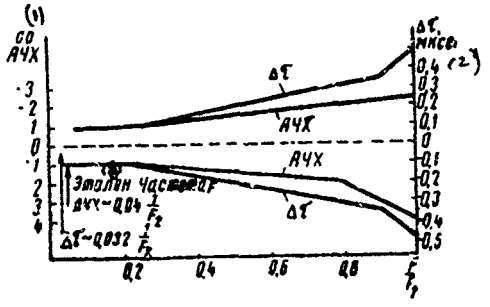

Fig. 8.1.7. Stencil of amplitude-frequency characteristic (attenuation) and characteristic of group time lag.

Key: (1). to AChKh. (2). Δτ, μs. (3). Frequency standard.

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Let us note that sine-quadratic pulses are very convenient form of test signal. These pulses have a form, described by equation $U(t)=\sin^2 2\pi F_н t$, shown in Fig. 8.1.9a. The spectral characteristic of

this pulse/momentum (Fig. 8.1.9b) is described by the expression

$$A(F) = \frac{1}{\sqrt{2\pi}} \sin \frac{F}{4F_н} \left[\frac{1}{F} + \frac{F}{4F_н^2 - F^2} \right]. \qquad (8.1.1)$$

From Fig. 8.1.9b it follows that sine-quadratic pulse has sufficiently sharply limited spectrum, that advantage of such pulses/momenta of the fact that in sufficiently simple electronic circuits they are formed with high accuracy, facilitates visual evaluation/estimate of its distortions.

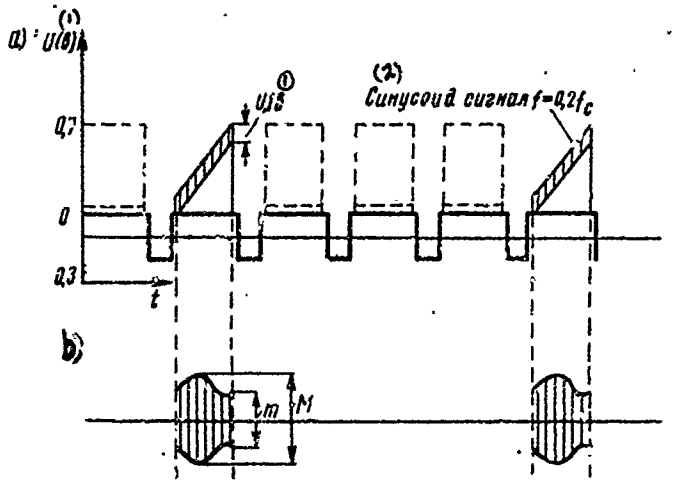

Fig. 8.1.8. a) test signal No. 3, b) envelope of sine wave.

Key: (1). V. (2). Sinusoids signal.

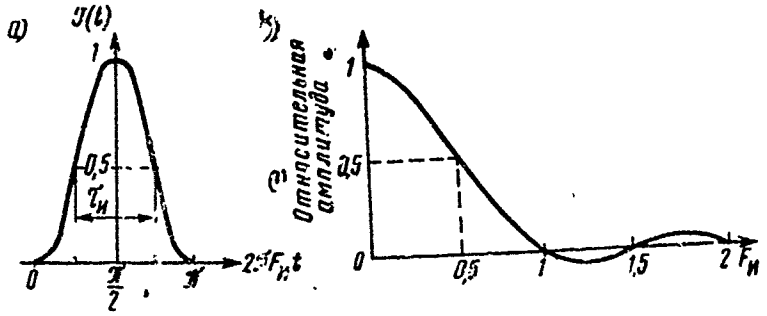

Fig. 8.1.9. a) sine quadratic pulse/momentum, b) spectral characteristic.

Key: (1). Relative amplitude.

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With passage of sine-quadratic pulses along communications link with insufficient bandwidth their relative height/altitude (spread/scope) and relative width (duration) will change. Fig. 8.1.10 [8.6] gives the curves, which characterize the relations of spread/scope (curve a) and duration (curve b) of sine-quadratic

pulses, at the output and at the input of the perfect filter of low frequencies with the limited band. The pulse duration at the input is equal to 0.17 (scale A), 0.1 (scale B) and 0.05 μs (scale C); on these scales the cutoff frequencies ФНЧ are indicated.

During tests pulse duration usually is selected

$$\tau_{и} = \frac{1}{F_{и}} = \frac{1}{4F_{в}}. \qquad (8.1.2)$$

In recommendations of MKKR, which relate to communication systems through ISZ, requirements for channel of sound tracking of television program are not specified. In connection with this it is appropriate to give requirements for the long-distance circuit of the I class, organized along the radio relay or cable lines for the transmission of broadcast programs [8.5]:

- the reproducible frequency band: 50-10^4 Hz with the nonuniformity of the frequency characteristic, calculated relative to the frequency of 800 Hz, ±1.74 dB in the band 0.2-6 kHz;

- root-mean-square coefficient of the harmonics: at frequencies to 100 Hz - not more than 2%, at frequencies it is more than 100 Hz - not more than 1-2%;

- ratio of signal to the background not less than 55 dB;

- ratio of signal to the interference not less than 63 dB;

- ratio of signal to the distinct transient interference 74 dB.

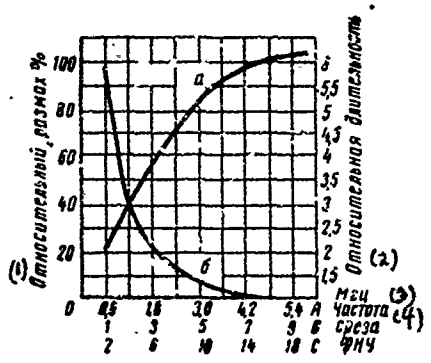

Fig. 8.1.10. Parameters of pulses/momenta at different frequency band.

Key: (1). Aspect ratio %. (2). Relative duration. (3). MHz. (4). Cutoff frequency.

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8.2. Transmission of one sideband (OBP).

Relation of power of unweighted thermal noise and signal at output of line of communications with OBP can be determined according to expression (7.2.3), if value ΔF is considered equal to width of band of TV signal, i.e.,

$$\frac{P_{ш}}{P_{с}} = k\Delta F\left[\frac{T_{26}\,B_{21}}{P_{пз}} + \frac{T_{23}\,B_{22}}{P_{пб}}\right]. \qquad (8.2.1)$$

Taking into account that lower cut-off frequency of TV signal F_1 is considerably lower than upper cut-off frequency F_2, it is possible to count

$$\Delta F = F_2 - F_1 \cong F_2. \qquad (8.2.2)$$

Since in recommendations of MKKR, given in Section 7.1, for TV signal is normalized ratio of spread/scope of picture signal to effective noise voltage, it is necessary to conduct appropriate conversions of expression (8.2.1).

Understanding by $U_ш$ effective value of noise voltage, it is possible to record

$$P_ш = \frac{U_ш^2}{R}. \qquad (8.2.3)$$

Since the actual voltage of the signal

$$U_с = \frac{U_р}{2\varkappa}, \qquad (8.2.4)$$

we will find

$$P_с = \frac{U_р^2}{4R\varkappa^2}. \qquad (8.2.5)$$

From expression (8.2.3) and (8.2.5) it follows

$$\frac{P_ш}{P_с} = \frac{4U_ш^2\varkappa^2}{U_р^2}. \qquad (8.2.6)$$

Substituting (8.2.5) and (8.2.2) in (8.2.1) and introducing visometric coefficient ϵ, which determines decrease of power of noises in transit through weighing circuit, we will obtain

$$\left(\frac{U_ш}{U_р}\right)^2 = \frac{kF_т}{4\varkappa^2}\left[\frac{T_{р6}}{P_{пт}}B_{т1} + \frac{T_{т3}}{P_{п6}}B_{т2}\right]\varepsilon. \qquad (8.2.7)$$

Let us note that frequency characteristic of weighing circuit corresponds to sensitivity of eye to perception of noises. Let us find the value of coefficient ϵ for the weighing circuit, Fig. 8.1.2. Attenuation characteristic of this circuit at the frequency $\Omega=2\pi F$ is determined by the relation

$$\frac{U_2}{U_1} = \frac{1}{1 + \frac{i\Omega L}{R}}. \qquad (8.2.8)$$

Here U_2 and U_1 - respectively voltage on input and output of circuit.

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Designating ratio $L/R=\tau$, we will obtain value of weighting function:

$$\Phi(F) = \left|\frac{U_2}{U_1}\right|^2 = \frac{1}{1 + 4\pi^2 F^2 \tau^2}. \qquad (8.2.9)$$

If on input of weighing circuit oscillations/vibrations (or noises) with spectral power density of G(F) act, then visually received power is determined by expression [8.6]

$$P_2 = \int_{F_1}^{F_2} G(F)\Phi(F)dF. \qquad (8.2.10)$$

Coefficient, which characterizes decrease of power of noises in transit through weighing circuit, is defined as ratio of visual-received power to power of noises at input of filter

$$e = \frac{P_2}{P_{ex}} = \frac{\int_{F_1}^{F_2} G(F)\Phi(F)\,dF}{\int_{F_1}^{F_2} G(F)\,dF}. \qquad (8.2.11)$$

If on input of weighing circuit fluctuation noise with uniform spectrum acts, spectral power density in numerator and denominator of latter/last expression can be carried out to integral sign, as a

result of which

$$\varepsilon = \frac{\int_{F_1}^{F_2} \Phi(F)\,dF}{F_2 - F_1}. \qquad (8.2.12)$$

With substitution (8.2.9), as a result of integration, we will obtain

$$\varepsilon = \frac{\operatorname{arc\,tg} 2\pi F_2 \tau - \operatorname{arc\,tg} 2\pi F_1 \tau}{2\pi\tau(F_2 - F_1)}. \qquad (8.2.13)$$

Assuming/setting F_2=5 MHz, F_1=0, τ=0.33 μs, we will obtain

$$\varepsilon = 0{,}148 (-8{,}5 \text{ dB}). \qquad (8.2.14)$$

With substitution (8.2.14) in (8.2.7) with values of $k=1.38\cdot10^{-23}$ J/deg, $F_2=5\cdot10^6$ Hz, we will obtain

$$\left(\frac{U_ш}{U_p}\right)^2 = 2{,}6\cdot10^{-18}\left[\frac{T_{э6} B_{э1}}{P_{пэ}(\varkappa)} + \frac{T_{э3} B_{э2}}{P_{пб}(\varkappa)}\right]. \qquad (8.2.15)$$

Is here marked $P_п(\varkappa)=P_п\varkappa^2$ - peak power of transmitter OBP.

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If the relation, which stands on the left side of equality (8.2.15), is assigned for each section of the line of communications, then it is possible to find the power of the transmitters

$$P_{п.}(\varkappa) = 2{,}6\cdot10^{-18} T_э B_э \left(\frac{U_p}{U_ш}\right)^2. \qquad (8.2.16)$$

Indices, which determine sections of connection/communication, in latter/last expression are omitted for generality of recording. The recording of indices can carried out in accordance with (8.2.15). Let

us note that the peak power of transmitter is determined by expression (8.2.16).

Bandwidth on SHF with OEP is equal to width of spectrum of video signal, i.e. will be

$$\Delta f_{\cdot} = \Delta F = 5 \text{ MHz}. \qquad (8.2.17)$$

8.3. Transmission with the frequency modulation.

In accordance with (5.5.6) and (5.3.13)

$$\frac{P_{ш}}{P_c} = \frac{\varkappa^2 k \Delta F^2}{3\Delta f_{m2}^2}\left[\frac{\Delta f_6 B_{\Sigma 1} T_{\Sigma 6}}{P_{ш3} \nu_6 z^2} + \frac{\Delta f_3 B_{\Sigma 2} T_{\Sigma 3}}{P_{ш6} \nu_3 z}\right]. \qquad (8.3.1)$$

Here value $z = \frac{\Delta f_{m1}}{\Delta f_{m2}}$, and Δf_{m1} and Δf_{m2} determine maximum value of deviation of frequency respectively in section Earth-satellite and satellite-Earth.

Considering that on the left side of equality (8.3.1) stands relation of effective power, and also taking into account that $\Delta f_6 = \nu_6 \Delta F$ and $\Delta f_3 = \nu_3 \Delta F$, we will obtain

$$\left(\frac{U_{ш}}{U_c}\right)^2 = \frac{\varkappa^2 k \Delta F \Delta F^2}{3\Delta f_{m2}^2}\left[\frac{T_{\Sigma 6} B_{\Sigma 1}}{P_{ш3} z^2} + \frac{T_{\Sigma 3} B_{\Sigma 2}}{P_{ш6}}\right]. \qquad (8.3.2)$$

Passing from effective value of video voltage to peak spread/scope of signal, should be considered equality

$$U_c = \frac{U_p}{2\varkappa}. \qquad (8.3.3)$$

Deviation of frequency Δf_m, proportional to half of spread/scope of picture signal, and spread/scope changes in frequency Δf_p, determined by TV signal, are connected as follows (see Fig. 6.1.1):

$$\Delta f_m = \frac{0{,}7\Delta f_p}{2}. \qquad (8.3.4)$$

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Substituting (8.3.3) and (8.3.4) in (8.3.2) after conversions and introduction of coefficient $\varepsilon_{чм}$, that considers noise transmission through weighing filter, we will obtain expression, which determines square of ratio of effective value of noise voltage to spread/scope of voltage of TV signal

$$\left(\frac{U_ш}{U_p}\right)^2 = \frac{kF_в}{1{,}5}\left(\frac{F_в}{\Delta f_p}\right)^2\left[\frac{T_{зб}B_{Σ1}}{P_{пз}z^2} + \frac{T_{Σз}B_{Σ2}}{P_{пб}}\right]\varepsilon_{чм}. \qquad (8.3.5)$$

In this of expression ΔF it is replaced by value F_2, since $\Delta F = F_2 - F_1 \cong F_2$, if $F_1 \cong 0$.

Let us determine value of coefficient $\varepsilon_{чм}$. With the frequency modulation in accordance with (5.3.9) the spectral noise density is proportional to the square of frequency.

Consequently, in expression (8.2.11) it is possible to consider that

$$G(F) = aF^2. \qquad (8.3.6)$$

With substitution in (8.2.11) expressions (8.3.6) and (8.2.9) with $F_1=0$ we will obtain

$$\varepsilon_{\text{чм}} = \frac{3(2\pi F_2 \tau - \operatorname{arc\,tg} 2\pi F_2 \tau)}{8\pi^3 F_2^3 \tau^3}. \qquad (8.3.7)$$

Assuming/setting F_2=5 MHz; τ=0.33 μs, let us find

$$\varepsilon_{\text{чм}} = 0{,}0234\ (-16{,}3\ \text{дБ}). \qquad (8.3.8)$$

Substituting in (8.3.5) value $\varepsilon_{\text{чм}}$ from (8.3.8) and accepting $k=1.38\cdot10^{-23}$ J/deg; F_2=5 MHz, let us find

$$\left(\frac{U_{\text{ш}}}{U_{\text{р}}}\right)^2 = 3\cdot10^{-17}\left(\frac{1}{\Delta f_{\text{р}}}\right)^2\left[\frac{T_{\text{з6}} B_{\text{з1}}}{P_{\text{пз}}} + \frac{T_{\text{з3}} B_{\text{з2}}}{P_{\text{пб}}}\right]. \qquad (8.3.9)$$

In this expression value $\Delta f_{\text{р}}$ (in MHz) determines spread/scope of change in frequency, caused by transmission of TV signal (from white level to apexes/vertexes of timing pulses).

According to expression (8.3.9), knowing assigned relation $\left(\frac{U_{\text{ш}}}{U_{\text{р}}}\right)^2_{\text{зад}}$ for each section of line of communications, it is possible to find power of terrestrial and onboard transmitters:

$$P_{\text{пз}} = = 3\cdot10^{-17}\left(\frac{1}{\Delta f_{\text{р1}}}\right)^2\left(\frac{U_{\text{р}}}{U_{\text{ш}}}\right)^2_{\text{зад}} T_{\text{з6}} B_{\text{з1}}, \qquad (8.3.10)$$

$$P_{\text{пб}} = 3\cdot10^{-17}\left(\frac{1}{\Delta f_{\text{р2}}}\right)^2\left(\frac{U_{\text{р}}}{U_{\text{ш}}}\right)_{\text{зад}} T_{\text{з3}} B_{\text{з2}}. \qquad (8.3.11)$$

Here $\Delta f_{\text{р1}}$ and $\Delta f_{\text{р2}}$ - spread/scope of change in frequency (in MHz), of called party by transmission of TV signal respectively for section of connection/communication Earth-satellite and satellite - Earth (in MHz).

Width of spectrum in the case of ChM can be determined according

to approximation formula:

$$\left.\begin{array}{ll}\Delta f \cong \Delta f_p, & \text{если (1) } \Delta f_p \gg 2F_в \\ \Delta f \cong (3 \div 4) F_в, & \text{если (2) } \Delta f_p \cong F_в\end{array}\right\}. \qquad (8.3.12)$$

Key: (1). if.

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8.4. Transmission with KIM

Let us examine transmission of TV signal for case to Kims.

In accordance with (5.5.6) and (5.4.19), in the absence of pulse regeneration on board satellite and measurements with visometric filter (to weighing circuit), with coefficient ϵ, let us find

$$\frac{P_ш}{P_с} = \left[\frac{x^2}{2\sqrt{q_{вх1}}\, e^{\frac{q_{вх1}}{2}}} + \frac{x^2}{2\sqrt{q_{вх2}}\, e^{\frac{q_{вх2}}{2}}}\right]\epsilon. \qquad (8.4.1)$$

Here:

$$q_{вх1} = \frac{P_{пв}}{kT_{з6}\Delta f_6 B_{з1}}, \qquad (8.4.2)$$

$$q_{вх2} = \frac{P_{пб}}{kT_{з3}\Delta f_в B_{з2}}. \qquad (8.4.3)$$

We convert expression (8.4.1) in accordance with norms of MKKR, which determine ratio of effective value of noises to spread/scope of picture signal. Taking into account (8.2.6) and (8.2.14), let us find

$$\left(\frac{U_ш}{U_p}\right)^2 = 2\cdot 10^{-2}\left[\frac{1}{\sqrt{q_{вх1}}\, e^{\frac{q_{вх1}}{2}}} + \frac{1}{\sqrt{q_{вх2}}\, e^{\frac{q_{вх2}}{2}}}\right]. \qquad (8.4.5)$$

During calculations on expressions (8.4.5) we enter as follows:

after determination according to expressions (8.4.2) or (8.4.3) value $q_{вх}$, value $q_{вх}e^{0,5q_{вх}}$ can be determined according to graphs/curves for idealized or real values (see Fig. 7.4.2).

For determining power of onboard or terrestrial transmitter we convert expression (8.4.5), considering relation $\left(\frac{U_ш}{U_р}\right)^2$ as given one for each section. In this case

$$\sqrt{q_{вх}}e^{0,5q_{вх}} = 2\cdot 10^{-2}\left(\frac{U_р}{U_ш}\right)^2_{зад}. \qquad (8.4.6)$$

After computations on (8.4.6) value $q_{вх}$ we find through idealized or real curve (see Fig. 7.4.2), and then, in terms of known value $q_{вх}$ we determine power of transmitters in accordance with expressions (8.4.2) and (8.4.3).

In the case of KIM-FM analogous (with 8.4.1) we have

$$\frac{P_ш}{P_с} = \frac{\varkappa^2}{4}\left[\frac{1}{\sqrt{q_{вх1}}e^{2q_{вх1}}} + \frac{1}{\sqrt{q_{вх2}}e^{2q_{вх2}}}\right]\varepsilon. \qquad (8.4.7)$$

After substitution in (8.4.7) expressions (8.2.6) we will obtain

$$\left(\frac{U_ш}{U_р}\right)^2 = 10^{-2}\left[\frac{1}{\sqrt{q_{вх1}}e^{2q_{вх1}}} + \frac{1}{\sqrt{q_{вх2}}e^{2q_{вх2}}}\right] \qquad (8.4.8)$$

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For determining the power of transmitters we will obtain

$$\sqrt{q_{вх}}e^{2q_{вх}} = 10^{-2}\left(\frac{U_р}{U_ш}\right)^2_{зад}. \qquad (8.4.9)$$

Utilizing results of calculation according to expression (8.4.9), with the aid of curves (see Fig. 7.4.2) are determined values $q_{вх}$ and then through expressions (8.4.2) or (8.4.3) power of transmitters are located.

Bandwidth with to Kims in accordance with (5.4.8) will comprise

$$\Delta f \cong 2\Delta Fn. \qquad (8.4.10)$$

Here ΔF - bandwidth, occupied by TV signal, and n - number of pulses per in word. The research has shown [8.7] that the reproduction of television image was acknowledged good on 64 and 128 levels of quantization, i.e., with value of n=6 even 7. In this case the clearness of the boundaries, which correspond to the adjacent levels, into which is divided/marked off television image on the brightness in the case of KIM, can be substantially lowered during the addition to the video signal before the quantization of white noise. This is explained by the fact that during the introduction of the white noise of the boundary of quantization it will be divided/marked off according to the random law.

We will point out that selection of levels of quantization during transmission of television image details, those differing somewhat from each other in brightness. Therefore a number of levels of quantization must be determined by Weber-Fechner's law, i.e., on the logarithmic scale they must be uniform.

8.5. Comparison different of the methods of the transmission of TV signals according to the energy indices.

For estimating energy relationships during transmission of TV si nals calculation of power of onboard and terrestrial transmitters with values of parameters, given in Tables 7.7.1 and 8.5.1, is given below.

In accordance with calculations, led in Section 7.7, for antennas without protective hoods we have:

- weakenings of signal in sections of connection/communication

$$B_{021} = 127{,}2 \text{ dB } (5.3 \cdot 10^{12} \text{ times}), \quad (8.5.1)$$

$$B_{021} = 125{,}7 \text{ dB } (3.7 \cdot 10^{12} \text{ times}); \quad (8.5.2)$$

- noise temperatures of terrestrial and onboard receivers

$$T_{\text{пз}} = 100°\text{K}, \quad (8.5.3)$$
$$T_{\text{пб}} = 1000°\text{K}. \quad (8.5.4)$$

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Calculation with ОБП. In accordance with expression (8.2.16) and data of Table 8.5.1 we will obtain the peak values of the power of the transmitters:

$$P_{\text{пб}}(\text{к}) = 500 \text{ W}; \quad P_{\text{пз}}(\text{к}) = 14 \text{ kW}.$$

Frequency band, occupied in this case, will be equal to width of

band of TV signal, i.e.,

$$\Delta f = \Delta F = 5 \text{ MHz}.$$

Calculation with ChM. Calculation can be carried out with the substitution of data of Table 8.5.1 and values of (8.5.1)-(8.5.4) into formulas (8.3.10) and (8.3.11). During the computations we will consider that the band, occupied during the transmission Δf=50 MHz. Therefore in accordance with (8.3.12) we obtain

$$\Delta f_p = \Delta f = 50 \text{ MHz}.$$

In this case

$$P_{пб} = 2{,}6 \quad \text{W}; \qquad (8.5.5)$$

$$P_{пз} = 74 \quad \text{W}. \qquad (8.5.6)$$

Let us note that obtained value $P_{пб}$ will create at input of terrestrial receiver signal, close to threshold, therefore, if in this receiver measures for reduction in threshold value of signal (for example, will not be introduced negative feedback in frequency) will not be accepted, value $P_{пб}$ for confident work it must be increased 2.5-3 times.

Comparatively low power of transmitters is caused by large gain, which is obtained with selected spread/scope of change in frequency.

Table 8.5.1.

(1) № пп.	(2) Параметр	Величина (3)
1÷18	См. табл. 7.7.1 (4)	См. табл. 7.7.1. (4)
19	Отношение размаха сигнала изображения к эффективному значению напряжения непрерывного теплового взвешенного шума для 99% времени на выходе видеоканала (5)	$\left(\frac{U_p}{U_ш}\right)^2 = +55$ дБ (6) ($3{,}2\cdot10^5$ раз) (7)
20	То же отношение на участке спутник—Земля (8)	$\left(\frac{U_p}{U_ш}\right)^2_2 = 56$ дБ (9) ($4\cdot10^5$ раз) (7)
21	То же отношение на участке Земля—спутник (9)	$\left(\frac{U_p}{U_ш}\right)^2_1 = 62$ дБ (6) ($1{,}6\cdot10^6$ раз) (7)
22	Максимальная частота видеосигнала (10)	$F_в = 5$ Мгц (11)

Key: (1). No. (2). Parameter. (3). Value. (4). See Table 7.7.1. (5). Ratio of spread/scope of picture signal to effective value of voltage/stress of continucus thermal weighed noise for 99% of time on output of video channel. (6). dB. (7). times. (8). The same relation in section satellite - Earth. (9). The same relation in section Earth - satellite. (10). Maximum frequency of video signal. (11). $F_в$=5 MHz.

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If the occupied frequency band for any reasons cannot be more than 20 MHz, then the spread/scope of change in the frequency with the frequency modulation must be lowered to the value

$$\Delta f'_p = 15 \text{ MHz}.$$

In this case:

$$P_{пб} = 22 \text{ W},$$
$$P_{пз} = 630 \text{ W}$$

With the obtained values of the power of transmitters and the band of receiver, equal to 20 MHz, the value of signal will considerably exceed threshold value.

Calculation with KIMS-AM. In accordance with the data of Table 8.5.1 and expressions (8.4.9), (8.4.2) and (8.4.3) we will obtain:

$$P_{n6} = 15 \quad \text{W},$$
$$P_{n3} = 126 \quad \text{W}$$

In this case the bandwidth according to (8.4.10) in the case of six pulses/momenta in word

$$\Delta f = 2 \cdot 5 \cdot 6 = 60 \text{ MHz}.$$

Conclusions/outputs. comparing obtained values, it is possible to note that with identical occupied frequency band application of ChM leads to somewhat smaller power, than KIM. Use of OBP, although it makes it possible to considerably shorten the frequency band, nevertheless causes a considerable increase in the power of onboard and terrestrial transmitters.

8.6. Effect of the Doppler effect, time lag and switching of satellites.

Change in the frequency and strain of frequency spectrum, caused by the Doppler effect and examined in Section 3.5, during the

transmission of television image and sonic tracking prove to be unessential. This is confirmed by different experiments, carried out with the communication systems through the satellites. A certain adjustment of the Doppler effec can be required only during the transmission of colored image.

Signal lag during transmission of television programs does not have value in such a case, when signals of sonic tracking and image delay approximately equally.

Switching from one moving/driving satellite to another can lead to loss of part of information. The latter is explained by the possible disruption of the work of timing mechanism during the switching, which can cause the loss of several fields of image. Furthermore, a change in signal lag during the switching can somewhat destroy the ntinuity of sonic tracking.

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Since the period between the switchings in the real systems will exceed 30 min, disturbance/breakdown in the transmission of video- and sound signals, apparently, they will prove to be not very noticeable.

8.7. On the reception/procedure of television program through ISZ directly to the mass receivers.

During evaluation/estimate of communication system, which makes

it possible to carry out reception of television program with ISZ directly to mass receivers, which are located in population, one should proceed from conditions, under which cost/value of terrestrial receivers and antenna systems will be minimum. This means that in this system can be examined only such conditions of reception/procedure, with which can be used the simplest motionless terrestrial antennas, which make it possible to carry out regular reception of television programs with ISZ, at least, for several hours in a 24 hour period.

Solution of this question is possible only with work through stationary ISZ or through ISZ, which move at large distances from Earth along circular or elliptic orbits. Therefore subsequently, during the evaluation/estimate of the indices of this system, we will consider that it can be realized with the orbits by height/altitude about 40000 km, accepted for the calculations in Section 8.5.

Reception of television programs with ISZ directly to mass receivers can be realized in following frequency bands, diverted for telecast

1 range ... of 47-68 MHz

2 range ... of 87.5-100 MHz

3 range ... of 174-230 MHz

4 range ... of 470-662 MHz

5 range ... of 662-960 MHz

6 range ... of 11.7-12.7 MHz.

Taking into account curves of Fig. 4.3.2, it is possible to say that as a result of a considerable increase in the cosmic noises the first three ranges for the use in this system are unsuitable. On the other hand, from the calculations, carried out in Section 8.5 for the frequencies of 4000 MHz, it follows that with the considerable sizes/dimensions of terrestrial antennas (with the amplification approximately 60 dB) for the transmission and reception of television program the receivers with a small inherent noise level are required. Taking into account that the indices of mass receivers and antennas will be considerably worse than accepted during these calculations, it is obvious that the work in the sixth range (11.7-12.7 MHz) will require the extremely large power of the onboard transmitters, which it will difficultly realize in the next years.

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Let us conduct the evaluation/estimate of the power of onboard transmitter during the transmission of television program with ISZ directly to the mass receivers to 4 and the 5th ranges (470-968 MHz). In this case we will consider that the frequency modulation of oscillations is realized. This means that in the calculations to admissibly consider only thermal noises it is possible not to consider the man-made interferences.

It is obvious that in accordance with expression (3.2.2) upon transfer into this range of sizes of losses in free space will change

into number of times, proportional to square of change in frequency. Thus, attenuation, in comparison with the calculation, given in Section 7.7, used in Section 8.5, decreases approximately 70 times upon transfer for the frequency of 470 MHz, and approximately 17 times upon transfer for the frequency of 958 MHz, i.e. to 18.5 and 12 dB respectively.

If we consider that the reception/procedure of emissions with ISZ is conducted on the multielement antenna with the amplification 8-10 dB (instead of 58 dB, accepted during calculations in Section 8.5), and the value of receiver noise with the input device on tunnel diodes composes 500°K (instead of 50°K, accepted for the calculations in the Section 8.5), then we will obtain that the power of onboard transmitter upon transfer into the range 470 MHz must be increased on - 18.5+50+10=41.5 dB. Upon transfer into the range 958 MHz this increase will compose - 12+50+10=48 dB.

Allowing/assuming reduction in signal-to-noise ratio at output by 10 dB in comparison with norms of MKKR for communication systems and taking into account (8.5.5), it is possible to say that even with ChM at frequencies of 470 MHz and 958 MHz will be required onboard transmitters by power about 3.7 kW and 16 kW respectively. These values are obtained during the amplification of the onboard transmitting antennas 14 dB (see Table 7.7.1). If the amplification of onboard antennas will be increased to 30 dB, the power of onboard transmitters must be approximately 0.09 kW and 0.4 kW respectively.

It must be noted that with the antennas with this amplification the territory, operated by one ISZ, somewhat decreases. Let us note also that the creation of onboard antenna with the amplification 30 dB at the frequency of 470 MHz is hindered/hampered as a result of its large sizes/dimensions.

On the basis of given tentative calculations it is possible to consider that from technical side direct reception of TV signals from satellite can prove to be possible. However, during the construction of this system together with the technical side should be considered a whole series of organizational positions. To them should be related such questions, as the difference of standard time, which will hinder/hamper the composition of the common television program, simultaneously transmitted to the large territory, a difference in the languages, the need for the conversion of television standards in the mass receivers of different countries, etc. The enumerated difficulties considerably complicate the creation of a similar system. these difficulties even more greatly they will increase, if we consider the existing pool of the individual television receivers, for work of which in the system of the direct reception of transmissions with ISZ will be required the creation of special attachments (low-noise amplifiers, the switches of standards, antennas, etc.).

In connection with enumerated difficulties, apparently, is more advisable reception of emissions with ISZ to collective terrestrial relay receivers. In this case it is much easier to solve the

enumerated above questions, and appear the following advantages:

- Possibility of use in section satellite - Earth optimum for space communications frequencies independent of ranges, in which will occur reception to mass receivers.

- Coordination of schedule of international transmissions with schedule of local and national broadcasting.

- Possibility of use in section satellite - Earth of interference-free forms of modulation and any transformations of signals, which ensure improvement of noise immunity of reception and contraction of occupied frequency band.

- Possibility of decreasing interferences with ground-based radio services from emissions from satellite. This is caused by the fact that with the collective reception at the terrestrial stations it is possible to use the more directional antennas, the low-noise amplifiers, that it will make it possible to reduce the power of onboard transmitter.

- Decrease of power, weight, sizes and simplification in construction/design of onboard transmitter and devices/equipment of feed. This leads to an increase in the reliability, it makes it possible to enforce the redundancy of equipment and to increase the period of its service.

An example of a similar system is system the orbit, realized in

the USSR to 50-th anniversary of the great October Socialist Revolution. The description of system orbit given in Section 10.5.

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Chapter 9.

Joint use of frequency bands by communication systems through ISZ wit' other ground-based radio facilities.

9.1. Initial information.

Up to onset of communication systems through ISZ frequency band from 3 to 8 GHz, which is most advisable (see Sectio· 4.3), it was already distributed between different radio services. Therefore for the communication systems through ISZ in this range it proved to be possible to isolate only the separate frequency bands in essence on the base of joint use, since the displacement/movement of the radio aids operating at present into other frequency bands is extremely difficult over the technical and economic reasons.

Joint use of frequency band by several radio services leads to interferences and some limitations of those parameters of radio aids, on which depends interference level. To such parameters can be referred the radiated power, the angles of elevation of antennas, direction of routes, distance between those radio stations, which can create and receive undesirable radio transmissions, etc.

Examination of interferences showed that joint operation in general/common band of frequencies of communication systems through

ISZ and radio relay lines of sight (RRL) from technical side can be realized most simply. With the coincidence of frequencies not with RRL, but with some other services, which have the transmitters of larger power, highly sensitive receivers or changing radiation zones, considerable difficulties appear.

Therefore at extreme administrative conference of radio communication (ChAKR) in 1963 for communication systems through ISZ frequency bands, which coincide in were isolated essence with frequency bands, diverted for radio relay lines, which are placed in category of fixed/recorded services. The enumeration of the frequency bands, isolated for the communication systems through ISZ, is given in Appendix 3.

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Fig. 9.1.1 gives the diagram of the onset of the interfering signals (interferences) between the ground radio-relay stations and the communication system through ISZ. From this diagram it follows that the interferences can be broken into four groups:

A. Interferences from the ground stations with the receivers of terrestrial stations.

B. Interferences from ground stations with receivers of onboard stations.

C. Interferences from terrestrial stations with receivers of ground stations.

Д. Interferences from the onboard stations with the receivers of ground stations.

Certainly, besides enumerated forms of interferences, there can be another, for example, interferences between different communication systems, which use ISZ, interferences between several radio relay lines, etc.

During analysis of interfering signals it is necessary to keep in mind the fact that their level is not time-constant. This is explained by the following reasons: by change in the time of the value of attenuation of the interfering signal during its propagation along the surface of the Earth, by the transit of satellite in the space, and also by the motion of the antenna of the terrestrial station, directed in the side of satellite.

In connection with the fact that at input of receivers interfering signal levels are variable, changes in time and ratio of useful signal to interfering signal at output of system. In accordance with this within ISZ the values of this relation for different percentages of the total time of the reception of signals are established by the recommendations of MKKR for the communication systems.

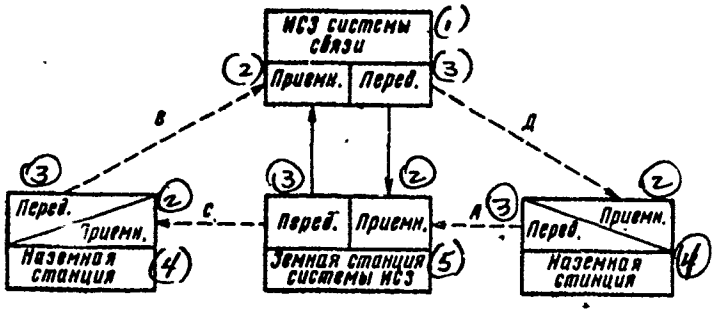

Fig. 9.1.1. Diagram of onset of interfering signals.

Key: (1). ISZ of the communication system. (2). Receiving. (3). Before. (4). Ground station. (5). Terrestrial station of system of ISZ.

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According to the recommendation of MKKR [9.1] the communication systems through ISZ, using frequency bands together with RRL, must be designed so that the power of the noise, created by all transmitters of RRL, at the point of the zero relative level of any telephone channel of the fundamental standard circuit of the communication system through ISZ would not exceed:

- 1000 pW of the psophometrically increased average/mean power in any hour,

- 1000 pW of the psophometrically weighted mean power in one minute during more than 20% of time of any month,

- 50,000 pW of the psophometrically weighted mean power in one

minute during more than 0.03% of time of any month.

These values must be part of total quantity of noises, given in Section 7.1.

Taking into account that psophometric measurements with respect to unweighted noises give decrease of power of noises on 2.5 dB, on the basis of given recommendation it is possible to obtain following ratios of power of experimental tone to power of unweighted noises at point with zero relative level for any telephone channel, which must not be exceeded:

- in any hour - $Q_{03}=57{,}5$ DBO,

- during is more than 20% time of any month - $Q_{03}=57{,}5$ DBO,

- during is more than 0.03% time of any month - $Q_{03}=40{,}5$ DBO.

For RRL, frequencies, as communication system working in the same bands through ISZ, is analogous recommendation; power of noises, caused by operation of terrestrial stations and communication satellites at point with relative zero level of any telephone channel of hypothetical standard circuit of RRL, by length of 2500 km, there must not exceed following values:

- 1000 pW of psophometrically weighted mean power in any hour,

- 1000 pW of psophometrically weighted mean power during are more than 20% of time of any month,

- 50,000 pW of psophometrically weighted mean power in minute during are more than 0.01% of time of any month.

Let us note that given values can be expressed in the form of ratio of power of test signal to unweighted power of noise analogous how it is given above for communication systems through ISZ.

Fig. 9.1.2 gives norms taking into account psophometric weighing of noises, recommended by MKKR [9.1] to the maximum permissible sound levels in telephone channel of hypothetical standard circuits, established/installed for communication systems through ISZ and for RRL.

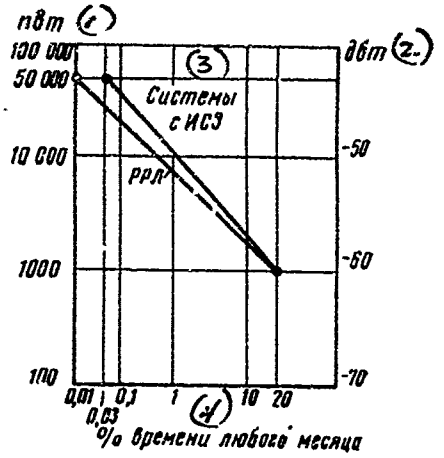

Fig. 9.1.2. Maximum permissible sound levels in telephone channel of communication systems through ISZ and RRL.

Key: (1). pW. (2). dBT. (3). Systems with ISZ. (4). Time of any month.

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On the basis Fig. 9.1.2 it is possible to determine the values of the noises, caused by the joint use of a frequency band for those percentages of the time of any month, which are located between the points described earlier.

For guaranteeing corrected values of noises of MKKR were accepted following recommendations to amount of radiated power:

1. For decreasing the interferences with the communication systems through ISZ it is recommended so that maximum power, effectively emitted antenna of RRL, would not exceed 55 dBW, and the power of transmitter at the input of this antenna would be not more

than 13 dBW.

2. In connection with the fact that communication systems through stationary ISZ will be irradiated by RRL continuously on fixed levels, for their protection from these emissions more rigid recommendations are provided for: norm 55 dBW relates only to that newly projected RRL, whose antennas are directed to region, distant behind orbit of stationary satellites more than on 2°. If antennas of RRL are oriented into the part of the space, which is within the limits of less than 0.5° from the orbit of stationary satellites, the power, emitted by RRL, must not exceed 47 dBW. With a change of orienting the antennas of RRL within the limits from 0.5 to 1.5° from the direction in orbit of stationary satellites the recommendation to the permissible radiated power is determined on the linear dependence in the limits from 47 to 55 dBW respectively.

So that values of noises, recommended for RRL, would not be exceeded, MKKR were developed recommendation, to which for communication systems through ISZ are limited power, emitted by terrestrial stations, and densities of power flux, created by onboard stations. These recommendations [9.1, 9.2, 9.17] are the following:

1. In the frequency bands from 1 to 10 GHz, which are together utilized by communication systems through ISZ and RRL, the maximum value of the density of power flux on the surface of the Earth, created by the transmissions of the space stations of the

communication systems through ISZ under all conditions and with methods of modulation, must not exceed values $\Phi = -152 + \frac{\theta}{15}$ dBW/m² in any band 4 kHz. Here value Θ determines the angle of arrival of oscillations/vibrations at the surface of the Earth, calculated off the horizontal plane. It is implied that the limitation of the density of power flux indicated on the surface of the Earth must be made during calculations of attenuation between the antennas under the conditions, analogous to free space.

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2. Virtual rating, emitted by terrestrial station in horizontal plane in any direction taking into account screening constant of terrain by obstruction, which limits β_{min} (see Table 9.2.1) in any band with width of 4 kHz, must not exceed +55 dBW, with exception of following two cases:

- if in certain direction distance from terrestrial station to boundary of territory of another administration exceeds 400 km, then limit, equal to +55 dBW in any band 4 kHz, can be in this direction increased on 2 dB for every 100 km of distance, which exceed 400 km;

- limit can be increased by agreement between interested and affected Administrations.

However, in these two cases power coefficient there must not exceed value of +65 dBW in any frequency band with width of 4 kHz.

3. Transmitting antennas of terrestrial stations must not emit electromagnetic energy at angles of increase of less than 3°, measured from horizontal plane to central axis of major lobe of antenna. Can be exception/elimination only those cases, which are coordinated between the interested or affected Administrations.

Let us note that in given recommendations is emphasized the fact that radiated power or density of power flux must not exceed values indicated in any frequency band, equal to 4 kHz. This means that in any section of the emitted frequency spectrum in the presence or in absence of modulation, the amount of radiated power must satisfy the values indicated. In order to satisfy this condition, in the systems with ChM in the absence of modulation (i.e. with the zero or with the time-constant modulating voltage/stress) it is necessary to use special diagrams or methods, which ensure scattering (dispersion) power at the carrier frequency along the spectrum. If in the communication system with ChM special measures for the dispersion of carrier output are not taken, then for fulfilling the requirement given above on all possible levels of the modulating voltage/stress (including with its equality to zero) will be required sharp reduction in the radiated power.

In recommendation of MKKR [9.1] is noted that for communication systems through ISZ, which work in ranges 3.4-4.2 GHz and 5.8-6.425 GHz, in the case of emitting uniform and wide spectrum selection of

the carrier frequencies makes no difference. But if in the communication systems through ISZ weak loading (for example, the decrease of a number of transmitted telephone conversations) is utilized by ChM, it is possible, and even desirable to then select the carrier frequencies so that they either would coincide with the carrier frequencies of RRL or they would be shifted relative to the carrier frequencies of RRL the value of the highest frequency of the group spectrum.

Table 9.2.1.

$\beta_{мин}$	Между 0° и 1° (2)	Между 1° и 2° (2)	Между 2° и 3° (2)	Между 3° и 4° (2)	4° и более (3)
B_{01}, дб (4)	0	10	17	23	25

Key: (1). Between. (2). and. (3). and more. (4). dB.

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Possibility of joint use of frequencies by two concrete/specific stations depends on following factors:

- tolerance level of interfering signal $P_{м\,вых}$ in telephone either television channel at output of system of communications, assigned for different percentages time of observations, or from permissible for different percentages time of observations of ratio of power of fundamental (useful) signal $P_{с\,вых}$ to power of interfering signal at output of channel of communication

$$Q_{вых} = \left(\frac{P_с}{P_м}\right)_{вых}; \qquad (9.1.1)$$

- permissible relation of power of fundamental and interfering signal at input of receiver for different percentages of time of observations

$$Q_{вх} = \left(\frac{P_с}{P_м}\right)_{вх}; \qquad (9.1.2)$$

- power of transmitter of fundamental and interfering signal $P_п$, $P_{пм}$;

- total attenuation of signal between output of transmitter of

fundamental signal and input of receiver $B_{\text{з}}$ (taking into account effective amplification by receiving and transmitting antennas in direction of propagation of fundamental signal and duration for time of observations);

- total attenuation of signal between output of transmitter of interfering signal and input of receiver $B_{\text{зм}}$ (taking into account amplification of receiving and transmitting antenna in direction of propagation of interfering signal and duration);

- coefficient of attenuating interference (KOP) - κ, which depends on special features of joint passage of fundamental and interfering signal through receiver and can be determined in the form of relation

$$\kappa = \frac{Q_{\text{вых}}}{Q_{\text{вх}}}. \qquad (9.1.3)$$

Let us note that value κ is determined by form of modulation of fundamental and interfering signal, by detuning between carrier frequencies of fundamental and interfering signal, special features of these signals (by parameters of telephone or TV signals, by form of multiplexing), by selectivity of receiver, etc. Value κ can substantially change near the threshold value of signal. Therefore subsequently will always count

$$Q_{\text{вх}} \gg q_{\text{вх пор}} > 1, \qquad (9.1.4)$$

where $q_{\text{вх пор}}$ is determined the ratio of the power of threshold signal to the power of the thermal noises of receiver. During calculations

of power capacity of the interfering signal should be taken the smallest value of value κ, since precisely this value will correspond to the worse/worst conditions.

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Let us agree by indices "0" to designate values, expressed in dB or dBW. Then, in accordance with the previously introduced designations, it is possible to record the following expression for determining the relation of received signal to the nterfering on the output of the system of communications

$$Q_{0\text{вых}} = Q_{0\text{вх}} + \varkappa_0 = P_{0\text{п}} - B_{0\Sigma} - (P_{0\text{пм}} - B_{0\Sigma\text{м}}) + \varkappa_0. \qquad (9.1.5)$$

In this expression $B_{0\Sigma}$ and $B_{0\Sigma\text{м}}$ it is determined in accordance with (3.1.2) or (3.1.4), and values $P_{0\text{п}}$ and $P_{0\text{пм}}$ are power of transmitters, expressed in dBW. Value $B_{0\Sigma\text{м}}$ depends both on propagation conditions and on a change of orienting the antennas, which radiates and which receives the interfering signal. For the interferences of group A and C (see Fig. 9.1.1) this change in the orientation is caused by the displacement of the antenna radiation pattern of terrestrial station in the space (with the tracking the motion of satellite) relative to the antenna radiation pattern fixed in the space of ground station. For interferences B and Д, besides a cnange of orienting by onboard receiving and transmitting the antennas relative to ground stations, it is necessary to consider change in the distance between the ground stations and ISZ. For the interferences groups A and C of the distance between the stations in question are constant/invariable (case of mobile radio-relay stations it is not

characteristic).

One should note that expression (9.1.5) relates to case of effect of one jamming station. In several jamming stations according to expression (9.1.5) it is necessary to determine $Q_{0\,вых}$ for each station and to sum up them taking into account the simultaneity of effect. If stations affect not simultaneously, then it is necessary to determine values $Q_{0\,вых}$ and corresponding percentages of the time of effect from the total time of observations.

Expression (9.1.5) makes it possible to decide not only direct problem - to find value $Q_{0\,вых}$, of, and reverse/inverse - from value $Q_{0\,вых}$, assigned on value and percentage of time of observations (which let us designate $Q_{0з}$), to determine minimum distance $L_{мин}$ between receiving and jamming station, with which this value will be provided. For this should be calculated value

$$B_{0\Sigma м} = Q_{0з} - P_{0п} + B_{ЭВ} + P_{0пм} - \varkappa_0, \qquad (9.1.6)$$

and then to find the permissible distance between stations $L_{мин}$, as function from $B_{0\Sigma м}$.

Value $L_{мин}$, with which $Q_{0\,вых} = Q_{0з}$ (where $Q_{0з}$ it is determined by recommendations of MKKR to value of noises, caused by work of systems in combined ranges), it is called coordination distance.

In the case of several jamming stations for each of them can be assigned the value $Q_{0з}$ and corresponding percentage of time, and then

necessary distance between stations is determined.

One should note that expressions (9.1.5) and (9.1.6) can be solved for band of frequencies of any width. In the materials of MKKR width of band, relative to which are conducted some calculations and normalization is realized, is accepted equal to 4 kHz.

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Therefore values P_{0n} and P_{0nm} must be substituted in (9.1.5) and (9.1.6) in dBW, calculated for that section of the emitted spectrum with a width of 4 kHz, in which their relation is smallest. If it is necessary to pass from one band 4 kHz to the next, occupied by telephone signal ΔF_k equal to 3.1 kHz, the obtained results it is necessary to multiply by value

$$z = \frac{3,1}{4} = 0,78, (-1,1 \text{ db}). \qquad (9.1.7)$$

9.2. Attenuation of the interfering signal between receiver and transmitter of terrestrial and ground station.

Attenuation of interfering signal between receiver and transmitter of terrestrial and ground station depends on following fundamental reasons:

- distances between stations,

- character of route, above which occurs propagation of

interfering signal (dry land, water surface),

- climatic conditions,

- special features of relief,

- mutual orientation of antennas.

In accordance with enumerated reasons total attenuation

$$B_{0\text{ум}} = B_{0\text{м}} + B_{0\text{l}} - \kappa_{0\text{пол}} + B_{0f\eta} - G_{0\text{з}}(\theta') - G_{0\text{м}}(\theta''). \quad (9.2.1)$$

Here $B_{0\text{м}}$ - fundamental losses, i.e., attenuation of interfering signal between isotropic antennas, not exceeded during percentages of time upon consideration of tropospheric scattering;

$B_{0\text{l}}$ - attenuation of interfering signal, caused by presence of natural obstacles between stations (see Table 9.2.1);

$\kappa_{0\text{пол}}$ - attenuation, caused by polarizational losses (see Section 3.4), i.e., not by conformity to polarization of interfering signal of antenna polarization;

$B_{0f\eta} = b_{0f\text{п}} + b_{0f\text{пр}} + e_{0\phi\text{п}} + e_{0\phi\text{пр}}$ - attenuation due to filters and feeders (see Sections 3.1 and 3.6);

θ' or θ'' - angles between axis of major lobe of antennas of

terrestrial and ground station under working conditions and straight/direct, connecting station;

$G_{0н}(\theta'')$ - antenna gain of ground station in direction of terrestrial station;

$G_{0з}(\theta'$, - antenna gain of terrestrial station in direction of ground station. Let us note that since the antenna of terrestrial station always must be directed toward the satellite, the value of angle θ', and consequently, and $G_{0з}(\theta')$ prove to be those changing in the time (with exception of the cases, which relate to the systems with the stationary communication satellites). For the calculations in expression (9.2.1) should be substituted value $G_{0з}(\theta')$, which will be exceeded during $\varepsilon_з$ the percentages of time.

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Determination of values $G_{0н}$ and $G_{0з}$ with value of efficiency of antennas, which differs little from one, can be carried out according to expressions (3.3.4) or (3.3.5), and also using the curves in Fig. 3.3.7. Fig. 9.21 shows the exemplary/approximate character of dependence $G_{0з}(\theta', \varphi')$ for different percentages of time. On this graph is shown also value $\theta_м$, whose selection is determined by the conditions, examined below (see Section 9.3).

Determination of value $B_{0м}$ for land routes above comparatively flat Earth in moderate and subtropical areas can be carried out

according to curves, Fig. 9.2.2 and 9.2.3 [9.3]. The curves of Fig. 9.2.2 characterize attenuation at the frequency of 4 GHz at the height of the antennas above the surface of the Earth 50 m and 15 m.

For determining value of attenuation with distance of x (km) between stations at height of antennas h_1 and h_2(m) in Fig. 9.2.2 should be taken distance $L = x + 45 - 4{,}1\sqrt{h_1(м)} + \sqrt{h_2(м)}$ км.

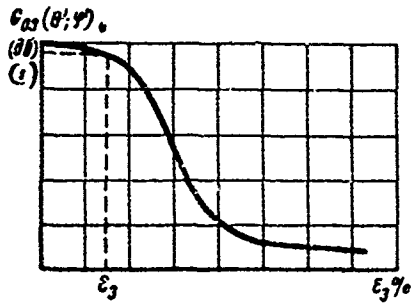

Fig. 9.2.1. Antenna gain of terrestrial station in direction of ground station, exceeded during ε_3 % time.

Key: (1). (dB).

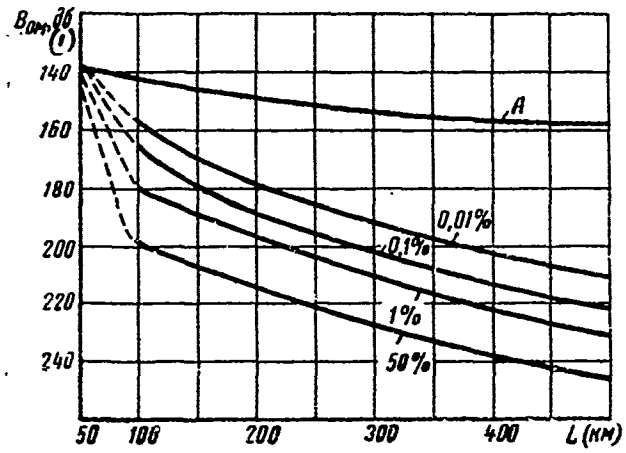

Fig. 9.2.2. Attenuation of signal between isotropic antennas at frequency of 4 GHz, not exceeded during percentage of time (moderate areas, dry land) indicated.

Key: (1). dB.

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In order to determine attenuation at frequencies, different from 4 GHz, to data, obtained from Fig. 9.2.2, should be added values, found for appropriate percentage of schedule time Fig. 9.2.3. Curve

A, given in Fig. 9.2.2, characterizes attenuation of signal at frequency of 4 GHz during free-space propagation. The correction curve for 0.01% of time, given in Fig. 9.2.3, is applicable and for by curve A, Fig. 9.2.2.

If between interfering and receiving stations is natural obstacle, which is observed in limits of angle of elevation β, and it is located at a distance from 5 to 16 km, attenuation $B_{0,1}$, introduced by this obstruction, can be found according to data of Table 9.2.1 [9.1].

From recommendations of MKKR, given in Section 9.1, it follows that upon transfer from hourly mean power of noises to power of noises, which can be exceeded during low percentages of time (0.03 and 0.01%), is allowed/assumed increase of power of noises into number of times, equal to 50,000/1000=50 (17 dB). On the basis of the curves, shown in Fig. 9.2.2 and 9.2.3, it is possible to conclude that the transition/junction from attenuation lengths, which are not exceeded in the long period (for example, during 50% of time), to the attenuations, not exceeded during 0.1 or 0.01% time, leads to the decrease of attenuation on 25-30 dB. This will cause an increase in the interfering signal level on 25—30 dB and the corresponding increase in the power of noises. Consequently, to satisfy the recommendation of MKKR for power capacity of noises, which can be exceeded during the low percentages of time, will be considerably more difficult than to the noises, exceeded during 20% time and more.

Therefore checking the fulfillment of the recommendations of MKKR to the power of noises, exceeded during 0.03 and 0.01% time, must be realized first of all. Designating the percentages of time, determined by the recommendations of MKKR through ϵ, we will obtain the condition

$$\varepsilon = \varepsilon_2 \varepsilon_3. \qquad (9.2.2)$$

Here value ε_2 determines parameter of curve in Fig. 9.2.2 and 9.2.3, and ε_3 - on graph/curve, analogous to Fig. 9.2.1, which relates to specific case.

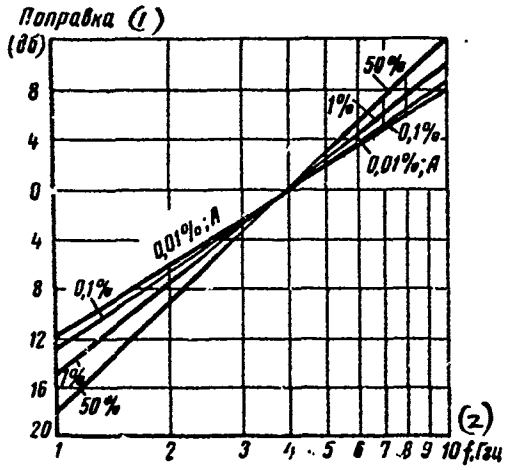

Fig. 9.2.3. Correction curves to graphs/curves Fig. 9.2.2.

Key: (1). Correction (dB). (2). GHz.

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Taking into account that according to Fig. 9.2.2 and 9.2.3 attenuation of signal can be found only for discrete/digital values $e_в$, in accordance with (9.2.2) we obtain, that if terrestrial station is receiving, and, consequently, ϵ=0.03%, then when $e_в$=0,1% antenna gain of terrestrial station must be determined for value

$$\varepsilon_з = 0{,}3\%. \qquad (9.2.3)$$

But if terrestrial station is transmitting, then according to recommendation of MKKR value ϵ=0.01%. Therefore when $e_в$=0,1% the antenna gain of terrestrial station must be determined for

$$\varepsilon_з = 0{,}1\%. \qquad (9.2.4)$$

If during calculations instead of values (9.2.3) or (9.2.4) average/mean antenna gain of terrestrial station in direction of

ground station will be determined, then for determination of $B_{ом}$ from curves of Fig. 9.2.2 should be taken value $\varepsilon_в=0{,}01\%$ of time.

During use of stationary satellites antenna of terrestrial station is fixed and therefore value $G_{00}(\theta')$ does not change in time. Since in this case $\varepsilon_э=1$, in accordance with (9.2.2) one should accept $\varepsilon_в \leqslant \varepsilon$. For this value $\varepsilon_в$ according to Fig. 9.2.2 and 9.2.3 the value of attenuation in the case of stationary satellites must be determined.

During use of expressions (9.2.3) and (9.2.4) it is necessary to have dependences, analogous to those given in Fig. 9.2.2, for different areas of terrestial globe when $\varepsilon_в=0{,}1\%$.

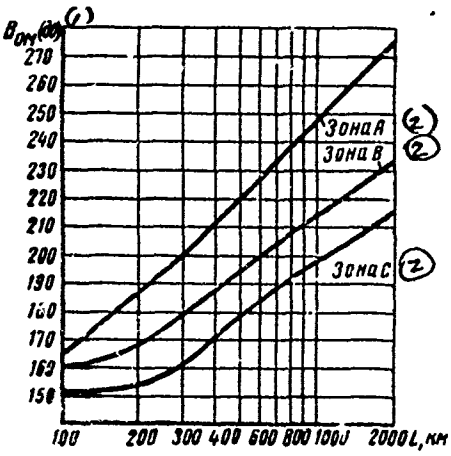

Fig. 9.2.4.

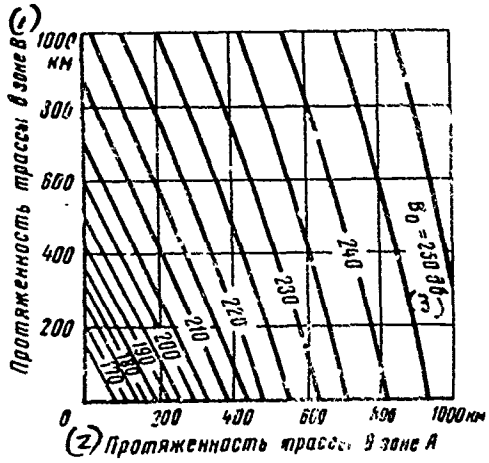

Fig. 9.2.5.

Fig. 9.2.4. Attenuation of the signal between the isotropic antennas at the frequency of 4 GHz, not exceeded during 0.1% of time.

Key: (1). (dB). (2). Zone.

Fig. 9.2.5. Auxiliary graph/curve.

Key: (1). Extent of route in zone B. (2). Extent of route in zone A. (3). db.

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Such dependences for the oscillations/vibrations with a frequency of 4 GHz are given in Fig. 9.2.4 [9.2], moreover:

- zone A - corresponds to propagation conditions for oscillations/vibrations along the dry land,

- zone B - to propagation above the sea on the latitudes of more than 23.5° N or 23.5° S,

- zone C - to propagation above the sea on the latitudes between 23.5° NL and 23.5° SL.

For determining attenuation at frequency, not equal to 4 GHz, it follows to use correction, found according to Fig. 9.2.3 at value ε_B=0,1%. The value of correction should be accumulated with value B_{0M}, obtained from the curves of Fig. 9.2.4.

In the case, if route between stations proves to be mixed, i.e., simultaneously are passed through several different zones, calculation of attenuation can be conducted approximately with the aid of auxiliary graphs/curves Figs 9.2.5, 9.2.6 and 9.2.7 [9.2], [9.4]. On these graphs/curves the dependence of attenuation B_0 (dB) on the extent of route in each of two zones individually is shown.

9.3. Attenuation of the interfering signal between the receiver and the transmitter of ground and onboard station.

Total attenuation of interfering signal between receiver and transmitter of onboard and ground station can be determined according to expressions (3.1.2) and (3.1.4) in accordance with recommendations, given in Chapter 3.

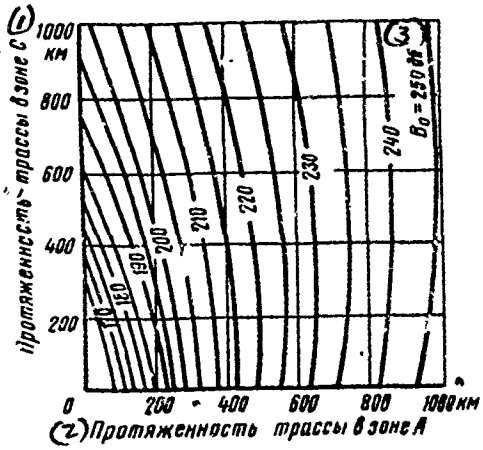

Fig. 9.2.6.

Fig. 9.2.7.

Fig. 9.2.6. Auxiliary graph/curve.

Key: (1). Extent of route in zone C. (2). Extent of route in zone A. (3). dB.

Fig. 9.2.7. Auxiliary graph/curve.

Key: (1). Extent of route in zone C. (2). Extent of route in zone B. (3). dB.

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Thus, in general form we have

$$B_{0\text{см}} = B_0 - \kappa_{0\text{пол}} + B_{0\text{ф}} - G_{0\text{н}}(\Theta') - G_{0\text{б}}(\Theta''). \qquad (9.3.1)$$

Here

$$B_0 \cong B_{0\text{св}} = 10\lg\left(\frac{4\pi L}{\lambda}\right)^2, \qquad (9.3.2)$$

L - distance between ground and onboard station (m), λ - wavelength of interfering signal (m), $G_{0\text{н}}(\Theta')$ and $G_{0\text{б}}(\Theta'')$ - amplification of ground-based and onboard antenna in direction Θ' and Θ'' relative to

axis of major lobe.

In connection with the fact that during motion of satellites continuously change the parameters, in (9.3.1), obtaining general/common expressions is difficult. The solution of specific problems proves to be simpler, although it also requires sufficiently bulky and tedious calculations.

Principles of these calculations can be explained in examination of following two cases:

- interaction of ground station and onboard station, established/installed on satellite, which moves along random circular orbit.

- interaction of ground station and onboard station, established/installed on satellite, which moves along equatorial orbit.

Interaction of ground station and station of satellite, which moves along random circular orbit.

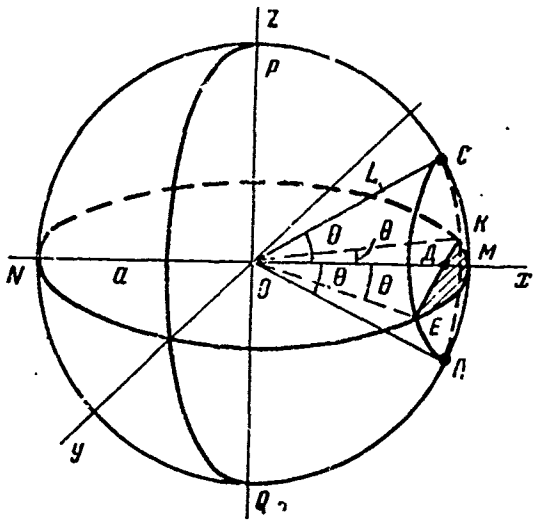

Fig. 9.3.1. To conclusion/output of formula (9.3.9).

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Let us assume that the satellite with the nondirectional antenna moves along the circular orbit, so distant from the surface of the Earth that it is possible a sufficient approximation/approach not to consider with its sizes/dimensions. This case is shown in Fig. 9.3.1, where through MNPQ is designated the orbital surface of radius L, in center of which point O designated the Earth. We will consider that the antenna radiation pattern of ground station is characterized by a certain angle Θ, in limits of which the antenna gain exceeds value of G, moreover when $\Theta=0$, $G=G_{\text{макс}}$. Tak as antenna is established/installed on the Earth, reception or transmission of the signal of such of antenna can be realized only in the region, the located above horizontal plane, i.e., only in region *CKME*. This means that in the random position of satellite in the space the probability of the

reception of signals from the satellite (or transmission to the side of satellite) with the antenna gain, which exceeds G, will be determined by the relation of the part of the surface of sphere *CKME*, characterized by angle Θ, to the entire surface of sphere MNPQ, i.e.

$$p(G) = \frac{S(CKME)}{S(MNPQ)} = \frac{0{,}5S(CKAEM)}{S(MNPQ)} = \frac{\pi L(DM)}{4\pi L^2}.$$

Taking into account that $ДМ$=L-L cos Θ, we will obtain

$$p(G) = 0{,}25(1-\cos\Theta). \qquad (9.3.3)$$

For determining level of interfering signals, which arrive from different directions, let us consider special features of antenna radiation pattern of ground station. We will consider that the form of major lobes of the antenna of ground station approaches Gaussian curve. In this case the probability of reception or transmission, the antenna of signal with level G in the direction Θ can be described by expression [9.5]

$$p(G) = \frac{A}{G(\Theta)} + B\lg G(\Theta) + C, \qquad (9.3.4)$$

where A, B and C - constant coefficients.

We will consider that the numerical values of the antenna gain are equal to:

$$\left.\begin{array}{l} G(\Theta) = 10^4 \ (1)\text{при } \Theta = 0° \\ G(\Theta) = 5\cdot 10^3 \ (1)\text{при } \Theta = 0{,}7° \\ G(\Theta) = 1 \quad (1)\text{при } \Theta \geqslant 20° \end{array}\right\}. \qquad (9.3.5)$$

Key: (1). with.

Let us note that level of lateral and rear lobes/lugs in region, where Θ≥20°, is accepted somewhat greater than as follows from Fig.

3.3.7, since the antennas of ground stations are usually made less thoroughly than the antenna of terrestrial stations, to which the graph/curve indicated relates.

After equating (9.3.4) and (9.3.3) with substitution of values (9.3.5), we will obtain system of equations, from which let us find:

$$A=1{,}47\cdot10^{-2};\quad B=-7{,}61\cdot10^{-4};\quad C=3{,}03\cdot10^{-4}.$$

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Therefore expression (9.3.4) in limits from $0\leq\Theta\leq20°$ will take form

$$p(G)=\frac{1{,}47\cdot10^{-2}}{G}-7{,}61\cdot10^{-4}\lg G+3{,}03\cdot10^{-4}. \qquad (9.3.6)$$

Taking into account uniform motion of satellite along circular orbit and assigning different values for G in interval $1\leq G\leq10^4$, according to expression (9.3.6) it can be calculated percentage of time, during which interfering signal will be received (or transmitted), antenna ground station from different direciton within limits of $0<\Theta<20°$. The results of this calculation are given in Fig. 9.3.2.

For determining analogous dependence in region of $20°\leq\Theta\leq180°$ let us appear as follows.

Analogously with antenna radiation pattern accepted in Fig. 3.3.7 we will consider that

$$\left.\begin{aligned}G(\Theta)&=1 \quad \text{при } \Theta=20°\\ G(\Theta)&=0{,}1 \quad \text{при } \Theta=180°\end{aligned}\right\}. \qquad (9.3.7)$$

Key: (1). with.

We will approximate probability of reception (or transmission) of antenna ground station of signal in region of $20° \leq \Theta \leq 180°$ in the form of expression

$$p(G) = MG(\Theta) + NG^2(\Theta) + K. \qquad (9.3.8)$$

Determination of coefficients of M, N and K let us satisfy, on the basis of two conditions: first so that at $\Theta=20°$ derivatives expressions (9.3.5) and (9.3.8) would not have gap, and, secondly, since below lines of horizon (lower than plane x-y in Fig. 9.3.1) signal cannot be received by antenna station O, into function of density distribution of probability of reception of signals on boundary with plane x-y should be introduced component in the form of function of switching on/inclusion, equal to 0.5. From (9.3.6) we have

$$-\frac{dp(G)}{dG} = \frac{1{,}47 \cdot 10^{-2}}{G^2(\Theta)} + \frac{7{,}61 \cdot 10^{-4}}{G(\Theta) \ln 10} \text{ (6) при } 1 \leqslant G(\Theta) \leqslant 10^4. \qquad (9.3.9)$$

Key: (1). with.

From (9.3.8) it follows

$$\frac{dp(G)}{dG} = M + 2NG(\Theta). \qquad (9.3.10)$$

Equating, according to first condition, (9.3.9) and (9.3.10) with value of $G(\Theta)=1$, let us find

$$M + 2N = -1{,}47 \cdot 10^{-2} - \frac{7{,}61 \cdot 10^{-4}}{\ln 10} = -1{,}473 \cdot 10^{-4}. \qquad (9.3.11)$$

Taking into account (9.3.10) and introducing, according to second condition function of switching on/inclusion, equal to 0.5 we will obtain

$$\int_{0,1}^{1} [M + 2NG(\Theta)]\, dG(\Theta) = 1 - [0{,}5 + p(G = 1)].$$

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Substituting p(G=1) from (9.3.6), let us find

$$0{,}9M + 0{,}99N + K = 0{,}5 - 1{,}47 \cdot 10^{-2} - 3{,}03 \cdot 10^{-4} = 0{,}485. \quad (9.3.12)$$

Taking into account condition of equality of expressions (9.3.6) and (9.3.8) with G(Θ)=1 (i.e. at Θ=20°), we obtain

$$M + N + K = 1{,}47 \cdot 10^{-2} + 3{,}03 \cdot 10^{-4} \approx 15 \cdot 10^{-3}. \quad (9.3.13)$$

Solving together system of equations (9.3.11)-(9.3.13), let us find

$$\left.\begin{aligned} M &= -4{,}935 \\ N &= 2{,}46 \\ K &= 2{,}49 \end{aligned}\right\}. \quad (9.3.14)$$

Thus,

$$\left.\begin{aligned} p(G) = 2{,}46G^2(\Theta) - 4{,}935G(\Theta) + 2{,}49 \quad &\text{when} \quad 20^\circ \leqslant \Theta \leqslant 180^\circ \\ &\text{or.} \quad 0{,}1 \leqslant G(\Theta) \leqslant 1 \end{aligned}\right\}. \quad (9.3.15)$$

The results of the calculations of this value are given in Fig. 9.3.2.

Thus, according to Fig. 9.3.2 it is possible to judge probability of interfering signal level at output of receiving antenna of ground station for different percentages of time of observation in random position of satellite in space and its motion by circular orbits.

For calculating weakening signals according to expression (9.3.1) value $G_{0н}(\Theta)$ should be substituted in accordance with values of Fig. 9.3.2. In this case in expression (9.3.1) values $G_{0б}(\Theta'')=1$, $\kappa_{0\,поп}=0$.

since this conclusion/output was made for the case of isotropic onboard antenna and without taking into account polarizational losses.

Interaction of equatorial satellites with ground station. In accordance with appendix 7 coordinates of the points, at which are established/installed the terrestrial and onboard antennas interacting between themselves satisfy the following expressions (without taking into account the refraction of waves):

$$\operatorname{tg} A \sin \zeta = \operatorname{tg} \varphi \tag{9.3.16}$$

$$\cos \varphi = \frac{\cos [\arccos (k \cos \beta) - \beta]}{\cos \zeta}, \tag{9.3.17}$$

$$L^2 = (R + H)^2 [1 + k^2 - 2k \cos \varphi \cos \zeta]. \tag{9.3.18}$$

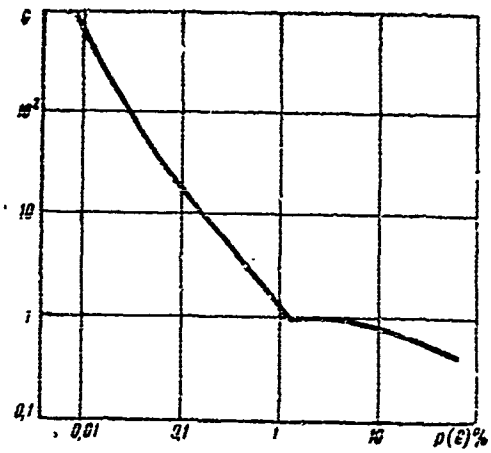

Fig. 9.3.2. Probability of changing factor of amplification of ground-based antenna for different percentage of time with circular orbits.

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Here L - distance between stations,

A - azimuth of direction to equatorial satellite for ground station,

φ - difference in longitude of points, in which is located ground station and equatorial satellite,

ζ - geographic latitude of point of ground station $k=(R/R+H)$,

R and H - radius of Earth and orbit altitude of satellite above surface of Earth respectively.

On the basis of expressions (9.3.16)-(9.3.18) it is possible to determine geographical reference of interacting stations and distance

between them, and then to calculate weakening and value of interfering signal. The antenna gain of ground station during these calculations should be determined for the direction, which with respect to the fundamental antenna bearing (to the axis of the major lobe of radiation pattern) composes the following angles: along the azimuth

$$A_0 = A - A_A, \qquad (9.3.19)$$

on, to angle of elevation

$$\beta_0 = \beta - \beta_A. \qquad (9.3.20)$$

Here A_A and β_A - azimuth and the angle of elevation of the antenna of ground station.

It is obvious that for purposes of decrease of interferences it is necessary so to orient antennas of ground stations, in order to value A_A and β_A as greatly as possible differed from analogous values, entering expressions (9.3.16) and (9.3.17).

Amplification of onboard antenna must be located for direction, determined by straight line, that connect point, at which at moment of time in question is located satellite, with point, where ground station is located. Its coordinates are characterized by latitude ζ and longitude/length, which differs from the longitude/length of satellite by value φ.

It is obvious that in general case during motion of satellite will change position of plane of polarization of interfering oscillations/vibrations, and also values H, φ, A, β, L and antenna

gain in direction of propagation of interfering signal. A change in the interfering signal level in the time with the repetition period, equal to the circling time of satellite along the equatorial orbit, will be the consequence of this. The calculation of a change in the interfering signal level in the time can be carried out for the specific cases upon consideration of the given formulas.

In the case of stationary satellite of value H, φ, A and β they will be constant/invariable in time and therefore interfering signal level at input of receivers will remain constant.

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Calculated relationships/ratios for the communication system, which uses stationary satellites, can be obtained of (9.3.16)-(9.3.18), if we consider k=0.151, and R+H=42400 km. In this case with β=0 without taking into account the phenomenon of refraction are obtained the very simple expressions:

$$\cos A \cong 0{,}151 \operatorname{tg} \zeta \qquad (9.3.21)$$

$$\cos \varphi = \frac{0{,}151}{\cos \zeta} \qquad (9.3.22)$$

$$L \cong 42 \cdot 10^3 \text{ км}. \qquad (9.3.23)$$

From (9.3.21) and (9.3.22) it follows that with β=0 with stationary satellites they can interact, creating interferences, ground station, which are located within limits of latitudes $\zeta = \pm 82°20$.

Thus, one stationary Earth satellite, with width of radiation pattern of receiving antenna of approximately 17° (see Section 2.2), can prove to be that subjected to interfering effect of very large

number of ground stations.

Is correct reverse/inverse position - stationary satellite, which has transmitting antenna with width of radiation pattern approximately 17°, can create interferences with those ground stations, whose coordinates and azimuth of major antenna lobe satisfy expressions (9.3.21) and (9.3.22) or more exact expressions (9.3.16) and (9.3.17). Let us again focus attention on the fact that all these expressions are approximate, since in them the phenomenon of refraction is not considered.

9.4. Possibility of applying one and the same frequencies in different communication systems, which use ISZ [artificial earth satellite].

Question about work of several communication systems through ISZ in one and the same frequency bands is very important, since its solution determines economical utilization of frequency bands. It is completely obvious that in this case the supplementary interferences with both the radio relay lines and with communication systems through ISZ can arise.

Let us examine first group of satellites, moving relative to Earth and which relate to one system, in which, for guaranteeing continuity of connection/communication, switching antennas (see Fig. 1.1.3) is fulfilled. In this system the part of the satellites can

provide the connection/communication between the stations, which are found on one territory (for example, in the western hemisphere of the Earth), whereas another part of the satellites will be utilized for the realization of the connection/communication between the terrestrial stations, arranged/located on another territory. With the transit of satellites switching the antennas of terrestrial stations is realized. It is obvious that in the described case all satellites can work both in the overall frequency band and in different.

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Another case - group of satellites, which relate to different communication systems. An essential difference in this case from the first is the fact that for the satellites of different communication systems different orbital parameters will be selected, whereas for the satellites of one communication system it is more expedient to utilize identical orbits and completely specific distribution of satellites in the space. Therefore the probability of the simultaneous entry/incidence of the communication satellites of different systems into the space, which is located in the region of the major lobe of the radiation pattern of the receiving or transmitting antennas terrestrial station, will prove to be greater. Upon the simultaneous entry/incidence of several satellites into the ray/beam of one antenna is unavoidable the onset of interferences, especially in that case, if satellites work in one frequency band.

Thus, although both that and other cases of work of satellites to

some frequencies lead to identical conclusions about possibility of interferences, nevertheless probability of interferences in first case is less.

One should note also that communication systems, in which stationary satellites are distant from each other up to angular distance, exceeding width of radiation pattern of major lobe of terrestrial antennas, can work in one and the same frequency bands without considerable interferences.

For evaluation/estimate of order of magnitudes let us find permissible angular distance δ between stationary satellites C_1 and C_2, counting antenna radiation pattern of symmetrical in space. Let stations A and Д (Fig. 9.4.1) support between themselves communication, utilizing the stationary satellite C_1. Therefore to side C_1 the major lobe of the antenna radiation pattern of station Д with the angle of elevation β is directed.

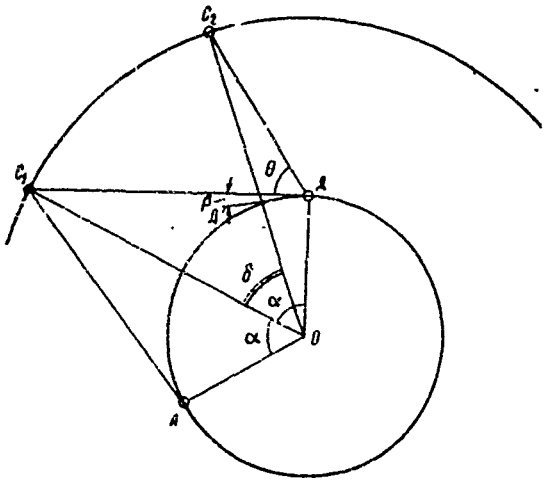

Fig. 9.4.1. To determination of permissible angle between stationary satellites.

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We will consider that the maximum antenna gain of terrestrial station Д comprises $G_{\text{макс}}$ (dB), but for antijamming from satellite C_2 the necessary difference in the amplification of this antenna in direction C_1 and C_2 must comprise not less than R, dB. Thus, it is necessary to determine angle δ from the condition that to the satellite C_2 are directed the minor lobes of the antenna of terrestrial station with the level, which differs to value $G_{\text{макс}}-R$, dB from the amplification of isotropic antenna.

From Fig. 9.4.1 we have

$$\measuredangle C_1ДC_2 = \Theta.$$

Line Д'Д is horizontal to point Д; therefore:

$$\sphericalangle Д'ДО = \frac{\pi}{2},$$

$$\sphericalangle C_2ДО = \sphericalangle Д'ДО + \beta + \sphericalangle C_1ДC_2 = \frac{\pi}{2} + \Theta + \beta,$$

$$\sphericalangle OC_2Д = \pi - \sphericalangle C_2OД - \sphericalangle C_2ДО = \frac{\pi}{2} - (\alpha - \delta + \beta + \Theta).$$

Utilizing theorem of sines for $OC_2Д$, after conversions let us find

$$\cos(\alpha - \delta + \beta + \Theta) = \frac{OД}{OC_2}\cos(\beta + \Theta) + \frac{R}{R+H}\cos(\beta + \Theta). \quad (9.4.1)$$

Considering $\beta=5°$, on the basis of expressions (2.2.1) and (2.2.2) we will obtain $\alpha=76°20'$. Accepting maximum antenna gain equal to 50 dB, and the necessary difference in the antenna gain in direction C_1 and C_2, equal to R=45 dB, on the basis Fig. 3.3.7 let us find $\Theta=14°$. After substitution in (9.4.1) we will obtain the value of the unknown angle $\delta=13°40'$.

If we under the same conditions consider that amplification of terrestrial antenna is 60 dB, then value of $\delta=5°$. In the case, when antenna gain is equal to 50 dB, and difference in the amplification in directions C_1 and C_2 must be 30 dB, value of $\delta\cong3°$.

Let us note that value of angle δ, found from expression (9.4.1), is approximate, since in given examination refraction and errors in orientation of antennas is not considered. Nevertheless the given calculations show that the use of one and the same frequency bands for different stationary satellites is possible with their diversity up to the appropriate angular distances.

9.5. Coefficient of weakening interference.

Coefficient of weakening of interference (KOP), determined by expression (9.1.3), shows, how much varies ratio of average/mean power of useful signal P_c to power of interfering $P_м$ with simultaneous passage of these signals from input of radio receiving equipment to output of communication channel.

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During the transmission of multichannel telephone signal value $Q_{вых}$, entering in (9.1.3), is defined at the output of telephone channel at point with the zero relative level as the ratio of the power of experimental tone (1 mW) to the psophometric power of the interfering signal.

Calculation of KOP can be brought to determination of change in relation of energy spectra of useful and interfering signal at input of receiver and output of channel. During the determination of value κ for purposes of simplification in the analysis the thermal noises at the input of receiver is not considered; at the same time it is always intended that inequality (9.1.4) is fulfilled. This means that during the determination of KOP the conditions are observed:

$$\left.\begin{aligned} P_{свх} &\gg P_{мвх} \\ U_{свх} &\gg U_{мвх} \end{aligned}\right\}. \qquad (9.5.1)$$

where $U_{свх}$ and $U_{мвх}$ - voltage of the fundamental and interfering signal on the input of receiver. During the determination of KOP it is assumed that width of band of receiver exceeds the width of the

spectrum of the interfering signal; therefore the selective elements of receivers are not considered. Thus, in accordance with (9.5.1) we will count

$$\varkappa = \frac{Q_{вых}}{Q_{вх}} = \frac{P_{свых}}{P_{мвых}} : \frac{P_{свх}}{P_{мвх}}. \qquad (9.5.2)$$

Ratio of average/mean power of signal to average/mean output power of interference can be expressed through relation of spectral power of signal and interference. For the determination of the latter should be preliminarily determined the correlation function of the total oscillation/vibration, formed by the fundamental and interfering signal. The computation of correlation functions and spectral power of total signal for some special cases of ChM oscillations/vibrations is carried out in the series/row of the articles, to the bases from which one should relate [9.6, 9.7, 9.8, 9.9]. The results, obtained in these articles, show that with the reception of fundamental signal the effect of the interfering signal can be considered as interaction between the spectra of these signals and interaction of carriers. Especially noticeably interference of interference precisely in second case [9.7, 9.8]. In connection with this for decreasing interferences with ChM it is necessary to prevent the possibility of the concentrated emission of energy in the narrow bands.

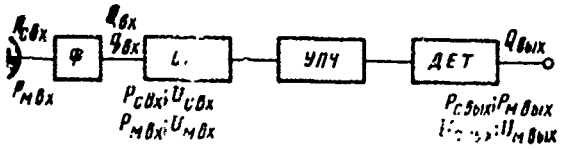

Fig. 9.5.1. To determination of KOP.

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For this it is necessary to resort to the special measures (see Section 9.7), which ensure scattering (dispersion) for carrier output in the wide frequency spectrum in such cases, when the modulating voltage (for example, with the decrease of a number of transmitted conversations) is absent. Let us note that for the communication systems through ISZ of MKKR [International Radio Consultative Committee] he recommends to introduce the dispersion of the carrier output frequency and therefore the normalization of radiated power is conducted for the frequency band, equal to 4 kHz (see Section 9.1).

Since soon in communication systems through ISZ and in majority of main-line radio relay lines will be utilized ChM oscillations/vibrations, cases of interaction of fundamental and interfering signal with frequency modulation are of greatest interest.

Frequency modulation of oscillations of both signals. In this case the calculation of KOP according to the formulas, of the enumerated above articles, leads to the very cumbersome calculations, but the tables and the graphs/curves, given in these works, do not

encompass some cases of interferences, which are obtained with the work in one frequency band of RRL and communication systems, which use ISZ.

Therefore below for calculations of KOP in the case of ChM oscillations are given simplified formulas, based on work [9.11], utilized in documents of MKKR [9.1, 9.12].

Simplification in these formulas is connected in essence with the fact that with frequency modulation of oscillations for communication systems through ISZ are accepted sufficiently large indices of modulation, with which distribution of power of ChM oscillations along spectrum corresponds to Gaussian distribution.

With frequency modulation of fundamental and interfering signal, when for radio relay lines and communication systems through ISZ spectral distribution of power is subordinated to normal law, value of KOP can be determined according to expression

$$x_\Phi = -10\lg \frac{F_к^2 \Delta F_к}{W \Delta f_к^2 \Delta f_з 2\sqrt{2\pi}} \left[\exp\left(-\frac{(f_р - F_к)^2}{2\Delta f_з^2}\right) + \exp\left(-\frac{(f_р + F_к)^2}{2\Delta f_з^2}\right) \right] (дБ) \qquad (9.5.3)$$

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Here $F_к$ - frequency in group spectrum of fundamental received signal, MHz.

$\Delta F_к = 0{,}0031$, MHz - width of band of telephone channel.

W - coefficient of predistortions (see page 191).

$\Delta f_к$ - effective deviation of frequency with transmission of experimental tone along primary system (not interfering), MHz.

$$\Delta f_з^2 = \Delta f_{с\,ср}^2 + \Delta f_{м\,ср}^2, \qquad (9.5.4)$$

$\Delta f_{с\,ср}$, $\Delta f_{м\,ср}$ - effective deviation of frequency during the transmission of the examined number of telephone channels along the primary system of communications (index "c") and on that interfering (index "m"), determined in accordance with expression (7.3.7) for $\Delta f_{ср}$ (MHz).

$f_р$ - separation between carrier frequencies of systems (MHz) in question.

In the case, when value of separation between carriers $f_р = 0$, expression (9.5.3) are simplified. In this case the calculation of KOP for $F_к = F_2$ can be carried out according to the expression

$$x_0 = 10\lg \frac{809\,\Delta f_к^2 W_з \Delta f_з \exp\left(\dfrac{F_2^2}{2\Delta f_з^2}\right)}{F_2^2}. \qquad (9.5.5)$$

Here $W_з$ - coefficient of predistortions for upper channel with frequency F_2; $\Delta f_к$, $\Delta f_з$ F_2 - in MHz.

Calculations according to expression (9.5.5) show that value x_0 for communication systems through ISZ is determined by values, given in Table 9.5.1 and virtually it does not depend on number of channels of RRL, if it is small.

If in primary and interfering system value of predistortions is selected by such, that frequency modulation of oscillations proves to be identical phase modulation, then calculation of value of KOP must be carried out through following expressions [9.12]:

$$x_0 = 15 + 10\lg\frac{S}{J}, \qquad (9.5.6)$$

$$S = 3\frac{\Delta f_{\text{ccp}}^2}{F_2^3}, \frac{1}{2u}, \qquad (9.5.7)$$

$$J = Z_1 + Z_2, \frac{1}{2u}. \qquad (9.5.8)$$

Last expression relates to case, when dispersion of energy of carrier is used. If there is no dispersion of energy of carrier, then value

$$J = 0{,}5\exp[-(D_c + D_{\text{м}})]. \qquad (9.5.9)$$

Table 9.5.1.

Число каналов в системе связи через ИСЗ (1)	около 240 (2)	300	600	1200
x_0, дб (3)	33,5	32,4	29,7	28,0

Key: (1). Number of channels in the communication system through ISZ. (2). about. (3). dB.

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Remaining values, entering given formulas, take following form:

$$Z_1 = \frac{1}{2}\frac{\exp(-D_м)}{\sqrt{2\pi}\Delta f_{ccp}}\left\{\exp\left[-\frac{(f_p - F_к^2)}{2\Delta f_{ccp}^2}\right] + \exp\left[\frac{-(f_p + F_к)^3}{2\Delta f_{ccp}^2}\right]\right\}, \quad (9.5.10)$$

$$Z_2 = \frac{1}{2\sqrt{\pi}}\frac{\exp(-D_м)D_м}{2F_{2м}}\left\{\int_a^b \exp(-x^2)\,dx + \int_c^d \exp(-x^2)\,dx\right\}, \quad (9.5.11)$$

$$D_м = 3\frac{\Delta f_{мср}^2}{F_{2м}^2};\ D_c = 3\frac{\Delta f_{ccp}^2}{F_{2c}^2};$$

$$a = \frac{-f_p - F_{2м} + F_к}{\sqrt{2}\Delta f_{ccp}};\ b = \frac{-f_p + F_{2м} + F_к}{\sqrt{2}\Delta f_{ccp}};$$

$$c = \frac{-f_p - F_{2м} - F_к}{\sqrt{2}\Delta f_{ccp}};\ d = \frac{-f_p + F_{2м} - F_к}{\sqrt{2}\Delta f_{ccp}}.$$

where $F_{2м}$ - frequency of upper channel in group spectrum of interfering system,

$F_к$ - average frequency of channel in group spectrum of primary communication system.

In Appendix 8 in the form of tables are given results of calculations of KOP for communication systems through ISZ and for RRL, carried out according to expression (9.5.6) with different number of channels in these systems and different detuning between carrier

frequencies. Data of these calculations for the communication systems through ISZ coincide with values of Table 9.5.1. Laboratory measurements showed that values κ_0, given in Appendix 8, differ from those measured not more than by 4-6 dB for the upper channels and they coincide well for lower channels [9.13]. In these measurements the fact that during the removal/taking of modulation in systems RRL the interference in the upper part of the group spectrum (534-550 kHz) grew/rose, focuses attention, this contradicts theoretical examination. With the separation of the carrier frequencies the removal/taking modulation led to reduction in the interference to the value less than the calculated.

Let us note that with substitution of value κ_0 into expression (9.1.4) should be taken smallest value κ_0, found for specific case in question, since precisely smallest value κ_0 in accordance with (9.1.4) will correspond to worst case.

During transmission of television program determination of value of KOP is given in work [9.14] for case, when interference is created by system of communications, which uses ISZ, during transmission of telephone conversations along 1200 channels with effective value of deviation of frequency of 7.5 MHz. The results of calculation κ_0 for the television channel of RRL with the band 5 MHz taking into account of predistortions and weighing filter are given in Fig. 9.5.2 in the form the curve A.

From the examination by this curve it follows that with a change of the separation between carriers f_p within the limits to ±10 MHz value of KOP changes little. Curve Б in Fig. 9.5.2 characterizes calculated change of KOP for the channel of chromaticity, which occupies band from 3.5 to 5.5 MHz.

In the case, when interfering signal from ISZ is not modulated, but is realized dispersion of energy of carrying, KOP radio relay communication system, along which transmission of television is realized, it is characterized by curves B (channel 0-5 MHz) and Г (channel of chromaticity with band 3.5-5.5 MHz). In this case it was accepted that the interfering signal has the uniform frequency spectrum, which lies on 400 kHz on both sides from the carrier. The examination of the curve Г of Fig. 9.5.2 leads to the conclusion that in the narrow-band interfering system and with the separation between carriers on the order of 4.5 MHz, which corresponds to the middle of the band of the channel of chromaticity, value of KOP sharply is reduced, which can lead to an increase in the interferences.

Code-pulse modulation. The theoretical relationships/ratios, which determine the appearance of noises or errors in the channels of system with KIM from the interfering signal in the various forms of modulation it is located in the stage of investigations. Approximately in this case it is possible to consider that the appearance of errors in channels with KIM with the four-phase carrier

modulation is determined by the dependence Fig. 9.5.3 [9.1]. During calculations it was accepted that the interferences have Gaussian distribution.

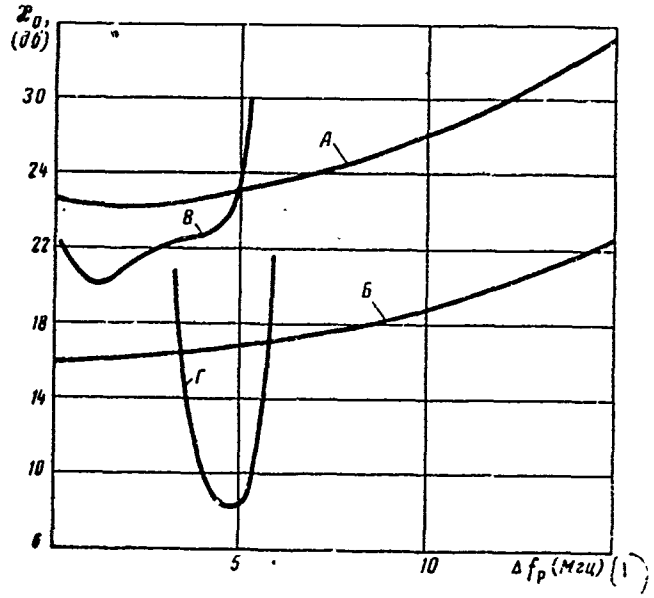

Fig. 9.5.2. Dependence $\varkappa_0$ during transmission of television on separation between carriers.

Key: (1). MHz.

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For obtaining the errors of order 10^{-4} according to Fig. 9 5.3 should be had a value $Q_{вх}=11$ dB; however, as experiments are shown, it is expedient to have a reserve not less than 3 dB. Thus, for obtaining the errors not more than 10^{-4} it is necessary to have a value $Q_{вх}=14$ dB. This value $Q_{вх}$ there must not exceed the assigned percentage of time in the course of the month.

Calculation of interferences from systems with angle modulation proves to be very complicated. Some special cases are given in works [9.15, 9.16].

Systems with OBP. Interferences with system with OBP can be calculated, comparing the spectral power density of the interfering signal, undertaken in the band of the corresponding channel of system with OBP, with the power of the received signal of system with OBP, undertaken in the same frequency band.

Interferences from system with OBP to frequency-modulation system approximately can be calculated, considering signal from system with OBP as thermal noise. Thus, adding the power of the signal of system with OBP to the thermal noises, which affect the input of receiver, it is possible to obtain the value of total noises in appropriate telephone channels according to expressions (7.3.1) and (7.3.2).

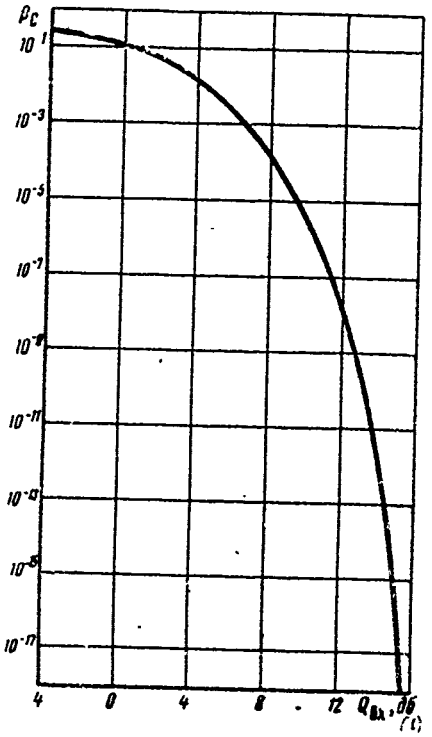

Fig. 9.5.3. Dependence of error in binary systems on value $Q_{вх}$

Key: (1). dB.

9.6. Determination of coordination distance.

In Section 9.1 it was noted that for determination of coordination distance expression (9.1.6) can be used. Substituting (9.2.1) in (9.1.6), we will obtain the general/common expression in the form

$$B_{0м} = Q_{03} - P_{0п} + B_{02} - \varkappa_0 + P_{0пм} - B_{01} + \varkappa_{0пол} - B_{0f_1} + G_{03}(\Theta') + G_{0м}(\Theta''). \quad (9.6.1)$$

Here value B_{02} is determined in accordance with (3.1.2) or (3.1.4), value $\varkappa_0$ is located according to Section 9.5, and remaining values are determined in Section (9.1) and (9.2).

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One should note that if values κ_0 and $Q_{0з}$ are determined for band with width of 4 kHz, power of transmitters $P_{0ш}$ and $P_{0м}$ also must be related to band 4 kHz. During calculations of coordination distance for the system, by which are transmitted telephone signals, should be considered expression (9.1.7).

After determination according to expression (9.6.1) of value using graphs/curves Fig. 9.2.2-9.2.6, it is possible to find coordination distance.

Let us examine determination of coordination distance for case of interferences with terrestrial station at frequency of 4 GHz from radio relay line with given value $Q_{0з}=40{,}5$ dBo during 0.03% of any month at unweighted power of noises in band of telephone channel. During calculations let us accept the values, given in Table 9.6.1.

Taking into account that on terrestrial station act three stations RRL nonsimultaneously, we obtain, that value $Q_{0з}$ from one station must be observed for time, half less than given one, i.e., for 0.01%. Since the antenna gain in 0.1% of time of any month, weakening signal must obtained be found in 0.3% of time; however, due to the absence on Fig. 3.2.2 of corresponding curve we will determine coordination distance, using curve for 0.1 % time.

Table 9.6.1.

Параметры (1)	Величина (2)
(3) Число радиорелейных станций, создающих помехи	3
(4) Мощность, излучаемая каждой станцией РРЛ в сторону земной станции,—максимальная. Поэтому, согласно рекомендации МККР (см. разд. 9.1) $P_{0пм} + G_{оп}(\theta'')$	55 *дбвт* (5)
(6) Мощность бортового передатчика	18 *дбвт* (5)
(7) Затухание между передатчиком и входом земного приемника согласно (7.7.5) $B_{0\Sigma}$	127,2 *дб* (8)
(9) Затухание мешающего сигнала в фидерах и фильтрах передатчика РРЛ и приемника земной станции B_{0f_1}	3 *дб* (8)
(10) Усиление антенны земной станции в направлении радиорелейных станций в течение 0,1% времени любого месяца $G_{оз}(\theta')$	20 *дб* (8)
(11) Поляризационные потери κ_0 пол	3 *дб* (8)
(12) Потери за счет экранирования местности B_{0i}	4 *дб* (8)
(13) Число телефонных каналов, передаваемых по радиорелейной линии	600
(14) Число каналов системы связи через ИСЗ	600
(15) Разнос несущих частот Δf_p	0
(16) Частота группового спектра, на которой определяется $B_{0м}$	2 *Мгц* (17)

Key: (1). Parameters. (2). Value. (3). Number of radio-relay stations, which create interferences. (4). Power, emitted by each station RRL in direction of terrestrial station, maximum. Therefore, according to the recommendation of MKKR (see Section 9.1) .... (5). dBW. (6). Power of onboard transmitter. (7). Attenuation between transmitter and input of terrestrial receiver according to (7.7.5) .... (8). dB. (9). Fading interfering signal in feeders and filters of transmitter RRL and receiver of terrestrial station .... (10). Antenna gain of terrestrial station in direction of radio-relay stations during 0.1% of time of any month .... (11). Polarizational losses .... (12). Losses due to shadowing of locality/terrain B_{0i}. (13). Number of telephone channels, transmitted by radio relay line. (14). Number of channels of communication system through ISZ. (15). Separation of carrier frequencies:. (16). Frequency of group

spectrum, at which it is determined:. (17). MHz.

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Determination of κ_0 must be conducted according to expressions (9.5.3), (9.5.5) or (9.5.6). Considering it possible for the present instance to use (9.5.6), in accordance with appendix 8 (Tabl. П. 8.2) we have κ_0=30.3 dB.

After substitution of enumerated values into expression (9.6.1) we find for 0.1% of time

$$B_{0м} = 183{,}4 - 3{,}6 \cong 180 \text{ dB}.$$

Value - 3.6 dB is determined by transition/junction from one band 4 kHz to the next 3.1 kHz according to (9.1.7) and by account of psophometric filter.

In accordance with graphs/curves Fig. 9.2.4 in terms of obtained value $B_{0м}$ can be determined value $L_{мин} \cong 170$ km, which will be coordination distance. In the case, if between RRL station and terrestrial station route will pass not along one zone, but on two, value L (km) must be determined according to the graphs/curves Fig. 9.2.5-9.2.7. Analogous with the given example can be designed coordination distances, also, for other cases.

9.7. Dispersion of power along the spectrum.

In Section 9.1 it was noted that for purposes of reduction in interferences by recommendations of MKKR for communication systems, which use ISZ, is normalized radiated power or density of power flux in frequency band, equal to 4 kHz. Let us examine conditions, with which can be fulfilled this recommendation in the case of the communication systems with the frequency modulation of oscillations.

During transmission of telephone conversations level of multichannel signal, by which is realized frequency modulation of oscillations, can vary in considerable limits - from maximum, which corresponds to busy hour, to zero - when conversations are absent. The latter leads to the decrease of the deviation of frequency, the reduction of the occupied frequency band, and consequently, to an increase in the power of oscillations at the carrier frequency. Since with the frequency modulation the power of the emitted oscillations is constant and does not depend on amplitude or spectrum of the modulating signal, change in the occupied frequency band will lead to the fact that the power of ChM oscillations, in reference to any band of frequencies (for example, to the frequency band with a width of 4 kHz), will change in the dependence on the value of charging. Therefore value $P_{\text{изм}}$ in expression (9.1.5) can substantially increase.

This will lead to decrease $Q_{0\,\text{изм}}$ to inadmissibly high values of interferences.

During transmission of television image is obtained somewhat different position. In this case the exchange of the subject of image leads to a change in the spectrum of the modulating signal, and consequently, to a change in width of band of ChM oscillations. However, even in the absence of picture signal (for example, during the transmission only of the level of white or level of black) TV signal contains the signal of synchronization, which consists of the pulses of synchronization and extinguishing pulses/momenta, which realize a change in the frequency in accordance with Fig. 9.7.1a, i.e., within the limits from the frequency ω_1 (corresponding to white level), to ω_3 (determining the transmission of the apexes/vertexes of timing pulses). The same figure gives the values of some parameters of TV signal. Let us note that according to the standard, accepted in the USSR, the repetition period of line timing pulses, the duration of the extinguishing pulses and their clock frequency are equal to the following values:

$$\left.\begin{aligned} T &= 64 \text{ мксек}^{(1)} \\ \tau &\cong 11 \text{ мксек}^{(1)} \\ F &= \frac{\Omega}{2\pi} = \frac{1}{T} = 15625 \text{ гц}^{(2)} \end{aligned}\right\}. \qquad (9.7.1)$$

Key: (1). μs. (2). Hz.

Let us find value of components of oscillation, frequency-modulated, with transmission of level of white and the presence only of line pulses/momenta. Since change the phases φ and the frequency ω are connected with relationship/ratio $\varphi = \int \omega dt$, on the

basis Fig. 9.7.1 it is possible to record

$$\left.\begin{aligned}\varphi &= \omega_2 t && \text{(1) при } 0 < t < 0{,}1\tau\\ \varphi &= \omega_3 t + 0{,}1\tau(\omega_3 - \omega_2) && \text{(2) при } 0{,}1 < t < 0{,}55\tau\end{aligned}\right\}. \quad (9.7.2)$$

$$\left.\begin{aligned}\varphi &= \omega_2 t + 0{,}5\tau(\omega_3 - \omega_2) && \text{(1) при } 0{,}55\tau < t < \tau\\ \varphi &= \omega_1 t + 0{,}5\tau(\omega_2 + \omega_1 - 2\omega_1) && \text{(1) при } \tau < t < T\end{aligned}\right\}. \quad (9.7.2)$$

Key: (1). with.

This change in phase is given in Fig. 9.7.1b. ChM oscillation can be recorded in the form

$$u = U_0 \sin\varphi. \quad (9.7.3)$$

Representing this oscillation in the form of series/row

$$u = U_0 \sum_{\kappa=1}^{\infty} C_\kappa \sin(\omega t + \varphi_\kappa),$$

it is possible to find values C_κ and φ_κ:

$$C_\kappa = \sqrt{a_\kappa^2 + b_\kappa^2}, \quad \operatorname{tg}\varphi_\kappa = \frac{a_\kappa}{b_\kappa}. \quad (9.7.4)$$

$$\left.\begin{aligned}a_\kappa &= \frac{1}{T}\int_0^T \sin\varphi\cos\kappa\Omega t dt,\\ b_\kappa &= \frac{1}{T}\int_0^T \sin\varphi\sin\kappa\Omega t dt\end{aligned}\right\} \quad (9.7.5)$$

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This method of determining the components ChM of signal is not complex, but it requires the high expenditure of time for calculations. If we take into account that during the transmission of video signals in the communication systems through ISZ [ИСЗ - artificial earth satellite] and in RRL the spread/scope of a change in the frequency composes several MHz, then for the determination of the components of ChM oscillation/vibration, modulated in accordance with Fig. 9.7.1a, it is possible to use simpler and more demonstrative method.

This method is based on the Blackman-Middleton theorem [9.18]: energy spectrum of ChM oscillation with very large indices of modulation in the case of transmission of stationary, determined or serrated signals is proportional to probability density of modulating process.

In case in question, if ω_1-ω_3 (see Fig. 9.7.1a) it composes several MHz, and parameters of determined modulating signal are determined by values (9.7.1), index of modulation will be very large. The probability density $\omega_1(v)$ of the modulating signal v with the repetition period T in accordance with Fig. 9.7.1a comprises

$$\left.\begin{aligned} \omega_1(v) &= \frac{T-\tau}{T} \quad \text{(1) при } \tau \leqslant t < T \\ \omega_1(v) &= \frac{0{,}55\tau}{T} \quad \text{(1) при } 0 \leqslant t \leqslant 0{,}1\tau \text{ (2) и } 0{,}55\tau \leqslant t \leqslant \tau \\ \omega_1(v) &= \frac{0{,}45\tau}{T} \quad \text{(1) при } 0{,}1\tau \leqslant t \leqslant 0{,}55 \end{aligned}\right\}. \qquad (9.7.6)$$

Key: (1). with. (2). and.

Using the Blackman-Middleton theorem, it is possible to say that energy spectrum of oscillation examined/considered by ChM will be described by three components, proportional to probability densities, which are determined according to (9.7.6).

In accordance with this of relation of power of oscillations with frequencies of ω_1, ω_2, ω_3 they will be

$$P(\omega_1):P(\omega_2):P(\omega_3) = \frac{T-\tau}{T} : \frac{0{,}55\tau}{T} : \frac{0{,}45\tau}{T}.$$

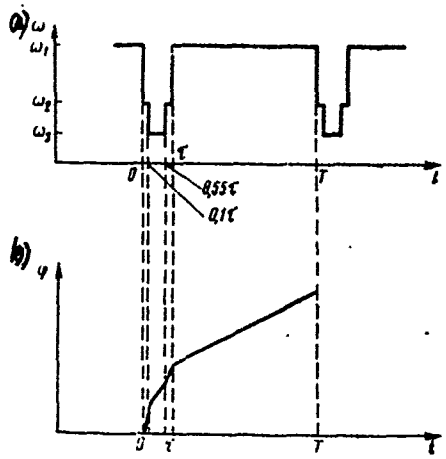

Fig. 9.7.1. Change in frequency (a) and phase (b) of oscillation, frequency-modulated by TV signal.

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With the substitution into this expression of the parameters of those determined (9.7.1), we will obtain

$$P_1(\omega_1):P(\omega_2):P(\omega_3)\approx 0{,}83:0{,}092:0{,}078. \qquad (9.7.7)$$

Fig. 9.7.2 gives energy spectrum of oscillation examined/considered by ChM, which corresponds to relations (9.7.7).

Let us note that from (9.7.7) it follows that power at frequency ω_1 in comparison with power of unmodulated oscillations is reduced into number of times, equal to 1/0.83=1.2 (0.8 dB). With other words, the transmission of white level and of line pulses/momenta in accordance with Fig. 9.7.1 does not lead to considerable reduction in the amplitude of the oscillations of the carrier frequency, and, consequently, is not created the uniform distribution of the power of

oscillations along the spectrum. Similar conclusions can be drawn also for the case, when it is realized the transmissions of black level, and also during the transmission of line and frame timing pulses.

Consequently, during transmission of TV signal, just as during transmission of multichannel signals, for satisfaction of condition for uniform distribution of power along spectrum it is necessary to introduce special signals of dispersion. This can be done with the help of the special device/equipment, which subsequently we will call dispersive. Let us consider the principle of the work of dispersive device/equipment ДУ on the functional diagram in Fig. 9.7.3.

In Fig. 9.7.3 it is shown that dispersive device/equipment is switched on between equipment for multiplexing АУ and ЧМ transmitter and consists of analyzers of group spectrum АС, amplifiers У and signal generator of dispersion ГСД. Depending on amplitude and character of the group spectrum, with the help of the analyzer of spectrum $АС_1$ is regulated the amplification $У_1$, through which is passed the special signal of dispersion, created by the generator ГСД.

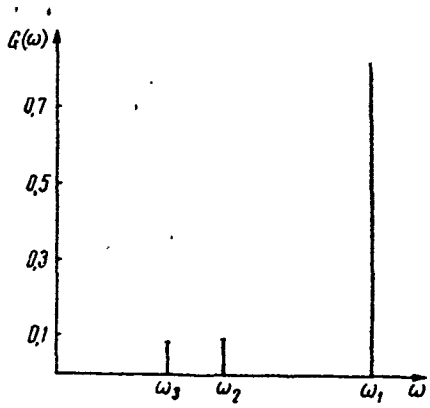

Fig. 9.7.2.

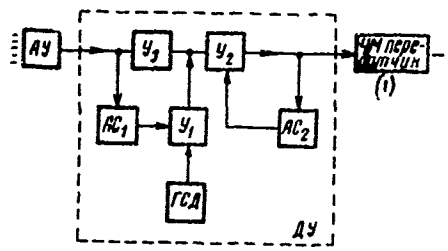

Fig. 9.7.3.

Fig. 9.7.2. Energy spectrum with ChM modulation by signal of synchronization.

Fig. 9.7.3. Schematic of dispersive device/equipment.

Key: (1). ChM transmitter.

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Thus, at the input of amplifier $У_2$ can be obtained the total signal, which consists of the signal, taken from the output of equipment for multiplexing, and the signal of dispersion, whose parameters are chosen so that at the output of ChM transmitter independent of a change in the signal level, taken from the equipment for multiplexing, would be obtained the uniform (or close to the uniform) distribution of the power of ChM oscillation along the spectrum.

At the output of amplifier $У_2$ an analyzer of spectrum $АС_2$, which regulates amplification of $У_2$, is connected.

Thus, by gain control $У_1$ is automatically established/installed necessary signal level of dispersion, and by change in amplification of $У_2$ is ensured the setting of necessary for deviation frequency by total signal. The presence of the adjustable amplifier $У_2$ makes it possible to support more uniform spectral power density even in such a case, when in the circuit $АС_1$-$У_1$ there is a certain "threshold", due to which with a small change in the charging the amplifier $У_1$ can prove to be still closed. One should note that ДУ can contain only the part of the blocks, given in Fig. 9.7.3. In this case there can be the following versions of construction of ДУ:

1. In the absence of $АС_1$ and $АС_2$ the constant supply of the signal of dispersion with the constant/invariable level is realized. In this case a certain simplification of ДУ unavoidably leads to a change in the group signal level and, consequently, also to a change in the spectral power density of ChM signal.

2. In the absence of $АС_1$, $У_1$, and ГСД with change in charging with the help of $АС_2$ will occur change in factor of amplification of $У_2$, as a result of which deviation of frequency at the output of transmitter can be maintained by constant. However, this version of diagram cannot be used under the conditions, when charging can prove to be equal to zero, and also during the transmission of such television images, in which little is changed the brightness. It is obvious, also that the transition/junction from the lower channels to the upper, or vice versa, will lead to a change in the spectral power

density of ChM oscillation. It is necessary to say that a change in the amplification of $У_2$ will produce level change at the output of channel and therefore it will require the appropriate automatic adjustment also at the reception/procedure.

3. In the absence of $АС_2$ the uniformity of spectrum on output of transmitter in process of work can somewhat be changed, especially if start of amplifier $У_1$ by analyzer $АС_1$ is realized after decrease of charging before certain "threshold value".

Let us note that together with versions of diagram of ДУ examined can be proposed others.

Let us now move on to comparison of several signals of dispersion from point of view of guarantee of uniformity of power in spectrum of ChM oscillation. Comparison let us lead for the cases, when as the signal of dispersion is used sinusoidal oscillation, noise signal with the normal distribution and the repetitive pulses of triangular form.

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During this comparison should be rated/estimated possibility of eliminating signal of dispersion at point of reception/procedure with the help of any methods, for example, by filtration or by introduction of special compensating signals.

Signal of dispersion in the form of sinusoidal oscillation. In this case in the absence of charging and during the supplying only of the signal of dispersion with the frequency $\Omega=2\pi F$ of ChM oscillation is determined (5.3.5), i.e.

$$u = U_0 \cos[\omega_0 t + m \sin \Omega t].$$

Here $m = \frac{\Delta\omega_m}{\Omega} = \frac{\Delta f_m}{F}$ is index of modulation, and $\Delta\omega_m=2\pi\Delta f_m$ and $\omega_0-2\pi f_0$ are respectively amplitude value of deviation of angular frequency and carrier angular frequency.

If m>>1, then the width of band Δf, occupied by ChM oscillation, is equal about to

$$\Delta f \cong 2\Delta f_m = 2mF.$$

In this case those appearing in process of modulation components are located from each other in frequency on value F.

Converting (5.3.5) to form

$$u = U_0 [\sin \omega_0 t \cos (m \sin \Omega t) + \cos \omega_0 t \sin (m \sin \Omega t)]$$

and using formulas for Bessel functions:

$$\cos (x \sin \varphi) = I_0 (x) + 2 \sum_{n=1}^{\infty} I_{2n} (x) \cos 2n \varphi,$$

$$\sin (x \sin \varphi) = 2 \sum_{n=0}^{\infty} I_{2n+1} (x) \sin (2n + 1) \varphi,$$

$$\cos (x \cos \psi) = I_0 (x) + 2 \sum_{n=1}^{\infty} (-1)^n I_{2n} (x) \cos 2n \psi,$$

$$\sin (x \cos \psi) = 2 \sum_{n=0}^{\infty} (-1)^{(n)} I_{2n+1} (x) \cos [(2n + 1) \psi],$$

let us find

$$u = U_0 \left\{ I_0(m) \sin \omega_0 t + \sum_{\kappa=1}^{\infty} I_\kappa(m) [\sin(\omega_0 + \kappa\Omega) t + (-1)^\kappa \sin(\omega_0 - \kappa\Omega) t] \right\}. \quad (9.7.8)$$

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From this expression it follows that ChM oscillation consists of an infinite number of components, distant behind each other in the frequency to the value, equal to frequency of the modulating. The energy spectrum of this process also consists of the separate components, the exemplary/approximate form of envelope of which is given in Fig. 9.7.4 for different indices of modulation [9.18]. As it follows from this figure, with an increase in the index of modulation the distance between the maximum values of the envelope of energy spectrum is expanded. With the large indices of modulation the envelope of energy spectrum in the first approximation, is characterized by following values [9.18]:

$$\left.\begin{aligned} G(\omega) &= \frac{U_0^2}{\Omega}\left[m^2 - \left(\frac{\omega_0 - \omega}{\Omega}\right)^2\right]^{-\frac{1}{2}} && \text{(1) при } |\omega_0 - \omega| < m\Omega \\ G(\omega) &= \infty && \text{(1) при } |\omega_0 - \omega| = m\Omega \\ G(\omega) &= 0 && \text{(1) при } |\omega_0 - \omega| > m\Omega \end{aligned}\right\}. \quad (9.7.9)$$

Key: (1). with.

From Fig. 9.7.4 and expressions (9.7.9) it follows that with $m \gg 1$ on edges of band value of energy spectrum very large. This is explained by the fact that the part of the time, during which ChM oscillation with an instantaneous frequency of $\omega_0 + \Delta\omega_m \cos \Omega t$ passes any

narrow frequency band, will be considerably more in such a case, when this band is located on the edges of the region of deviation of frequency, but not in the center of region. In accordance with this the power of oscillation in this band so will be more then, when band is located nearer to the edge.

Thus, dissipation of power with the help of sine wave of dispersion will be not effective: with large indices of modulation significant part of power will be concentrated on edges of band, but with small indices of modulation - in center of band. The use of indices of modulation, close to one, also does not solve a question.

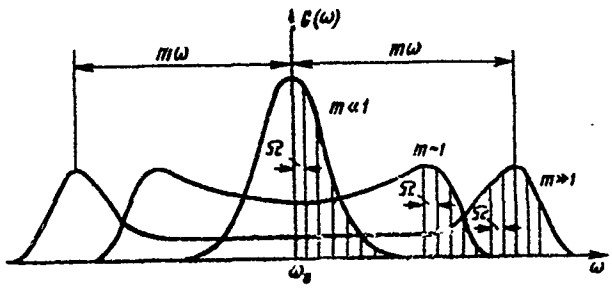

Fig. 9.7.4. Energy oscillation spectrum, modulated in frequency by sinusoid.

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The signal of dispersion in the form of limited on band thermal noise, with uniform spectrum in band from zero to frequency $F_{макс}$ if the modulation frequency with the effective index of modulation is realized by this stationary normal process,

$$m_э = \frac{\Delta f_э}{F_{макс}}, \qquad (9.7.10)$$

where $\Delta f_э = \frac{\Delta\omega_э}{2\pi}$ - effective deviation of frequency by the noise signal of dispersion, moreover

$$m \gg 1, \qquad (9.7.11)$$

that energy spectrum $G_{чм}(\omega)$ of ChM oscillation has a form of gaussian curve with the apex/vertex at point $\omega=\omega_0$ [9.19]. Thus,

$$G_{чм}(\omega) = \frac{U_0^2}{2} \frac{\sqrt{2\pi}}{\Delta\omega_э} e^{-\frac{(\omega-\omega_0)^2}{2\Delta\omega_э^2}}. \qquad (9.7.12)$$

Here U_0 - amplitude of unmodulated carrier.

Width of spectrum band of ChM oscillation with signal of

dispersion in question is determined by expression

$$\Delta f = \sqrt{2\pi}\, m_э F_{макс}. \qquad (9.7.13)$$

It is obvious that with this energy spectrum highest efficiency of ChM oscillation in frequency band, equal to 4 kHz, $P_{(4)\,макс}$ will be concentrated near the carrier frequency ω_0.

Consequently,

$$P_{(4)\,макс} = \int_{\omega_0-2\cdot 2\pi}^{\omega_0+2\cdot 2\pi} G_{чм}(\omega)\, d\omega = \frac{\sqrt{2\pi}\, U_0^2}{2\Delta\omega_э} \int_{\omega_0-4\pi}^{\omega_0+4\pi} e^{-\left(\frac{\omega-\omega_0}{\sqrt{2}\Delta\omega_э}\right)^2} d\omega. \qquad (9.7.14)$$

In this expression all frequencies are expressed in kHz. Introducing the variable $t=\frac{\omega-\omega_0}{\Delta\omega_э}$, we will obtain

$$P_{(4)\,макс} = \left[\Phi\left(\frac{4\pi}{\Delta\omega_э}\right) - \Phi\left(\frac{-4\pi}{\Delta\omega_э}\right)\right]\pi U_0^2, \qquad (9.7.15)$$

where

$$\Phi(z) = \frac{1}{\sqrt{2\pi}} \int_0^z e^{-\frac{x^2}{2}}\, dx. \qquad (9.7.16)$$

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Taking into account that $|z|<1$, and also that in this case

$$\Phi(-z) = 1 - \Phi(z),$$

in accordance with (9.7.15) we will obtain

$$P_{(4)\,макс} = \left[2\Phi\left(\frac{2}{\Delta f_э}\right) - 1\right]\pi U_0^2. \qquad (9.7.17)$$

Let us note that value $\Phi\left(\frac{2}{\Delta f_э}\right)$ can be found according to tables of probability integrals, and the value U_0^2 characterizes amplitude

of oscillation at carrier frequency in the absence of frequency modulation.

It is of interest to determine power change in band 4 kHz, situated symmetrically relative to carrier frequency, in the case, when modulation is realized only by multichannel communication/report or only signal of dispersion in the form of noise. We will consider in this case that the deviation of frequency, created by the signal of dispersion, is ten times less than during the supplying of multichannel signal, and the peak factors of each signal are identical.

Taking into account that effective deviation of frequency, created by multichannel signal, is determined by unbroken curves of Fig. 7.3.6 and, consequently, is satisfied condition $\Delta f_{cp} > F_2$ or $m_0 > 1$, where F_2 - maximum frequency of group spectrum, it is possible to cons.der that during transmission of multichannel signal energy spectrum has form of gaussian curve. Therefore, taking into account (9.7.17), it is possible to record the following expression for the relation of power in the frequency band with a width of 4 kHz during the transmission of multichannel signal $P_{(4)N}$ or during the transmission of the signal of dispersion in the form of thermal noise $P_{(4)m}$:

$$\frac{P_{(4)\text{ш}}}{P_{(4)N}}=\frac{2\Phi\left(\frac{2}{\Delta f_{\text{э}}}\right)-1}{2\Phi\left(\frac{2}{\Delta f_{\text{ср}}}\right)-1}. \qquad (9.7.18)$$

Here $\Delta f_{\text{э}}$ and $\Delta f_{\text{ср}}$ must be substituted in kHz.

If we consider that the peak factor of noise signal and multichannel signal are identical, and band, occupied by ChM oscillation with modulation by signal of dispersion in the form of thermal noise is ten times less than band with frequency modulation by multichannel signal, then in expression (9.7.18)

$$\Delta f_{\text{э}}=0{,}1\,\Delta f_{\text{ср}}.$$

Therefore

$$\frac{P_{(4)\text{ш}}}{P_{(4)N}}=\frac{2\Phi\left(\frac{20}{\Delta f_{\text{ср}}}\right)-1}{2\Phi\left(\frac{2}{\Delta f_{\text{ср}}}\right)-1}. \qquad (9.7.19)$$

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Calculations according to expression (9.7.19) with values of Δf, given in Fig. (7.3.6), show that with N>50 relation $\frac{P_{(4)\text{ш}}}{P_{(4)N}}\cong 9$ of times (9.5 dB) and does not depend on number of telephone channels. Thus, during the introduction of the signal of dispersion in the form of the thermal noise, whose band composes 10% of the band of multichannel signal, power in the band 4 kHz in the absence of charging will exceed power in the hours of maximum charging approximately on 9.5 dB, if a number of channels is more than 50.

Signal of dispersion in the form of pulses/momenta of triangular form (saw-tooth voltage). In this case, the instantaneous value of frequency is changed in accordance with Fig. 9.7.5a, that in the analytical form it is possible to record in the form

$$\omega = \frac{\omega_1 + \omega_2}{2} + \frac{\omega_2 - \omega_1}{T} t. \qquad (9.7.20)$$

Here T - pulse repetition period, and $-T/2) \leq t \leq (T/2)$.

Introducing designations of average/mean value of frequency and spread/scope of frequency

$$\left.\begin{aligned} \frac{\omega_1 + \omega_2}{2} &= \omega_0 \\ \omega_2 - \omega_1 &= \Delta\omega_p \end{aligned}\right\}, \qquad (9.7.21)$$

let us find change in phase of oscillation

$$\varphi = \int_0^t \omega\, dt = \omega_0 t + \frac{\Delta\omega_p}{2T} t^2. \qquad (9.7.22)$$

This expression in the form of graph is shown in Fig. 9.7.5b. Consequently, for the oscillation, modulated in the pulse frequency of triangular form (saw-tooth voltage), we have

$$u = U_0 \sin\left(\omega_0 t + \frac{\Delta\omega_p}{2T} t^2\right).$$

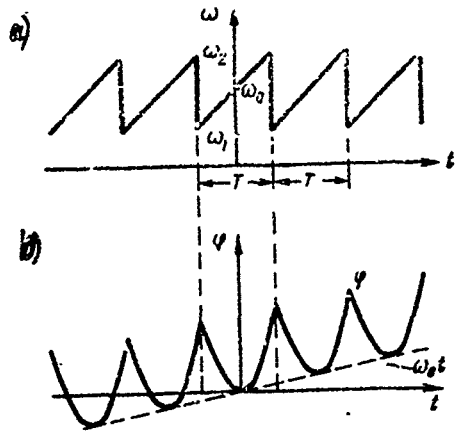

Fig. 9.7.5. Change in frequency (a) and phase (b) of oscillations, modulated in frequency by saw-tooth voltage.

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Let us substitute this oscillation in the form of sum of components; then amplitude of component with number "k"

$$u_к = U_0 \sqrt{a_к^2 + b_к^2}. \tag{9.7.23}$$

Here

$$\left.\begin{aligned} a_к &= \frac{1}{T}\int_{-\frac{T}{2}}^{+\frac{T}{2}} \sin\left(\omega_0 t + \frac{\Delta\omega_0}{2T} t^2\right)\cos к\Omega t dt \\ b_к &\frac{1}{T}\int_{-\frac{T}{2}}^{+\frac{T}{2}} \sin\left(\omega_0 t + \frac{\Delta\omega_p}{2T} t^2\right)\sin к\Omega t dt \end{aligned}\right\}. \tag{9.7.24}$$

Representing products of trigonometric functions, which stand under integral in the form of sum, for coefficient $a_к$ we will obtain

$$\sin\left(\omega_0 t+\frac{\Delta\omega_p}{2T}t^2\right)\cos\kappa\Omega t=\frac{1}{2}\left[\sin\left(\alpha t+\gamma^2 t^2\right)+\sin\left(\beta t+\gamma^2 t^2\right)\right]=$$
$$=\frac{1}{2}\left\{\sin\left[\left(\frac{\alpha}{2\gamma}+\gamma t\right)^2-\frac{\alpha^2}{4\gamma^2}\right]+\sin\left[\left(\frac{\beta}{2\gamma}+\gamma t\right)^2-\frac{\beta^2}{4\gamma^2}\right]\right\}=$$
$$=\frac{1}{2}\left[\sin\left(\frac{\alpha}{2\gamma}+\gamma t\right)^2\cos\frac{\alpha^2}{4\gamma^2}-\cos\left(\frac{\beta}{2\gamma}+\gamma t\right)^2\sin\frac{\beta^2}{4\gamma^2}\right]. \quad (9.7.25)$$

Here

$$\alpha=\omega_0+\kappa\Omega;\ \beta=\omega_0-\kappa\Omega;\ \gamma=\frac{\Delta\omega_p}{2T}.$$

Analogous relationships/ratios can be obtained for product of integrands in expression for b_κ. After integration [9.10] we will obtain:

$$a_\kappa=\frac{1}{2T\gamma}\left(\cos\frac{\beta^2}{4\gamma^2}C_y+\sin\frac{\beta^2}{4\gamma^2}S_y-\cos\frac{\alpha^2}{4\gamma^2}C_x-\sin\frac{\alpha^2}{4\gamma^2}S_x\right), \quad (9.7.26)$$
$$b_\kappa=\frac{1}{2T\gamma}\left(\cos\frac{\beta^2}{4\gamma^2}S_y-\sin\frac{\beta^2}{4\gamma^2}C_y+\cos\frac{\alpha^2}{4\gamma^2}S_x-\sin\frac{\alpha^2}{4\gamma^2}C_x\right). \quad (9.7.27)$$

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Here

$$\left.\begin{aligned} C_x&=\int_{x_1}^{x_2}\cos x^2dx;\quad C_y=\int_{y_1}^{y_2}\cos y^2dy\\ S_x&=\int_{x_1}^{x_2}\sin x^2dx;\quad S_y=\int_{y_1}^{y_2}\sin y^2dy;\end{aligned}\right\} \quad (9.7.28)$$
$$x_1=\frac{1}{2}\left(\frac{\alpha}{\gamma}-\gamma T\right);\ y_1=\frac{1}{2}\left(\frac{\beta}{\gamma}-\gamma T\right);$$
$$x_2=\frac{1}{2}\left(\frac{\alpha}{\gamma}+\gamma T\right);\ y_2=\frac{1}{2}\left(\frac{\beta}{\gamma}+\gamma T\right). \quad (9.7.29)$$

After substitution (9.7.26) and (9.7.27) in (9.7.23) we will obtain

$$u_\kappa=\frac{1}{2T\gamma}\left\{C_y^2+S_y^2+C_x^2+S_x^2-2\cos\frac{\beta^2+\alpha^2}{4\gamma^2}(C_yC_x-S_yS_x)-\right.$$
$$\left.-2\sin\frac{\beta^2-\alpha^2}{4\gamma^2}(C_yC_x+S_yS_x)\right\}^{\frac{1}{2}}. \quad (9.7.30)$$

Values C and S, in this expression - Fresnel's integrals.

Fig. 9.7.6 shows character of change in value of amplitude of different components in accordance with expression (9.7.30); hence follows that greatest values of amplitude have on edges of band, in limits of which is realized deviation of frequency. With an increase in the deviation of frequency the difference between the maximum and minimum values of the amplitude of oscillations is reduced. In the limit, when the index of modulation

$$m = \frac{0{,}5\Delta\omega_p}{\Omega} \to \infty, \qquad (9.7.31)$$

the difference between the maximum and minimum values of amplitudes becomes equal to zero, i.e., all components have identical amplitudes. By other words, with $m \to \infty$ energy spectrum becomes uniform and it can be described by following expression [9.18]:

$$\left.\begin{aligned} G(\Omega) &= \frac{U_0^2\pi}{2\Omega m} = \frac{U_0^2\pi}{\Delta\omega_p} = \frac{U_0^2}{2\Delta f_p} \text{ при } 0 \leqslant |\omega - \omega_p| < \frac{\Delta\omega_p}{2} \\ G(\Omega) &= 0 \text{ при } |\omega - \omega_0| > \frac{\omega_p}{2} \end{aligned}\right\}. \qquad (9.7.32)$$

Key: (1). with.

Here

$$\Delta f_p = \frac{\Delta\omega_p}{2\pi}.$$

One should note that, as show experiments [9.20], with values of

m>100 it is possible to consider that the energy spectrum is described by expression (9.7.32). In this case the power of oscillations, created on the single resistor/resistance in the band of frequencies of 4 kHz, will be determined by the expression

$$P_{(4)} = 4\frac{U_0^2}{2\Delta f_p} = \frac{2U_0}{\Delta f_p}, \qquad (9.7.33)$$

where Δf_p in kHz.

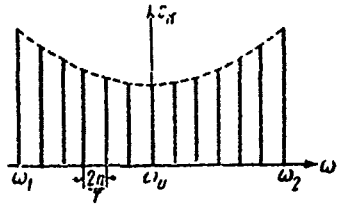

Fig. 9.7.6. Dependence of amplitude of components with frequency modulation by saw-tooth voltage.

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Thus, effective dispersion of carrier output along spectrum by saw-tooth wave form can be realized with values of m>100.

Let us lead comparison of power emitted in band 4 kHz, during maximum charging of communication system by multichannel signal with case, when in the absence of charging is introduced signal of dispersion in the form of pulses/momenta of triangular form with $m \to \infty$ and with such spread/scope of change in frequency, with which band, occupied by this signal, composes 10% of band of multichannel signal.

Counting the peak factor of multichannel signal equal to $\varkappa = 12$ dB (4 times), we obtain, that peak deviation of frequency by multichannel signal will compose $\Delta f_{\text{п}} = 4\Delta f_{\text{ср}}$. Consequently, according to the condition accepted the spread/scope of a change in the signal frequency of dispersion must compose

$$\Delta f_{\text{р}} = 2\left(\frac{4\Delta f_{\text{ср}}}{10}\right) = 0{,}8\Delta f_{\text{ср}}. \qquad (9.7.34)$$

Therefore relation of power in band of frequencies of 4 kHz upon consideration (9.7.33), (9.7.34) and expressions of similar (9.7.17) will take form

$$\frac{P_{(4)}}{P_{(4)N}} = \frac{1}{0,4\pi\Delta f_{cp}\left[2\Phi\left(-\frac{2}{\Delta f_{cp}}\right)-1\right]}. \qquad (9.7.35)$$

Calculations according to this expression show that independent of number of transmitted telephone channels

$$\frac{P_{(4)}}{P_{(4)N}} = 0,6 \text{ once } (-2.0 \text{ dB}).$$

Thus, with identical frequency bands use of saw-tooth voltage as signal of dispersion gives best results, than use of thermal noises, limited along spectrum. In this case, however, is necessary a good form of serrated signal, since the disturbance/breakdown of linearity and r unding the apexes/vertexes of saw-tooth voltage can lead to the power increase $P_{(4)}$.

One should note that removal/distance of signal of dispersion from group spectrum in place of reception/procedure can be carried out by method of compensation. For this at the point of reception/procedure must be generated the local signal of the same form as the signal of the dispersion (or the signal of dispersion it must be transmitted by one of the channels of the line of communications) and then be introduced in the antiphase for the

compensation.

Compensation can be realized both on video spectrum and in intermediate frequency, via modulation of heterodyne in signal frequency, opposite on phase to signal of dispersion.
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Latter/last method is more preferable, since it makes it possible to decrease the band of frequencies of receiver in the case of the continuous transmission of the signal of dispersion.

During transmission of TV signals saw-tooth voltage can be introduced into those time intervals, during which are transmitted rows (image). This introduction of the signals of dispersion during the transmission of white level is shown in Fig. 9.7.7. Possibly also the introduction of the signals of dispersion with the repetition frequency of frames/personnel.

It was previously noted that effective dispersion of power with the help of saw-tooth wave form was possible in the case of m>100. This condition during the transmission of TV signal means that the spread/scope of the frequency deviation, created by saw-tooth voltage with the repetition period of rows must be not less

$$\Delta f_p > 2m\frac{1}{T} = 2\cdot 100\frac{1}{64\cdot 10^{-6}} \sim 3 \text{ MHz}.$$

Consequently, if spread/scope of change in frequency, created by complete TV signal, including synchronizing pulses, will exceed 43 MHz, then maximum level of saw-tooth voltage, shown in Fig. 9.7.7, will correspond to one tenth of video signal, i.e., one gradation of brightness. With this spread/scope of a change in the frequency the presence of the saw-tooth wave form of dispersion will lead to a smooth change in the brightness from the left to the right side of the kinescope to one gradation, i.e., in effect it will not be noticed. This you mean that even the partial compensation for the signal of the dispersion in the place of reception/procedure will lead to the satisfactory results.

Let us note that in selection of repetition frequency of saw-tooth waves form of dispersion to equal frequency $F_{кд}$=50 Hz, even obtaining m>10^4 will correspond to spread/scope frequency, equal to

$$\Delta f_p > 2mF_{кд} = 2 \cdot 10^4 \cdot 50 = 1 \text{ MHz}.$$

This means that in the absence of compensation for signal of dispersion change in image brightness on vertical line to one gradation will be observed, if spread/scope of change in frequency by TV signal is approximately 14 MHz.

Let us note that selection of signal of dispersion in the form of triangular pulse/momentum is feasible also during transmission of multichannel telephone signals. In this case the duration and the repetition period of sawtooth pulses must be determined from the condition so that their spectrum would be concentrated in the frequency band, which lies lower than the minimum frequency F_1 of the group spectrum.

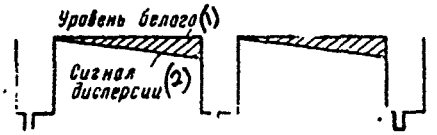

Fig. 9.7.7. Introduction of signal of dispersion during transmission of picture signal.

Key: (1). White level. (2). Signal of dispersion.

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One should note that nonlinear distortions in circuit of transmission (nonlinearity of amplitude characteristic of group amplifiers, phase-frequency characteristic of UPCh, etc.) they can lead to the increase of noises in the communication channels, caused by the distortions of the signal of dispersion. In the case of the transmission of television program the nonlinearity of circuit can lead to beatings of the components of video signal and signal of dispersion.

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10.3. Antenna systems of terrestrial stations.

Requirements, which are imposed on antennas of terrestrial stations of communication systems through ISZ, reduced to obtaining of great directive gains [KND], which correspond to amplification factor of 45-55 dB and more, on low side-lobe level. At the same time the construction/design of antenna must allow for the possibility of continuous tracking of ISZ, i.e., it must have the appropriate rotary devices/equipment. The latter must realize displacement/movement of antenna in the space, either over the previously comprised program or over the maximum value of the signal adopted with ISZ. It is obvious that the second method can be directly realized on the receiving antennas, from which the data, which characterize the direction of receiving antenna to the satellite, can be transmitted to the system, the manager of the motion of the transmitting antenna. One should note that during the transfer of these data in them the corresponding corrections, which consider both the certain territorial separation of the receiving and transmitting antenna, and their structural/design nonidentity, must be introduced.

Let us note that coincidence of functions of reception and

transmission in by one antenna to system is very expedient, since this to a considerable ext t makes it possible to shorten initial costs of terrestrial stations.

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Nevertheless this association/unification is hindered by the creation of appropriate separation filters, which ensure, first of all, maximally low losses in the forward directions (receiver-antenna; transmitter-antenna) and, secondly, as great as possible crosstalk attenuation from the transmitter to the receiver. Furthermore, during the use of a common antenna for the transmission and the reception usually the length of feeder is obtained greater than in the case of the separate antennas: this leads to additional power losses of signal and increase in the noises at the input of receiver.

At present at terrestrial stations are utilized several types of antennas: parabolic, horn-parabolic, antennas of Cassegrain, etc.

In spite of the fact that terrestrial antenna systems must work in the range of frequencies 3.4-8 GHz (see appendix 3), their sizes/dimensions, for guaranteeing high amplification factors, they prove to be sufficiently large. This to a considerable degree impedes the creation of the rotary devices/equipment, which ensure the displacement/movement of antenna in the space with the high accuracy.

Tentative sizing of antennas according to given values of factor of amplification and frequency can be carried out on the basis of expressions and graphs/curves, led in sections. 3.3. Representation about sizes/dimensions and appearance of antennas they can give Fig. 10.3.1-10.3.6. The data about the radiation patterns of some antennas are cited in Fig. 3.3.5 and 3.3.6.

Let us examine now question about rotary devices/equipment of antenna systems.

Fig. 10.3.1. Parabolic antennas of communication system Molniya-1 Moscow- Vladivostok.

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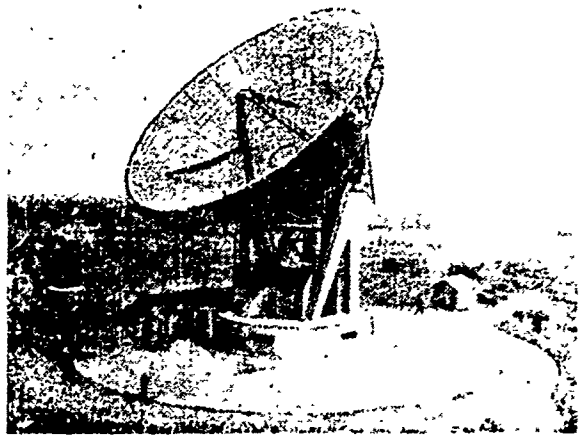

Fig. 10.3.2. General view of station and antenna in Goonhilly Downs [10, 14].

Fig. 10.3.3. Ground station in Nutley [10, 14].

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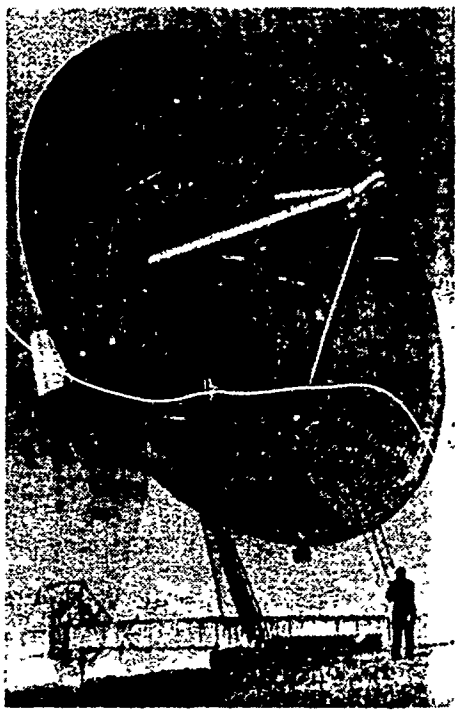

Fig. 10.3.4. Parabolic antenna in Holmdel [6].

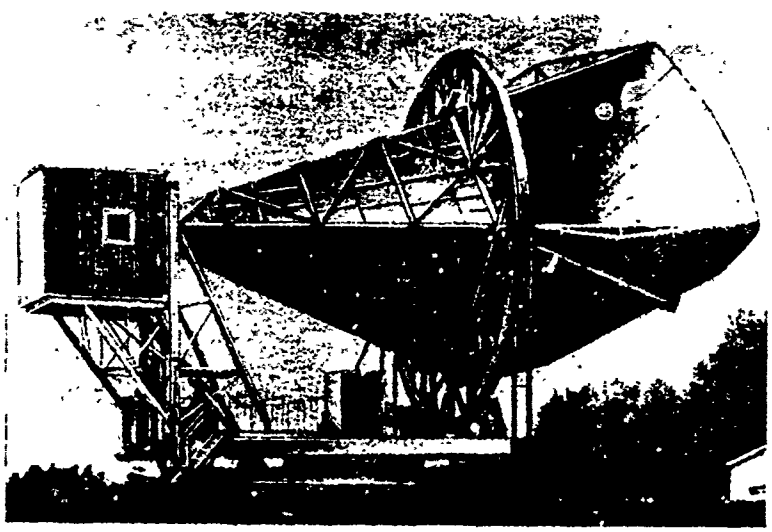

Fig. 10.3.5. General view of station in Holmdel [6].

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So that antenna during motion of satellite would be always directed to its side, at present are utilized two types of rotary devices/equipment: azimuthal-elevation, with which rotation of antennas is realized around x and z axes (see Fig. 10.3.7a), and the system, in which the antenna rotates around x and y axes (see Fig. 10.3.7b).

Analysis of work of elevation system [10.13, 10.14] shows that with tracking satellite angular antenna scan rate ω_z relative to azimuthal z axis will increase with increase in angle of elevation.

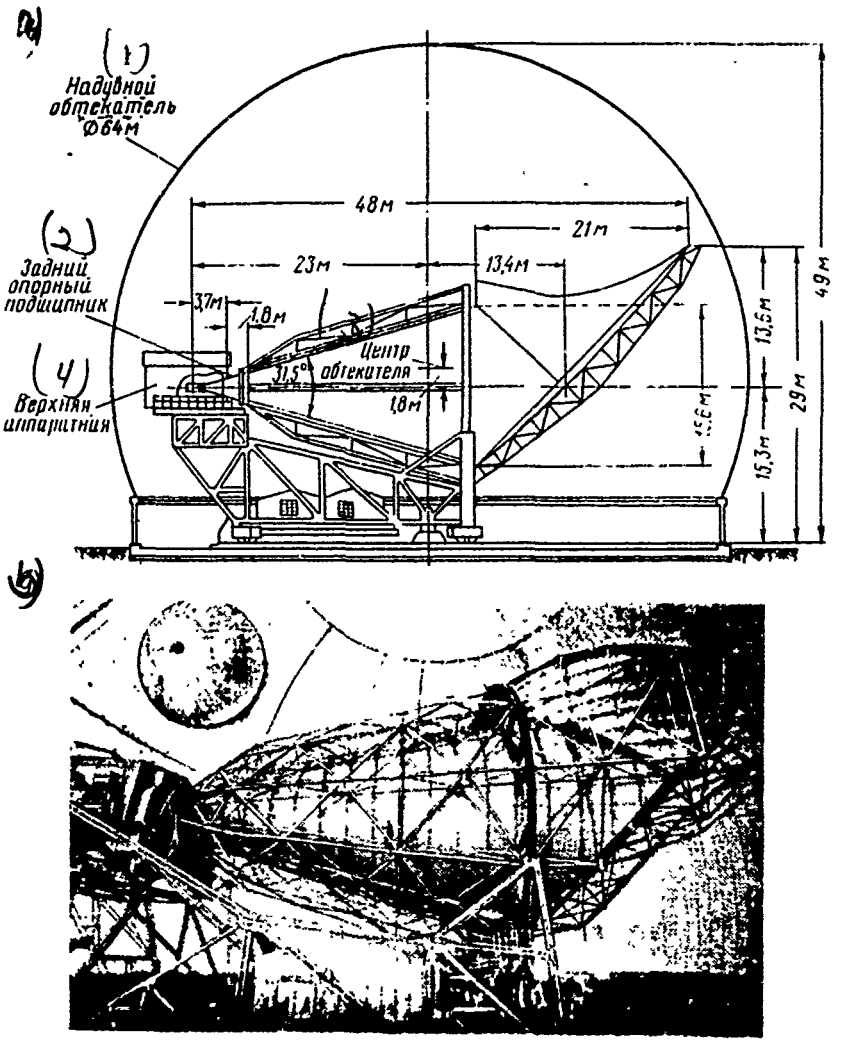

Fig. 10.3.6. Horn-parabolic antenna in Andover: a) diagram; b) general view [10.1].

Key: (1). Inflatable dome. (2). Rear radial bearing. (3). Center of dome. (4). Upper apparatus.

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At the angles of elevation, which approach 90°, the angular velocity

indicated approaches very high values, moreover the large it is, the lower the orbit of satellite. This is limited the possibility of applying such rotary devices/equipment, since near the angles of elevations, whose values approach 90°, due to the limited speed of the motion of antenna it is obtained certain "dead zone", in which the tracking the satellites proves to be impossible (region a in Fig. 10.3.8).

At orbit altitude in several thousand kilometers dead zone decreases to units of degrees and less and in many instances virtually it does not have value. Elevation turning system is used in horn-parabolic antennas, Fig. 10.3.5 and 10.3.6.

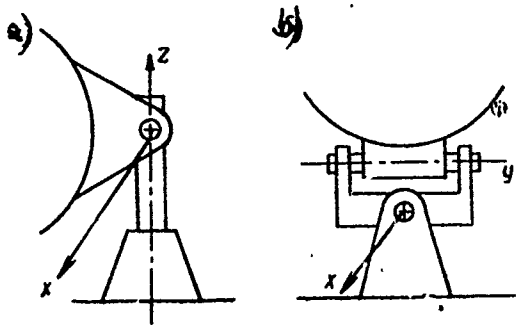

Fig. 10.3.7. Types of rotary devices/equipment of antennas: a) azimuthal-elevation (x, z), b) system of axes xy.

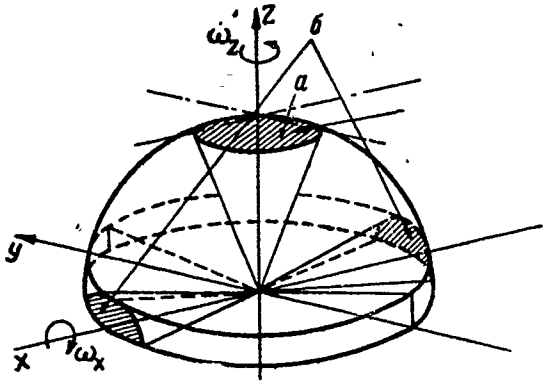

Fig. 10.3.8. Dead zones of rotary devices/equipment: a) azimuthal-elevation (x, z), b) system of axes xy.

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Rotary system with axes xy just as elevation, has dead zone. However, for system of xy dead zone is displaced into the region of small angles of increase (region b in Fig. 10.3.8), at which the communication system through ISZ usually is not utilized as a result of the considerable increase of atmospheric noises.

From comparison of rotary systems it follows also that system xy with any orbits requires only one capture of satellite near horizon/level on entire way of accompaniment, while system xz with orbits, passing through "dead zone" at zenith, it requires two captures of satellite - on horizon/level and at output from "dead zone". As a result of an unavoidable deterioration in the parameters of the communication channel in the period of capture of satellite, during the use of system xz is possible a certain loss of information, whereas in system of xy there is no repeated capture; therefore this deficiency is absent. Furthermore, system xy frees/releases from the need for the change-over of antenna around the azimuthal axis, which in system xz serves as one additional source of the appearance of a "dead zone" as a result of limited angular velocity ω_r. The "dead zone", connected with the change-over, is essential for the low-flying satellites and has low value for the satellites, which move in high orbits.

Thus, differences in systems of rotation xy, xz especially sharply are developed with motion of satellites at low altitudes. In the communication systems through ISZ, using orbit altitudes 10000 km and more, both types of rotary devices/equipment are approximately equivalent.

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12.2. Special features of work with the diversity of signals in the frequency.

Multistation work with diversity of broad-band channels at frequency (MCh) it is possible to explain Fig. 12.2.1. In this case for the transfer of each broad-band channel is utilized a specific carrier frequency - $f_1, f_2, \ldots, f_n$. The diversity between any pair of the adjacent carrier frequencies must be selected so as to exclude the mutual overlap of the spectra.

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During use MCh can be used any forms of modulation - OBP, ChM, KIM, etc.

Let us examine multiplex operation of repeater, with which is realized simultaneous reception/procedure and amplification of all broad-band channels/trunks by general receiving-transmitting equipment. It is obvious that in comparison with the mode of single operation in this case the equipment of repeater must be more broadband. Besides this, it is necessary that the levels of the signals, which arrive at different frequencies to the receiver of repeater, would be identical. The nonobservance of this condition leads to the suppression of weak signals by strong. Thus, with the

multiplex operation in the case MCh is necessary adjusting the signal level of terrestrial stations, which can be carried out by the appropriate change in the power output of the transmitter of terrestrial station. Adjusting must be realized both in accordance with the change in the distance between ISZ and this station and in accordance with a change in the parameters in the section of connection/communication Earth - ISZ (for example, as a result of inaccuracy of orientation of onboard antenna, appearance of sediments/ residues in area of terrestrial station, etc.).

With multiplex operation in systems with MCh due to nonlinearity of characteristics of receiving-transmitting equipment general/common for all broad-band channels/trunks transient noise between different broad-band channels/trunks will appear. So that the total quantity of noises at the output of the communication channels would answer the recommendations of MKKR and would be preserved by the same, as with the single work of repeater, it is necessary to ensure reduction in the value of thermal noises. This means that upon transfer from the single work of repeater to the repeated is necessary an application of antennas with the higher amplification factor or an increase in the power of the transmitters of terrestrial and onboard station and the decrease of the total equivalent noise temperature, in reference to the input of receiver.

Thus, transition/junction to multiplex operation in systems with MCh requires higher energy indices.

12.3. Special features of work with the diversity of signals in time.

In systems with MV with synchronous working at any moment of time through repeater they will pass signals, emitted only by one terrestrial station.

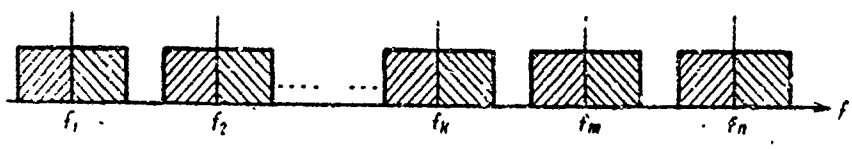

Fig. 12.2.1. Arrangement/position of spectra with MCh.

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Therefore all terrestrial stations can work at one carrier frequency, but they must have the general/common timing mechanism, which ensures the strictly by-turn/alternate switching on of transmitters.

Fig. 12.3.1. gives graph/curve of work of transmitters of three stations, which are located in points/items A, B and C when distance from these stations to the relaying repeater are indentical. According to 12.3.1, for time τ each station emits oscillations/vibrations of carrier frequency, modulated by signal, which comes from multiplexing equipment; through T_i in Fig. 12.3.1 clock period is designated, and $\tau_з$ - shielding time interval, which prevents simultaneous operation of two terrestrial stations. The described version, which relates to the case of the synchronous working of terrestrial stations, will be for the contraction designated by MVS (multistation system with the diversity in the time of the synchronously switched on stations). Possibly also realization MV with asynchronous operation of terrestrial stations, that we will designate through MVA.

In the case MVA each terrestrial station emits

oscillations/vibrations into time intervals, not synchronized with other stations. In this case the stations differ one from another from in terms of code combinations. The code can be constructed both according to the temporary/time sign/criterion and according to the frequency.

With MVA through receiving-transmitting equipment of satellite into some time intervals simultaneous possible passage of signals not of one, but several terrestrial stations. It is obvious that at MVA the receiving-transmitting equipment of satellite must possess somewhat different parameters, than with MVS, for example, the characteristics of onboard relay equipment must be such that during the simultaneous amplification of the signals of several terrestrial stations transient interferences between these signals would be absent.

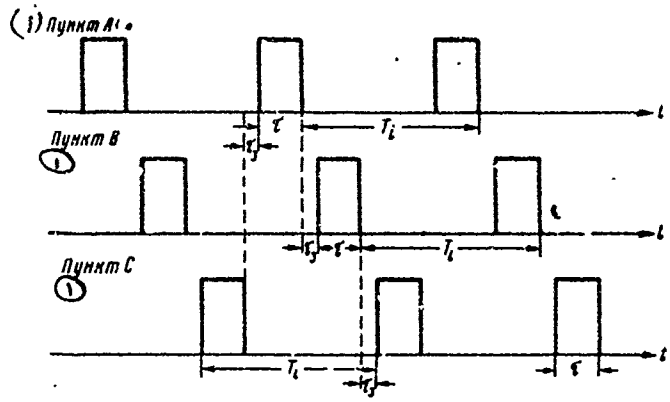

Fig. 12.3.1. Schedule of operation of points/items with MVS.

Key: (1). Point/item.

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One should note that with MVA as a result of the random combinations of the signals of different stations the spurious signals can appear. However, in spite of the difficulties, which appear during the creation MVA, they are of considerable interest, since they possess the series/row of advantages. Fundamental from them - the absence of the synchronizers of operation of terrestrial stations and, consequently, considerable simplification in the equipment.

Let us note that with any form MV (synchronous and asynchronous) modulation of transmitters of terrestrial stations can be any (amplitude, frequency and phase), but identical for all stations. The form of modulation is selected from the conditions of guaranteeing the high of noise immunity and reliability of

connection/communication. At the same time with any form MV modulation of pulses/momenta, realized in the equipment for multiplexing, also can be any (ShIM, FIM, KIM, etc.). Let us examine the case MVS in connection with KIM.

We will consider that to input of repeater from n stations modulated oscillations/vibrations with identical carrier frequencies are supplied. Each terrestrial station works in the period τ. For this time one word is emitted.

Let us find repetition period T_i of impulses/transmissions of any of stations, by assuming that from each station by one broad-band channel/trunk is transmitted group spectrum by width $\Delta F=F_2-F_1$. In this case must be satisfied the condition

$$T_i \leqslant \frac{1}{2\Delta F}. \qquad (12.3.1)$$

Since during this time successive work all n of stations must be realized, time τ, abstracted/removed for operation of any station, must satisfy condition

$$\tau = \frac{T_i}{n} - \tau_з, \qquad (12.3.2)$$

where $\tau_з$ - shielding time interval (see fig. 12.3.1.). Introducing designation

$$\tau_з = \varepsilon\tau, \qquad (12.3.3)$$

from (12.3.2) we will obtain

$$n\tau(1+\varepsilon) = T_i. \qquad (12.3.4)$$

Time of emission of station τ and necessary band of frequencies Δf are connected with relationship/ratio

$$\Delta f = \frac{A}{\tau}, \qquad (12.3.5)$$

where coefficient A is determined by shape of pulse. With substitution (12.3.5) and (12.3.1) into expression (12.3.4) we will obtain:

$$\frac{nA}{\Delta f}(1+\varepsilon) \leqslant \frac{1}{2\Delta F},$$
$$\Delta f > 2\Delta F A (1+\varepsilon) n. \qquad (12.3.6)$$

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Total frequency band, occupied by repeater for reception or transmission of oscillations/vibrations, in the case to Kims will be doubly more than value, found from (12.3.6), i.e.,

$$\Delta f_{\Sigma} \quad 4\Delta F A (1+\varepsilon) n. \qquad (12.3.7)$$

Coefficient of use of repeater η, which characterizes reduction in its capacity upon transfer from single work to repeated, is determined by fact that with multiplex operation introduction of shielding time between emissions of stations is necessary. Therefore

$$\eta_{\text{I}} = \frac{\tau}{\tau + \tau_3} = \frac{\tau}{\tau + \varepsilon\tau} = \frac{1}{1+\varepsilon}. \qquad (12.3.8)$$

Let us note that values τ_3 and ϵ are determined by requirements for permissible value of transient noises between terrestrial stations, parameters of equipment (by time constants of different circuits, etc.), by possible change in propagation time in section between terrestrial stations and by ISZ and accuracy of work of timing mechanism of stations.

Is very interesting study of problem about application in communication systems through ISZ of temporary/time systems with memory/memorizing devices, or as it is accepted to call them [12.1], systems with compression in time [12.2]. It is possible to show that the application of such systems makes it possible to improve the coefficient of the use of a repeater and to somewhat decrease the occupied frequency band.

Let us examine principle of construction of these systems, on the basis of [12.1]. We will consider that at each terrestrial station there is the memory system, because of which separate words of pulses/momenta are not emitted during the interval of time τ, but "are stored" specific for a while. As a result of this for the storage time at each station will be accumulated m of intervals by duration τ, of the forming blocks from words pulses/momenta, by which then will modulated the transmitter.

By other words, with this form of work m of time interval, for each of which in previously example examined from each station occurred emission of one word (intervals of time 1.1', 1" ... for first station, intervals of time 2.2', 2" ... for second station and so forth - see Fig. 12.3.2a), are united at each station into one

overall time interval, called the block (see Fig. 12.3.2b).

As a result of this duration of time of emission of each station will increase in m of times ($\tau_m = m\tau$), in the same number of times to increase clock period, and duration of shielding interval between operation of stations can remain previous, equal to $\tau_з$. This makes it possible to increase the coefficient of the use of a repeater and to shorten the frequency band.

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On the spot of reception/procedure blocks accepted can be divided into intervals duration τ with the aid of special equipment, after which groups of pulses, which are contained in these intervals, they are demodulated.

Taking into account designations Fig. 12.3.2 analogous with expression (12.3.2), we will obtain $\tau_m = m\tau = \frac{T_m}{n} - \tau_з$. Hence upon consideration (12.3.3)

$$m\tau = \frac{T_i m}{n} - \varepsilon\tau.$$

Thus, keeping in mind (12.3.1) and (12.3.5), let us find that during introduction to compression to time frequency band, occupied by any station,

$$\Delta f_m \geqslant 2\Delta FA\left(n + \frac{\varepsilon}{m}\right). \qquad (12.3.9)$$

Total frequency band, occupied by repeater,

$$\Delta f_з = 2\Delta f_m > 4\Delta FA\left(n + \frac{\varepsilon}{m}\right). \qquad (12.3.10)$$

Coefficient of use of repeater in this case

$$\eta_m = \frac{m\tau}{m\tau + \varepsilon\tau} = \frac{1}{1+\frac{\varepsilon}{m}} = \frac{m}{m+\varepsilon}. \qquad (12.3.11)$$

Comparing (12.3.6) and (12.3.9), and also (12.3.8) and (12.3.10), we will obtain that application of time compression (memory elements) will make it possible to shorten required frequency band and to increase coefficient of use of repeater into number of times

$$a = m\frac{1+\varepsilon}{m+\varepsilon}. \qquad (12.3.12)$$

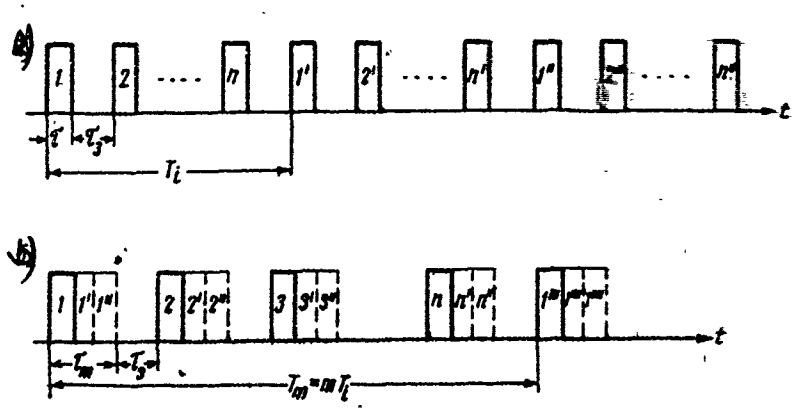

Fig. 12.3.2. Sequences of operation of stations.

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Dependence of coefficient a on m different values ϵ is given in Fig. 12.3.3.

Thus, application of system of temporary/time compression with MVS makes it possible to shorten frequency band and to improve use of repeater due to contraction of pauses. Furthermore, compression in the time makes it possible to simplify both the timing mechanism of stations and mode of their operation (station they are switched on more rarely, and they work in the more prolonged period). Especially essential results can be obtained with the large space it is memorable; however, this will lead to the complication of memory units and an increase in the cost/value of terrestrial stations.

Synchronization of terrestrial stations with MVS can be realized both on signals of world service of precise time and by

pulses/momenta, transmitted by central station of control. Equipment on ISZ can relay these pulses/momenta along the separate channel. After working/treatment at the terrestrial stations the pulses/momenta can be utilized for setting of the moments of switching on and disconnection of this station. During the determination of the moment of switching on it is necessary to consider the time of the emission of the signal from each station to ISZ so that the possibility of impulse flashing over of different stations would be excluded. This determination is complicated in the case of transient satellites and with the low value of shielding time.

Together with difficulties of system indicated with MVS they possess essential advantage over MCh, which consist in the fact that with successive operation of terrestrial stations through repeater transient noise between broad-band channels/trunks decreases, which can lead to certain improvement in signal-to-noise ratio at output of communication channel. Furthermore, with the successive work, in contrast to MCh, mutual regulation of signal levels is not required, which enter the input of repeater.

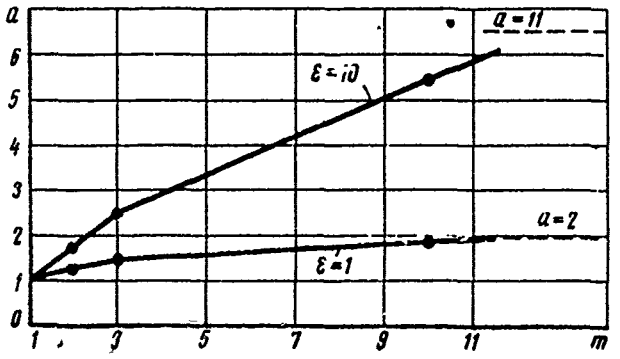

Fig. 12.3.3. Dependence of coefficient a on m.

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Application of system MVA facilitates requirements for synchronous working of ground stations, but ground equipment (coding and decoding) is more complicated and can arise transient noise between broad-band channels/trunks.

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12.2. Flood J. E. Urgnhart-Pulleu D. J. Time compression multeplex transmission. «Electronic Rec.», Apr. 1964, p. 647—668.

APPENDICES

Appendix 1.

Summary table of formulas for calculating the undisturbed motion ISZ the construction of the geocentric projections of the orbit of satellite.

During different calculations of motion of ISZ according to circular and elliptic orbits can prove to be useful given below expressions [2.1]. Designations utilized in this case are clear from Fig. 2.1.2.

Semimajor axis a

$$a = \frac{r_A + r_\Pi}{2} = \frac{r_\Pi}{|e-1|}.$$

Orbit eccentricity e

$$e=\frac{r_{\text{A}}-r_{\text{П}}}{r_{\text{A}}+r_{\text{П}}}=\frac{r_{\text{A}}}{a}-1=1-\frac{r_{\text{П}}}{a}.$$

True anomaly ;PCh

$$\vartheta=\arccos\left(\frac{\cos E-e}{1-e\cos E}\right)=2\operatorname{arc\,tg}\left(\sqrt{\frac{1+e}{1-e}}\operatorname{tg}\frac{E}{2}\right).$$

Radius-vector r

$$r=\frac{r_{\text{П}}(1+e)}{1+e\cos\vartheta}=\frac{r_{\text{A}}(1-e)}{1+e\cos\vartheta}=\frac{a|e^2-1|}{1+e\cos\vartheta}.$$

Eccentric anomaly E

$$E=2\operatorname{arc\,tg}\left(\sqrt{\frac{1-e}{1+e}}\operatorname{tg}\frac{\vartheta}{2}\right),$$

$$E=2\operatorname{arc\,sin}\left(\frac{\sqrt{2}\sin\frac{\vartheta}{2}}{\sqrt{a(1+e)}}\right)=2\operatorname{arc\,cos}\left(\frac{r\cos\frac{\vartheta}{2}}{\sqrt{a(1-e^2)}}\right),$$

$$E=\arccos\left(\frac{a-r}{ae}\right)=\arccos\frac{a-r}{a-r_{\text{П}}}=\arccos\left(\frac{r\cos\vartheta+ae}{a}\right).$$

Mean anomaly

$$\vartheta_{\text{ср}}=E-e\sin E.$$

time of motion of ISZ from one point of orbit to another

$$t_2-t_1=\frac{E_2-E_1-e(\sin E_2-\sin E_1)}{\sqrt{\frac{K}{a^3}}}.$$

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Here value K=kM. In accordance with the definitions, given for expression (2.1.1), $K=3.986\cdot10^{20}$ cm^3/s^2=$3.986\cdot10^5$ km^3/s^2.

Let us consider order of approximate construction of geocentric projection of orbit of satellite.

For construction of this projection they must be known:

- orbit altitude H (in the case of circular orbits) or height/altitude of apogee H_A and perigee $H_П$ (in the case of elliptic orbits);

- orbit inclination i;

- coordinate of projection of any point of orbit on surface of Earth (subsequently this point we will call "reference)". In the case of elliptic orbits it is expedient as the reference point to select the projection of perigee point;

- orbit eccentricity e;

- period of orbit T of satellite around the Earth;

- direction of the motion of satellite.

1. On globe, knowing orbit inclination, stretched by filament or wire, they plan section of surface of Earth by orbital plane (without taking into account rotation of earth/ground). This section must the cross line of equator at angle of i, pass through the projection of perigee (or the reference point O) and touch two parallels with coordinates $\zeta=i$ and $\zeta=-i$ (positive values ζ relate to north latitudes, negative - to south).

2. Coordinates of points, through which is passed section planned on globe as orbital plane, are transferred to ap/chart and are combined by line, called line of section.

3. For subsequent calculations we arbitrarily choose several values of eccentric anomaly E in limits from 0 to 2π. A number of such values is determined by the necessary accuracy of the construction of the projection of orbit and must not be less than 10-12. In a number of selective values E is desirable to include values $E_{\text{П}}=0$ and $E_{\text{П}}=2\pi$ (perigee point) and $E_{\text{А}}=\pi$ (apogee point). Thus, introducing in the form of indices the numbering of all selective values for the eccentric anomaly, beginning from the zero index for the perigee point, we will obtain the following arbitrarily selected values: $E_{\text{П}}=E_0=0$ (perigee); $E=E_1$; $E=E_2 \ldots E=E_{\text{А}}=\pi$ (apogee) $\ldots E=E_{\text{П}}=2\pi$. Let us number the points of orbit, which correspond to the selected values of E. Thus in accordance with the motion of satellite along the orbit we will obtain points 0, 1, 2, ..., A, ..., n.

4. For each value E_p according to given above formulas we determine value of true anomaly ϑ_p and time t_p, for which satellite will pass from perigee into each point p of orbit. If we for the zero value of time take the moment/torque of passage as the satellite of perigee, we will obtain a certain simplification in the formula. Thus, when $t_{\text{П}}=0$, $E_{\text{П}}=E_0=0$ let us find:

$$t_p=(E_p-e\sin E_p)\sqrt{\frac{a^3}{3{,}98\cdot 10^5}},\ \text{s}.$$

Here

$$a=\frac{r_{\text{А}}+r_{\text{П}}}{2}=\frac{2R+H_{\text{А}}+H_{\text{П}}}{2},\ \text{км},$$

$R=6400$ км; $H_{\text{А}}^{(1)}$ н $H_{\text{П}}^{(2)}$ в км.

Key: (1). and. (2). in.

For circular orbit for zero value of time it is expedient to take moment/torque of passage as satellite of reference point.

5. We determine geographic latitude of each point of orbit according to formula

$$\sin \zeta_p = -\sin i \cos \vartheta_p.$$

6. We compute angle, to which will turn itself Earth for time t_p (s), according to formula

$$\Lambda_з = \frac{360}{24 \cdot 3600} t_p = \frac{t_p}{240}, \text{ deg.}$$

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7. All values it is brought in into table.

(1) Номер точки орбиты	0	1	2	...	A	...	n
(2) Параметр	(3) Перигей				(4) Апогей		(5) Перигей
E	0	E_1	E_2		π		2π
ϑ	0	ϑ_1	ϑ_2		ϑ_A		$\vartheta_П=0$
t	0	t_1	t_2		$t_A = \frac{T}{2}$		T
ζ	$-i$	i_1	i_2		$+i$		$-i$
$\Lambda_з$	0	θ_1	θ_2		θ_A		$\theta_П$

Key: (1). Number of the point of orbit. (2). Parameter. (3). Perigee. (4). Apogee.

8. On line of section, constructed on map/chart according to p. 2, knowing geographic latitude ζ in accordance with tables p. 7 let us

plot points 0, 1, 2, ..., A, ..., П.

9. We shift/shear each point on line of section to appropriate value A_3 in accordance with table, filled out according to p. 7. The shift/shear of points is realized along the parallels of latitude from right to left. Thus, in accordance with the table, carried out on p. 7, will be obtained points 1', 2', ..., A', ..., П', moreover 1 and 1' will be apart by value Θ_1 degrees from each other, and points A and A' - to value Θ_A degrees from each other.

10. Points 0, 1', 2', ..., m', ..., Π' we combine by line, which will be geocentric projection of orbit of satellite taking into account rotation of Earth.

11. For construction of following section of projection of orbit should be repeated p. 1, after taking for reference perigee point П', and then to repeat paragraphs 2-10.

Appendix 2.

Derivation of formula to Section 2.3.

Fig. П.2.1 shows part of orbit of satellite, whose plane has angle of slope i to equatorial plane. Considering that orbit altitude above the surface H Earth, a radius R Earth, from the examination Δ Obc we have

$$Ob = R + H.$$
$$bc = z = (R + H)\sin\zeta. \qquad (\Pi.2.1)$$

From the examination Δ Oba

$$ab = Ob\sin\alpha = (R + H)\sin\alpha.$$

From Δ abc follows

$$bc = z = ab\sin i = (R + H)\sin\alpha\sin i. \qquad (\Pi.2.2)$$

Equalizing (П.2.1) and (П.2.2), we will obtain

$$\sin\zeta = \sin\alpha\sin i. \qquad (\Pi.2.3)$$

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With change in angle α on α+dα we will obtain respectively change in angle ζ to angle of ζ+dζ. In this case instead of expression (П.2.3) we will obtain

$$\sin(\zeta + d\zeta) = \sin(\alpha + d\alpha)\sin i.$$

After conversions of latter/last expression, by taking into account that values dζ and dα are low and therefore

$$\sin d\zeta = d\zeta;\ \cos d\zeta = 1;\ \cos d\alpha = 1;\ \sin d\alpha = d\alpha,$$

we will obtain

$$\sin i\cos\alpha\, d\alpha = \cos\zeta\, d\zeta + \sin\zeta - \sin\alpha\sin i. \qquad (\Pi.2.4)$$

Taking into account (П.2.3), we find

$$d\alpha = \frac{\cos\zeta}{\sin i\cos\alpha}\, d\zeta = \frac{\cos\zeta}{\sin i\sqrt{1-\sin^2\alpha}}\, d\zeta \qquad (\Pi.2.5)$$

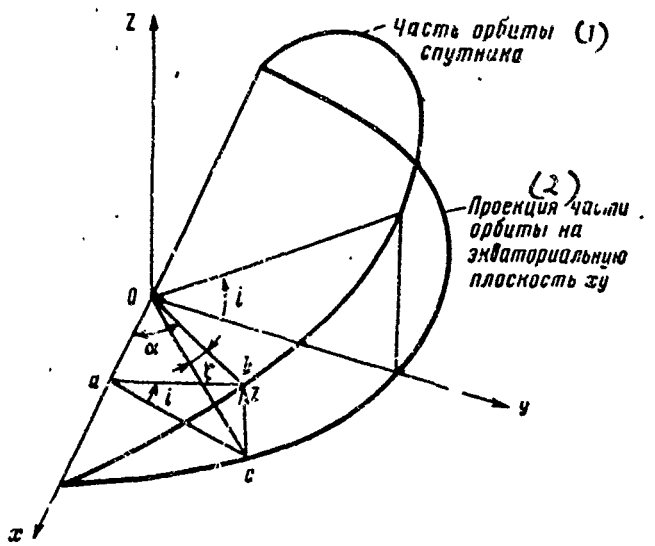
or, again taking into account (П.2.e), we will obtain

$$d\alpha = \frac{\cos\zeta\, d\zeta}{\sin i\sqrt{1-\left(\frac{\sin\zeta}{\sin i}\right)^2}}. \qquad (\Pi.2.6)$$

Consequently,

$$da = \frac{\cos\zeta\, d\zeta}{\sqrt{\sin^2 i - \sin^2\zeta}}. \qquad (\Pi.2.7)$$

Substituting this expression into formula (2.3.9), we obtain expression (2.3.10).



П.2.1. Explanation to the derivation of formula (2.3.10).

Key: (1). Part of the orbit of satellite. (2). Projection of part of orbit to equatorial plane XY.

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In Section 2.3 it was noted that with the values $\zeta_1=-i$ or $\zeta_2=i$ expression (2.3.10) leads to the uncertainty/indeterminacy. Let us consider the determination of integral for the case $\zeta_2=i$; a similar examination is applicable with $\zeta_1=-i$. Let us rewrite expression (2.3.10) in the form

$$p_k(1)=\frac{1}{2\pi^2}\int_{-\zeta_2}^{i-\varepsilon}\frac{\Delta\varphi(\zeta)\cos\zeta\,d\zeta}{\sqrt{\sin^2 i-\sin^2\zeta}}+\frac{1}{2\pi^2}\int_{i-\varepsilon}^{i}\frac{\Delta\varphi(\zeta)\cos\zeta\,d\zeta}{\sqrt{\sin^2 i-\sin^2\zeta}}, \quad (\text{П.2.8})$$

moreover we will consider ϵ as the very small angle. The first integral can be calculated graphically. For calculating the second let us assume that when change ζ near value of i the function

$$\Delta\varphi(\zeta)=a+b\sin(i-\zeta), \quad (\text{П.2.9})$$

where a and b - constants.

Designating second integral in expression (П.2.8) through $\Delta\rho_к$ (1) and taking into account (П.2.9), we will obtain

$$\Delta\rho_к(1)=\frac{1}{2\pi^2}\int_{i-\varepsilon}^{i}\frac{(a-b\cos i\sin\zeta+b\sin i\cos\zeta)\cos\zeta\,d\zeta}{\sqrt{\sin^2 i-\sin^2\zeta}}.$$

Substituting into first two addends $\sin\zeta=x\sin i$, and in last addend in accordance with (П.2.3) $\sin\zeta=\sin\alpha\sin i$, we will obtain

$$\Delta\rho_к(1)=\frac{1}{2\pi^2}\int_{x_\varepsilon}^{1}\frac{a-bx\sin i\cos i\,dx}{\sqrt{1-x^2}}+\frac{1}{2\pi^2}\int_{\arcsin x_\varepsilon}^{\frac{\pi}{2}}b\sin i\sqrt{1-\sin^2 i\sin^2\alpha}\,d\alpha,$$

here $$x_\varepsilon=\frac{\sin(i-\varepsilon)}{\sin i}.$$

As a result of integration let us find

$$\Delta\rho_к(1)=\frac{1}{2\pi^2}\left[a\arccos x_\varepsilon-\frac{b}{2}\sin 2i\sqrt{1-x_\varepsilon^2}+\right.$$
$$\left.+b\sin i\left\{E\left(\frac{\pi}{2};\ \sin i\right)-E(\arcsin x_\varepsilon;\ \sin i)\right\}\right]. \qquad (П.2.10)$$

Here $E\left(\frac{\pi}{2};\sin i\right)$ - second-order complete elliptic integral;

$E(\arcsin x_\varepsilon\cdot\sin i)$ - incomplete elliptical second-order integral.

If values $\Delta\varphi(\zeta)$ with $\zeta=i$ and $\zeta=i-\varepsilon$ are known and equal to respectively $\Delta\varphi(i)$ and $\Delta\varphi(i-\varepsilon)$, then coefficients, entering in (П.2.9), will be

$$a=\Delta\varphi(i),$$
$$b=\frac{\Delta\varphi(i-\varepsilon)-\Delta\varphi(i)}{\sin\varepsilon}.$$

In this case instead of (П.2.10) we will obtain

$$\Delta p_{\kappa}(i) = A\,\Delta\varphi(i) - B\,\Delta\varphi(i-\epsilon). \qquad (\text{П}.2.11)$$

Coefficients A and B, entering this expression, are functions i and ϵ. For value of ϵ=2.5° (0.0436 rad) coefficients A and B are represented by graphs in Fig. 2.3.2.

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Thus, knowing values of $\Delta\varphi(i)$ and $\Delta\varphi(i-2.5°)$ in accordance with carried out considerations instead of (П.2.8) we will obtain

$$p_{\kappa}(i) = \frac{1}{2\pi^2}\int_{-i}^{i-2,5°} \frac{\Delta\varphi(\zeta)\cos\zeta\,d\zeta}{\sqrt{\sin^2 i - \sin^2\zeta}} + A\,\Delta\varphi(i) - B\,\Delta\varphi(i-2,5°), \qquad (\text{П}.2.12)$$

where coefficients of A and B they are found from graphs Fig. 2.3.2 through known value of dip angle of orbit i.

Appendix 3.

Frequency bands, isolated to the communication systems through ISZ.

During determination of frequency band, necessary for communication systems through ISZ, were considered both future requirements for quantity of channels of communication, and possibilities of isolation/liberation of frequency sections in different ranges without disruption of work of existing radio aids. Taking into account this, in 1963 in Geneva was created the extreme administrative conference of radio communication (ChAKR), at which was

examined a question of the appropriation of the frequency bands for the space radiolink systems. Arrangement/position was based on the recommendations to reports of the X plenary assembly of MKKR. The fundamental technical principle, assumed as the basis of the appropriation of frequencies to space radiolink systems, is the possibility of the joint use of one frequency band several different radio services. The examination of a similar coincidence of different radio services in the general/common frequency band showed that the communication systems through ISZ it is technically expedient to combine with the radio relay lines of sight, which are placed in the category of the fixed/recorded services.

For decreasing interferences with each of radio services, worker in combined frequency band, are introduced specific limitations of radiated power, given in Chapter 9, and international coordination of appropriation of frequencies also is provided for.

Taking into account recommendations and reports of MKKR, for communication systems, which use ISZ, for different areas of world were isolated frequency bands, enumerated in Table П.3.1. Let us note that to area I pertains the USSR, the European countries and the country Africa, area II includes North and South America and Greenland; the remaining countries enter into area III.

Table П.3.1 shows services, with which communication systems through ISZ must together use different frequency bands on equal

rights/laws (on primary base). One should note that in some countries, except the services enumerated in the table, are some others. In the table italics printed the services, which use the appropriate frequency bands on the secondary base.

Stations, which use frequency bands on secondary base, must not cause harmful interferences with stations, which use frequency bands on primary base, they cannot require protection from harmful interferences from side of station, which use frequency bands on primary base, but they can require protection against harmful interferences from side of other stations of the same or other service, which use frequency bands on secondary base.

Let us give definitions of some services, enumerated in Table П.3.1. in accordance with regulations of radio communication, accepted in 1959 in Geneva.

Fixed/recorded service - radio communication service between specific fixed/recorded points/items.

Mobile service - radio communication service between mobile and ground stations or between mobile stations.

Service of radar - service of radio-determination, which realizes application of radar.

Amateur service - service of self-improvement, coupling and technical research, realized by lovers, i.e., by those, who have for this official permission/resolution and who are interested in radio engineering exclusively for personal purposes and without any material interest.

One should note that solutions, accepted by ChAKR, on the basis of which is comprised Table П.3.1. must be realized from 1 January, 1965.

Table п.з.1

(1) Район 1	(1) Район 2	(1) Район 3
3400÷3600 *Мгц* (2) (3) Фиксированная (5) Подвижная (6) Спутник связи (спутник—Земля) (8) *Радиолокация*	3400÷3500 *Мгц* (2) Радиолокация (4) (6) Спутник связи (спутник—Земля) (7) *Любители*	
3600÷4200 *Мгц* (2) (3) Фиксированная (4) Спутник связи (спутник—Земля) (5) Подвижная	3500÷3700 *Мгц* (2) (3) Фиксированная (5) Подвижная (8) Радиолокация (6) Спутник связи (спутник—Земля)	3500÷3700 *Мгц* (2) (8) Радиолокация (6) Спутник связи (спутник—Земля) (3) *Фиксированная* (5) *Подвижная*
	3700—4200 *Мгц* (2) (3) Фиксированная (5) Подвижная (6) Спутник связи (спутник—Земля)	
4400÷4700 *Мгц* (2) (3) Фиксированная (5) Подвижная (9) Спутник связи (Земля—спутник)		
5725÷5850 *Мгц* (2) (8) Радиолокация (10) Спутники связи (Земля—спутники) (7) *Любители*		
5850÷5925 *Мгц* (2) (3) Фиксированная (5) Подвижная (9) Спутники связи (Земля—спутник)		5850÷5925 *Мгц* (2) (3) Фиксированная (5) Подвижная (9) Спутники связи (Земля—спутник) (8) *Радиолокация*
5925÷6425 *Мгц* (2) (3) Фиксированная (5) Подвижная (9) Спутники связи (Земля—спутник)		
7250÷7300 *Мгц* (2) (6) Спутники связи (спутник—Земля)		
7300÷7750 *Мгц* (2) (3) Фиксированная (5) Подвижная (6) Спутники связи (спутник—Земля)		
7900÷7975 *Мгц* (2) (3) Фиксированная (5) Подвижная (9) Спутники связи (Земля—спутник)		
7975÷8025 *Мгц* (2) (9) Спутники связи (Земля—спутник)		
8025÷8400 *Мгц* (2) (3) Фиксированная (5) Подвижная (9) Спутники связи (Земля— спутник)		

Key: (1). Region. (2). MHz. (3). Fixed/recorded. (4). Radar. (5). Mobile. (6). Communication satellite (satellite-Earth). (7). Lovers. (8). Radar. (9). Communication satellite (Earth-satellite). (10). Communication satellites (Earth-satellites).

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Appendix 4.

Transmission levels.

In communication equipment it is accepted to express power levels, voltage and current not in absolute units (W, V, A), but in relative units (Np or dB), which characterize, on how much neper or decibel the examined power level, voltage or current (P; U; I) differs from appropriate initial values (P_0; U_0; I_0). Thus,

$$p = \frac{1}{2}\ln\frac{P}{P_0},\ \text{нп},\ \text{(1) или}\ p = 10\lg\frac{P}{P_0},\ \text{дБ}, \qquad (\text{П.4.1})$$

$$u = \ln\frac{U}{U_0},\ \text{нп},\ \text{(2) или}\ u = 20\lg\frac{U}{U_0},\ \text{дБ}, \qquad (\text{П.4.2})$$

$$i = \ln\frac{I}{I_0},\ \text{нп},\ \text{(3) или}\ i = 20\lg\frac{I}{I_0},\ \text{дБ}. \qquad (\text{П.4.3})$$

Key: (1). or.

Values in nepers and decibels are connected with relationships/ratios:

$$\left.\begin{aligned} 1\ \text{нп} &= 8{,}686\ \text{дБ} \\ 1\ \text{дБ} &= 0{,}115\ \text{нп}\ \text{(1)} \end{aligned}\right\}. \qquad (\text{П.4.4})$$

Key: (1). Np.

It is accepted to distinguish three forms of levels: absolute, relative and measuring. Let us consider the determination of these levels.

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If as initial level in expression (П.4.1) value P_0=1 mW is considered, this level is called absolute level and is expressed in dBm (decibels, calculated relative to 1 mW) or Np. Thus, absolute levels can be computed according to the following expressions:

$$\left.\begin{aligned} p &= \frac{1}{2}\ln\frac{P_{[мвт]}}{1_{[мвт]}} = \frac{1}{2}\ln P_{[мвт]}, \text{ нп } (1) \\ p &= 10\lg\frac{P_{[мвт]}}{1_{[мвт]}} = 10\lg P_{[мвт]}, \text{ дбм } (2) \end{aligned}\right\}. \qquad (П.4.5)$$

Key: (1). Np. (2). dBm.

For calculating absolute level of voltage U(v) or current I(ma) as original values in expressions (П.4.2) and (П.4.3) they take value of actual voltage U_0=0.775 V on resistor/resistance of 600 ohms or effective current I_0=1.29 mA, which takes place through resistor/resistance of 600 ohms. The selection of values U_0 and I_0 is determined by the fact that these the voltage and current on the resistor/resistance of 600 ohms develop the power of 1 mW. The absolute levels of voltage and current on any resistor/resistance of Z can be determined from the formulas:

$$u = \ln\frac{U_{[в]}}{0{,}775} - \frac{1}{2}\ln\frac{Z_{[ом]}}{600}, \text{ нп; } u = 20\lg\frac{U_{[в]}}{0{,}775} - 10\lg\frac{Z_{[ом]}}{600}, \text{ дб.} \qquad (П.4.6)$$

$$i = \ln\frac{I_{[ма]}}{1.29} + \frac{1}{2}\ln\frac{Z_{[ом]}}{600}, \text{ нп; } i = 20\lg\frac{I_{[ма]}}{1{,}29} + 10\lg\frac{Z_{[ом]}}{600}, \text{ дб.} \qquad (П.4.7)$$

If during calculation of transmission level for initial value is accepted power coefficient, voltages or current at point, which corresponds is accepted power coefficient, voltage or current at point, which corresponds to origin of circuit or system, then this level is called relative. Thus, relative level shows a difference in the levels (expressed in the nepers or the decibels) at the point in question and at the initial point of circuit.

Consequently, if at input of circuit absolute level of power is equal to $p_{вх}$ and at certain point of system absolute level is equal to p_1, then relative level will be determined by expression $p_1 - p_{вх}$.

The measuring level is called absolute level at point of circuit or system in question when in the beginning of it zero absolute level is established/installed, i.e., into point with zero absolute level signal is given by power of 1 mW.

One should note that during calculations sometimes it is necessary to determine amount of power in terms of known value of absolute transmission level p. In accordance with expressions (П.4.5) this can be carried out according to the formulas:

$P = e^{2p}$. mW (if p is determined in Np) (П.4.8);

$P = 10^{0.1p}$ mW (if p is determined in dBm) (П.4.9).

Appendix 5.

Characteristics of stationary random process with the zero

average/mean value and the normal distribution law.

According to [6.6] for stationary random process of u(t) with normal distribution law and with zero average/mean value

$$a = \overline{u(t)} = 0. \qquad (\text{П.5.1})$$

are satisfied following conditions.

1. Stability condition: function of distribution W of any order n remains constant/invariable with shift/shear of zero time reference to any interval τ, i.e.

$$W_n(u_1; u_2; \ldots; u_n; t_1; t_2; \ldots; t_n) = W_n(u_1; u_2; \ldots; u_n; t_1+\tau; \ldots; t_n+\tau). \qquad (\text{П.5.2})$$

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Here $u_1; u_2, \ldots, u_n$ - levels, which are not exceeded for any realization of random process at the moments of time $t_1; t_2, \ldots, t_n$.

2. One-dimensional function of distribution (probability density) does not depend on time:

$$W_1(u; t) = W_1(u; t+\tau) = W_1(u). \qquad (\text{П.5.3})$$

Taking into account that multichannel signal has normal distribution law and satisfies condition (П.5.1), we obtain, that one-dimensional function of distribution (one-dimensional probability density)

$$W_1(u) = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-\frac{u^2}{2\sigma^2}}. \qquad (\text{П.5.4})$$

This function has maximum, equal to $\frac{1}{\sqrt{2\pi\sigma^2}}$, with value of u=0; plotted functions for different σ are represented in Fig. П.5.1. From these graphs it follows that the less the parameter σ, the greater the maximum probability density and the more compact the weight values of function $W_1(u)$. Thus, with the decrease σ increases the probability of the determination of the process in the region of the assigned width in question, which includes point u=0.

Let us note that parameter σ - rms or standard deviation of random process.

3. Dispersion - is constant and does not depend on time. The value of the dispersion of process with the normal law distribution is equal to σ^2.

4. Two-dimensional distribution function can depend only on difference $\tau=t_2-t_1$:

$$W_2(u_1; u_2; t_1; t_2) = W_2(u_1; u_2; \tau). \qquad (П.5.5)$$

For normal distribution law with $\overline{u(t)}=0$

$$W_2(u_1; u_2; \tau) = \frac{1}{2\pi\sigma^2\sqrt{1-R^2(\tau)}} \exp \frac{-[u_1^2 + u_2^2 - 2R(\tau)u_1u_2]}{2[1-R^2(\tau)]\sigma^2}. \qquad (П.5.6)$$

Here $u_1=u(t_1)$, $u_2=u(t_2)=u(t_1+\tau)$, and value $R(\tau)$ - coefficient of correlation of stationary process:

$$R(\tau) = \frac{B(\tau)}{B(0)}, \qquad (П.5.7)$$

where $B(\tau)$ - correlation function of random process, found for value of $\tau=t_2-t_1$, and $B(0)$ - for value $\tau=0$.

5. Correlation function $B(t_1; t_2)$ - even, that depends only on difference in countdowns $t_2-t_1=\tau$:

$$B(t_1; t_2)=B(\tau)=\overline{u_1u_2}=\int_{-\infty}^{\infty}\int_{-\infty}^{\infty}u_1u_2W_2(u_1; u_2; \tau)\,du_1du_2. \qquad (\text{П.5.8})$$

Here $u_1=u(t_1)$; $u_2=u(t_1)=u(t_1+\tau)$.

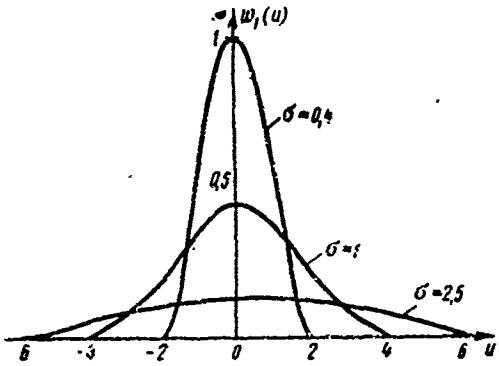

Fig. П.5.1. Dependence of $W_1(u)$ on σ.

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If stationary random process with zero average/mean value is determined by some regular (not random) function f(u) from value u, which is by chance in time, then correlation function of this process

$$B(\tau) = \overline{f(u_1) f(u_2)} = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} f(u_1) f(u_2) W_2(u_1; u_2; \tau)\, du_1 du_2. \quad (\text{П}.5.9)$$

Stationary processes, whose average/mean value and dispersion do not depend on time, but correlation function depends only on time difference τ, they are called stationary in the broad sense (stationary according to A. Ya. Khinchin). Let us note that the random processes, stationary in accordance with (П.5.2), will be stationary in the broad sense (but not vice versa).

If random process possesses property of ergodicity, correlation function can be calculated according to only realization, for example realization k of random process:

$$B(\tau) = \overline{u(t)\, u(t+\tau)} = \lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} u_k(t)\, u_k(t+\tau)\, dt. \quad (\text{П}.5.10)$$

Let us note that according to (П.5.8) correlation function characterizes degree of statistical connection/communication of adjacent elements of random process, distant to period τ.

For stationary processes, whose average/mean value is equal to zero, are observed conditions:

$$\lim_{\tau\to\infty} B(\tau) = B(\infty) = 0, \qquad (П.5.11)$$

$$\lim_{\tau\to 0} B(\tau) = B(0) = \sigma^2. \qquad (П.5.12)$$

On the other hand, assuming/setting in expression (П.5.10) $\tau\to 0$, let us find

$$\lim_{\tau\to 0} B(\tau) = B(0) = \overline{u^2(t)} = \lim_{T\to\infty} \frac{1}{T}\int_0^T u_к^2(t)\,dt = u_{к\,эфф}^2. \qquad (П.5.13)$$

Equalizing (П.5.12) and (П.5.13), we find

$$B(0) = \sigma^2 = \overline{u^2(t)} = u_{к\,эфф}^2. \qquad (П.5.14)$$

Consequently, correlation function with $\tau=0$ is equal to dispersion and is average/mean power of random process in question on load resistance/resistor, equal to 1 ohms.

6. Energy spectrum. As showed N. Wiener and A. Ya. Khinchin, the correlation function of the stationary random process $B(\tau)$ and spectral density (energy spectrum) $G(\Omega)$ were connected with Fourier transforms:

$$B(\tau) = \frac{1}{4\pi}\int_{-\infty}^{\infty} G(\Omega)\,e^{i\Omega\tau}\,d\Omega = \frac{1}{2\pi}\int_0^{\infty} G(\Omega)\cos\Omega\tau\,d\Omega, \qquad (П.5.15)$$

$$G(\Omega)=2\int_{-\infty}^{\infty}B(\tau)e^{-i\Omega\tau}d\tau=4\int_{0}^{\infty}B(\tau)\cos\Omega\tau\,d\tau. \qquad (\text{П}.5.16)$$

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Normal stationary processes can differ from each other in terms of form of correlation function or energy spectrum, given in each specific case by further conditions. The known amount of power ΔP, which falls to the band of frequencies $\Delta F=F_2-F_1$, where F_2 and F_1 - values of frequencies on the edges of band, can be one of such conditions.

In this case spectral density

$$G(\Omega)=\lim_{\Delta\Omega\to 0}2\pi\frac{\Delta P}{\Delta\Omega}=G_0. \qquad (\text{П}.5.17)$$

Correlation function $B(\tau)$ and spectral density $G(\Omega)$ of stationary random process possess all properties, inherent in pair of Fourier transforms. For example, during the expansion of energy spectrum $G(\Omega)$ correlation function is made by narrower more compact. By other words, for the family of the energy spectra of the assigned form the product of the time of correlation τ_0 to width of band ΔF of energy spectrum is constant value.

Let us note one additional property of functions $B(\tau)$ and $G(\Omega)$: correlation functions of two stationary random processes $\eta_1(t)$ and $\eta_2(t)=\frac{d\eta_1}{dt}=\eta_1'(t)$ are connected with dependence

$$B_2(\tau)=-B_1''(\tau), \qquad (\text{П}.5.18)$$

i.e. correlation function of process $\eta_2(t)$ is equal to second derivative of correlation function of process $\eta_1(t)$, undertaken with opposite sign. In this case must be satisfied condition $B''(0) \neq \infty$. Energy spectrum $G_2(\Omega)$ of random process $\eta_2(t)$ can be determined through energy spectrum $G_1(\Omega)$ of process $et_1(t)$ as follows:

$$G_2(\Omega) = \Omega^2 G_1(\Omega). \qquad (П.5.19)$$

Appendix 6.

Determination of the percentage distortions of the amplitude characteristic of amplifier.

Amplitude characteristic can be represented as $u_{вых} = f(u_{вх})$, where nonlinear dependence $f(u_{вх})$ is approximated by polynomial of form

$$f(u_{вх}) = a_1 u_{вх} + a_2 u_{вх}^2 + \ldots + a_m u_{вх}^m. \qquad (П.6.1)$$

Here $u_{вх}$ - voltage on the input of amplifier,

$u_{вых}$ - output potential of amplifier,

a_n - the coefficients of polynomial, which correspond to degree of n.

$$a_1 u_{вх} = u, \qquad (П.6.2)$$

we will obtain

$$u_{вых} = u + a_2 u^2 + a_3 u^3 + \ldots + a_m u^m. \qquad (П.6.3)$$

Here

$$a_2 = \frac{a_2}{a_1^2}; \; a_3 = \frac{a_3}{a_1^3}; \; \ldots; \; a_m = \frac{a_m}{a_1^m}. \qquad (П.6.4)$$

If we to the input of amplifier conduct the harmonic oscillation

$$u_{вх} = U_{вх} \cos \Omega t, \qquad (П.6.5)$$

at the output will have

$$u_{max} = U\cos\Omega t + \sum_{n=2}^{m} a_n U^n \cos^n \Omega t. \qquad (П.6.6)$$

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Here

$$U = a_1 U_{вх}. \qquad (П.6.7)$$

Being limited m=5 and taking into account that

$$\left.\begin{aligned} \cos^2 x &= \frac{1}{2}(\cos 2x + 1) \\ \cos^3 x &= \frac{1}{4}(\cos 3x + 3\cos x) \\ \cos^4 x &= \frac{1}{8}(\cos 4x + 4\cos 2x + 3) \\ \cos^5 x &= \frac{1}{16}(\cos 5x + 5\cos 3x + 10\cos x) \end{aligned}\right\}. \qquad (П.6.8)$$

After conversion and without terms with constant component instead of (П.6.6) we will obtain

$$u_{max} = \sum_{k=1}^{5} U_k \cos k\Omega t, \qquad (П.6.9)$$

moreover in accordance with (П.6.6) and (П.6.8)

$$\left.\begin{aligned} U_1 &= U + \frac{3}{4}a_3 U^3 + \frac{5}{8}a_5 U^5 \\ U_2 &= \frac{1}{2}a_2 U^2 + \frac{1}{2}a_4 U^4 \\ U_3 &= \frac{1}{4}a_3 U^3 + \frac{5}{16}a_5 U^5 \\ U_4 &= \frac{1}{8}a_4 U^4 \\ U_5 &= \frac{1}{16}a_5 U^5 \end{aligned}\right\}. \qquad (П.6.10)$$

Hence follows:

$$\left.\begin{aligned} a_5 &= 16\frac{U_5}{U^5} \\ a_4 &= 8\frac{U_4}{U^4} \\ a_3 &= 4\left(\frac{U_3}{U^3}-\frac{5}{16}\frac{a_5U^5}{U^3}\right)=\frac{4}{U^3}(U_3-5U_5) \\ a_2 &= 2\left(\frac{U_2}{U^2}-\frac{1}{2}\frac{a_4U^4}{U_2}\right)=\frac{2}{U^2}(U_2-4U_4) \end{aligned}\right\}. \qquad (П.6.11)$$

the value U, entering these formulas, can be determined from the expression for U_1:

$$U=U_1-\frac{3}{4}a_3U^3-\frac{5}{8}a_5U^5=U_1-3(U_3-5U_5)-10U_5=U_1-3U_3-5U_5. \qquad (П.6.12)$$

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Consequently, if the amplitudes of oscillations U_1-U_5 will be measured, then on (П.6.11), (П.6.12) can be obtained the values of coefficients of a_2-a_5.

Furthermore, if is known value of input voltage $U_{вх}$, then from (П.6.12) and (П.6.7) can be obtained value of coefficient $a_1=\frac{U}{U_{вх}}$, and from (П.6.4) and (П.6.11) - value of any coefficient $a_m=a_m a_1^m$, entering initial expression (П.6.1).

When characteristic of nonlinearity of amplifier can be approximated by polynomial of smaller degree, for example cubic, all coefficients can be calculated according to given formulas, if we in them assume that values, which have indices higher than third, are equal to zero. The obtained relationships/ratios make it possible to determine coefficients a or a on the known voltages U_1 and U_5 or to solve inverse problem. However, in practice it is frequently necessary to determine on coefficients of a or α the values of the

coefficients of nonlinear distortions or the value of the index of nonlinearity.

By coefficient of nonlinear distortions on n harmonic howls ratio of amplitude of this harmonic to amplitude of fundamental harmonic, measured at one and the same point of circuit during the supplying of specific level p_1 into starting point of circuit:

$$K_n(p_1)=\frac{U_n}{U_1}. \qquad (П.6.13)$$

Index of nonlinearity of order n at point, where absolute level of fundamental harmonic is equal to p_1, which corresponds to voltage U_1, it is determined by relationship/ratio

$$b_n(p_1)=\ln\frac{U_1}{U_n}. \qquad (П.6.14)$$

If we in (П.6.13) or (.P.6.14) substitute values of voltages from (П.6.10), then it is easy to see that coefficient of nonlinear distortions and attenuation index depend on stress level of fundamental harmonic, on which they are determined. Thus, for example, considering for simplification of calculations that $a_4=a_5=0$, and assuming/setting sufficient to low value of a_3, as a result of which $U\gg\frac{3}{8}a_3U^3$, from (П.6.10) let us find:

$$U_1=U;\ U_2=\frac{1}{2}a_2U_1^2;\ U_3=\frac{1}{4}a_3U_1^3.$$

Therefore in this case

$$K_2=\frac{1}{2}a_2U_1;\ b_2=\ln\cdot\frac{2}{a_2U_1},$$
$$K_3=\frac{1}{4}a_3U_1^2;\ b_3=\ln\frac{4}{a_3U_1^2}.$$

Let us note that from given relationships/ratios it follows that with change in input voltage U_1 Z times (or power Z^2 times) value of percentage distortions K_n will be changed Z^{n-1} times, and attenuation of nonlinearity b_n - to value, equal to $\ln \frac{1}{Z^{n-1}}$. This is correct under the condition of the low nonlinearity of the characteristic ($a_n \ll 1$ or $a_n \ll a_1$, where n=2, 3, 4.). Thus, depending on value U_1 will be obtained different values K_n and b_n In order to introduce certainty during the determination of values K_n and b_n, it is necessary to indicate the value of voltage level (or power level), with which were determined or must be determined these values. Specifically, this is stressed by the fact that in (П.6.13) and (П.6.14), and also in the subsequent expressions for the coefficients of nonlinear distortions and indices of nonlinear distortions in the brackets is indicated the power level of the fundamental harmonic, on which are determined K_n and b_n.

From (П.6.13) and (П.6.14) it follows that between K_n and b_n there are relationships/ratios:

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$$K_n(p_1) = e^{-b_n(p_1)}, \tag{П.6.15}$$

$$b_n(p_1) = \ln \frac{1}{K_n(p_1)}. \tag{П.6.16}$$

Taking into account (П.6.14), instead of (П.6.12) and (П.6.11), we will obtain:

$$U = U_1\left(1 - 3\frac{U_2}{U_1} + 5\frac{U_4}{U_1}\right) = U_1(1 - 3e^{-b_2(p_1)} + 5e^{-b_4(p_1)}) = U_1 A_1, \tag{П.6.17}$$

$$A_1 = 1 - 3e^{-b_2(p_1)} + 5e^{-b_4(p_1)}, \tag{П.6.18}$$

$$\left.\begin{aligned} a_2 &= \frac{2}{U_1 A_1^2}\left(\frac{U_2}{U_1} - 4\frac{U_4}{U_1}\right) = \frac{2}{U_1 A_1^2}\left(e^{-b_2(p_1)} - 4e^{-b_4(p_1)}\right) \\ a_3 &= \frac{4}{U_1^2 A_1^3}\left(\frac{U_2}{U_1} - 5\frac{U_3}{U_1}\right) = \frac{4}{U_1^2 A_1^3}\left(e^{-b_2(p_1)} - 5e^{-b_3(p_1)}\right) \\ a_4 &= 8\frac{1}{U_1^3 A_1^4}e^{-b_4(p_1)}; \quad a_5 = 16\frac{1}{U_1^4 A_1^5}e^{-b_5(p_1)} \end{aligned}\right\}. \tag{П.6.19}$$

If amplifier introduces low nonlinear distortions and, consequently, attenuations of nonlinearity of different orders large, it is possible to count

$$e^{-b_3(\rho_1)} \ll 1;\ e^{-b_5(\rho_1)} \ll 1;\ e^{-b_4(\rho_1)} \ll 1;\ e^{-b_6(\rho_1)} \ll 1. \qquad (\text{П}.6.20)$$

Inequalities (П.6.20) make it possible to considerably simplify obtained expressions, since in this case value $A_1=1$, and therefore instead of (П.6.17) let us find $U=U_1$. Considering that with voltage U_1 on resistor/resistance of R power P_1 is developed, on the basis of Appendix 4 we have

$$U = U_1 = \sqrt{2RP_1} = \sqrt{2R\,e^{2\rho_1}} = \sqrt{2R}\,e^{\rho_1} \qquad (\text{П}.6.21)$$

and respectively we will obtain

$$\left.\begin{aligned} a_3 &= 2\left\{e^{-[b_3(\rho_1)+\rho_1]} - 4e^{-[b_4(\rho_1)+\rho_1]}\right\}\frac{1}{\sqrt{2R}} \\ a_5 &= 4\left\{e^{-[b_5(\rho_1)+2\rho_1]} - 5e^{-[b_6(\rho_1)+2\rho_1]}\right\}\frac{1}{2R} \\ a_6 &= 8e^{-[b_4(\rho_1)+3\rho_1]}\frac{1}{2R\sqrt{2R}} \\ a_7 &= 16e^{-[b_6(\rho_1)+4\rho_1]}\frac{1}{4R^2} \end{aligned}\right\} \qquad (\text{П}.6.22)$$

With $\rho_1=0$, i.e., during the determinations at point with the zero absolute level, these expressions substantially are simplified.

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Expression (П.6.22) establishes connection/communication between indices of nonlinearity of order n and coefficients a_n. Keeping in mind (П.6.16) and (П.6.21), on the basis (П.6.22) it is easy to switch over to the coefficients of the nonlinear distortions:

$$\left.\begin{aligned} a_3 &= \frac{2}{U_1}[K_3(p_1) - 4K_4(p_1)] \\ a_3 &= \left(\frac{2}{U_1}\right)^3 [K_3(p_1) - 5K_5(p_1)] \\ a_4 &= \left(\frac{2}{U_1}\right)^3 K_4(p_1) \\ a_5 &= \left(\frac{2}{U_1}\right)^4 K_5(p_1) \end{aligned}\right\}. \qquad (П.6.23)$$

It is obvious that for determining coefficients a_n from expressions (П.6.22) or (П.6.23) it is necessary to fulfill measurements on all harmonics, including the fifth. This is realized only in such cases, when the cut-off frequencies of the group spectrum F_1-F_2 are connected with relationship/ratio $5F_1 \leq F_2$, which corresponds to the relative width of the group spectrum

$$\frac{F_2 - F_1}{0{,}5(F_2 + F_1)} > \frac{4F_1}{0{,}5 \cdot 6F_1} = 1{,}34.$$

When relative bandwidth is lower than this value, determination of coefficients a_n from given formulas cannot be fulfilled, since it will prove to be impossible to measure attenuation of nonlinearity or coefficients of nonlinear distortions of higher orders (for example, the fourth and the fifth). For determining the coefficients a_n in such systems it is necessary to resort to other methods of measurements, for example to the method of two tones. In this case to

the input of amplifier are supplied two oscillations/vibrations with the identical amplitudes and such frequencies of $F_a=\frac{\Omega_a}{2\pi}$ and $F_c=\frac{\Omega_c}{2\pi}$, which lie/rest at the band of the efficiently reinforced frequencies. Then

$$u_{ax}=U_{ax}[\cos\Omega_a t+\cos\Omega_c t]. \qquad (\Pi.6.24)$$

Analogously (П.6.6) at output of amplifier we will obtain

$$u_{вых}=U(\cos\Omega_a t+\cos\Omega_c t)+\sum_{n=2}^{m}a_n U^n(\cos\Omega_a t+\cos\Omega_c t)^n. \qquad (\Pi.6.25)$$

Here U is determined by expression (П.6.7).

From expression (П.6.25) it follows that together with the components of the frequencies of form $m\Omega_a$ and $m\Omega_c$ during the supplying to the input of the nonlinear amplifier of two oscillations/vibrations at the output will be present the components

$$\cos(p\Omega_a t)\cos(q\Omega_c t)=\frac{1}{2}\{\cos[(p\Omega_a+q\Omega_c)t]+\cos[(p\Omega_a-q\Omega_c)t]\}. \qquad (\Pi.6.26)$$

Here p and q - series/rows of independent natural numbers, moreover p+q=n, where n=2, 3, 4, 5.

Taking into account that relative width of band of amplifier is usually small, we will be bounded to examination of components with frequencies $\Omega_a-\Omega_c$; $2\Omega_a-\Omega_c$; $2\Omega_a-2\Omega_c$; $3\Omega_a-2\Omega_c$, formed with values p and q, respectively equal to p=1 and q=1; p=2 and q=1; p=2 and q=2; p=3 and q=2, and also components with frequencies of Ω_a and Ω_c, formed with values of p=1; q=0 and p=0; q=1.

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For these components from (П.6.25) we will obtain:

$$\left.\begin{aligned} U_{\Omega_a} &= U_{\Omega_c} = U + \frac{9}{4}a_3U^3 + \frac{25}{4}a_5U^5 \\ U_{\Omega_a-\Omega_c} &= a_2U^2 + 3a_4U^4 \\ U_{2\Omega_a-\Omega_c} &= \frac{3}{4}a_3U^3 + \frac{25}{8}a_5U^5 \\ U_{2\Omega_a-2\Omega_c} &= \frac{3}{4}a_4U^4 \\ U_{3\Omega_a-2\Omega_c} &= \frac{5}{8}a_5U^5 \end{aligned}\right\}. \qquad (П.6.27)$$

From these expressions let us find:

$$\left.\begin{aligned} a_5 &= \frac{8}{5}\frac{U_{(3\Omega_a-2\Omega_c)}}{U^5} \\ a_4 &= \frac{4}{3}\frac{U_{(2\Omega_a-2\Omega_c)}}{U^4} \\ a_3 &= \frac{4}{3}\left[\frac{U_{(2\Omega_a-\Omega_c)}}{U^3} - \frac{25}{8}\frac{a_5U^5}{U^3}\right] = \frac{4}{3U^3}\left[U_{(2\Omega_a-\Omega_c)} - 5U_{(3\Omega_a-2\Omega_c)}\right] \\ a_2 &= \frac{U_{(\Omega_a-\Omega_c)}}{U^2} - \frac{3a_4U^4}{U^2} = \frac{1}{U^2}\left[U_{(\Omega_a-\Omega_c)} - 4U_{(2\Omega_a-2\Omega_c)}\right] \end{aligned}\right\}. \qquad (П.6.28)$$

Value U can be determined through value U_{Ω_a} or U_{Ω_c} (from П.6.27):

$$U = U_{\Omega_a} - \frac{9}{4}U^3 - \frac{25}{4}a_5U^5 = U_{\Omega_a} - 3U_{(2\Omega_a-\Omega_c)} - 5U_{(3\Omega_a-2\Omega_c)}. \qquad (П.6.29)$$

Subsequently for brevity of recording we will lower in indices of designation Ω_a and Ω_c, keeping in mind, that in first place in index is written/recorded reference number of harmonic of frequency Ω_a, and in second place of index is written/recorded reference number of harmonic of frequency Ω_c. Thus, component $U_{(3\Omega_a-2\Omega_c)}$ will be registered as U_{3-2}, and designation $U_{\cdot}$ will determine component U_{Ω_a}.

Taking into account aforesaid and keeping in mind, which

$$U_1 = \sqrt{2RP_1} = \sqrt{2R\,e^{2p_1}} = \sqrt{2R}\,e^{p_1},$$
$$\frac{U_{p-q}}{U_1} = e^{-b_{p-q}(p_1)},$$

moreover $b_{p-q}(p_1)$ determines attenuation of nonlinearity of order n=p+q (in nepers) at point, where level of fundamental tone is equal to p_1, expressions (П.6.29) and (П.3.16) can be registered in the form:

$$U = U_1\left(1 - 3\frac{U_{2-1}}{U_1} - 5\frac{U_{3-2}}{U_1}\right) = U_1\left(1 - 3e^{-b_{2-1}} - 5e^{-b_{3-2}}\right) = U_1B_1, \quad (П.6.30)$$

$$B_1 = 1 - 3e^{-b_{2-1}} - 5e^{-b_{3-2}} \quad (П.6.31)$$

$$\left.\begin{aligned}
a_2 &= \frac{1}{U_1B_1^2}\left(\frac{U_{1-1}}{U_1} - 4\frac{U_{2-2}}{U_1}\right) = \frac{1}{U_1B_1^2}\left[e^{-b_{1-1}(p_1)} - 4e^{-b_{2-2}(p_1)}\right]\\
a_3 &= \frac{4}{3U_1^2B_1^3}\left(\frac{U_{2-1}}{U_1} - 5\frac{U_{3-2}}{U_1}\right) = \frac{4}{3U_1^2B_1^3}\left[e^{-b_{2-1}(p_1)} - 5e^{-b_{3-2}(p_1)}\right]\\
a_4 &= \frac{4}{3U_1^3B_1^4}\frac{U_{2-2}}{U_1} = \frac{4}{3U_1^3B_1^4}e^{-b_{2-2}(p_1)}\\
a_5 &= \frac{8}{5U_1^4B_1^5}\frac{U_{3-2}}{U_1} = \frac{8}{5U_1^4B_1^5}e^{-b_{3-2}(p_1)}
\end{aligned}\right\}. \quad (П.6.32)$$

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In the case of amplifiers with a small nonlinearity of value

$$e^{-b_{1-1}(p_1)} \ll 1;\ e^{-b_{2-1}(p_1)} \ll 1;\ e^{-b_{2-2}(p_1)} \ll 1;\ e^{-b_{3-2}(p_1)} \ll 1. \quad (П.6.33)$$

In this case of $B_1=1$ expressions are simplified:

$$U = U_1 = \sqrt{2R}\,e^{p_1} \quad (П.6.34)$$

$$\left.\begin{aligned}
a_2 &= \frac{1}{\sqrt{2R}}\left[e^{-(b_{1-1}+p_1)} - 4e^{-(b_{2-2}+p_1)}\right]\\
a_3 &= \frac{2}{3R}\left[e^{-(b_{2-1}+p_1)} - 5e^{-(b_{3-2}+p_1)}\right]\\
a_4 &= \frac{4}{3(2R)^{\frac{3}{2}}}e^{-(b_{2-2}+p_1)}\\
a_5 &= \frac{2}{5R^2}e^{-(b_{3-2}+p_1)}
\end{aligned}\right\}. \quad (П.6.35)$$

In these formulas for brevity of recording index p_1 in value, which determines attenuation of nonlinearity, is omitted, since absolute level of fundamental tone p_1 is written/recorded in the same expression.

Appendix 7.

Derivation of formulas, which determine coordinates of points of interacting antennas.

Let us at first consider case, when angle of elevation, at which occurs reception or emission of oscillation/vibration of terrestrial antenna to side of satellite, is equal to zero. Fig. П.7.1 shows the surface of the Earth radius R=OK=OM=OC with center O. Let us assume that at point K the antenna of terrestrial station is established/installed, and along the circular equatorial orbit ЕАД with a height/altitude of H=MA satellite moves. If we to point K lead tangential plane, then the orbit of satellite will cross plane P at two points - in A and D. With plane P the axis of the major lobe/lug of the antenna radiation pattern coincides. We will consider that the satellite is located at point A, and let us establish the geometric relationships/ratios, which determine the coordinates of points A and K depending on the orbit altitude H=MA and latitude of point K (which is determined by angle ζ=KOC). Let us consider that the zero meridian passes through point O'; therefore the longitude/length of point K is determined by angle $\measuredangle O'OC$, and the longitude/length of satellite - by angle $\measuredangle O'OA$. Let us designate difference in these longitudes/lengths through angle $\varphi = \measuredangle AOC$.

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After dropping perpendicular from point K to equatorial plane, we will obtain point N, moreover $KN \perp OC$. From Δ OKN follows

$$KN = R\sin\zeta, \qquad (П.7.1)$$

$$ON = R\cos\zeta. \qquad (П.7.2)$$

Let us connect points N and A and let us consider ΔONA, in which $\angle AON=\varphi$, OA=R+H, ON=R cos ζ; therefore according to law of cosines $(AN)^2=(AO)^2+(ON)^2-2(AO)(ON)\cos\varphi$, after substitution we obtain

$$(AN)^2=(R+H)^2+R^2\cos^2\zeta-2R(R+H)\cos\zeta\cos\varphi. \quad (\Pi.7.3)$$

In ΔAKN straight line AN, which lies at equatorial plane (orbit EAD it is equatorial), to perpendicular line KN. Therefore

$$(AK)^2=(AN)^2+(KN)^2.$$

Taking into account (П .7.1) and (П .7.3), after conversions we find

$$(AK)^2=(R+H)^2+R^2-2R(R+H)\cos\zeta\cos\varphi. \quad (\Pi.7.4)$$

Let us consider now ΔAKO. Since plane P at point K is tangential to the surface of the Earth, $\sphericalangle AKO=\frac{\pi}{2}$.

Therefore

$$(AK)^2=(AO)^2-(OK^2)=(R+H)^2-R^2. \quad (\Pi.7.5)$$

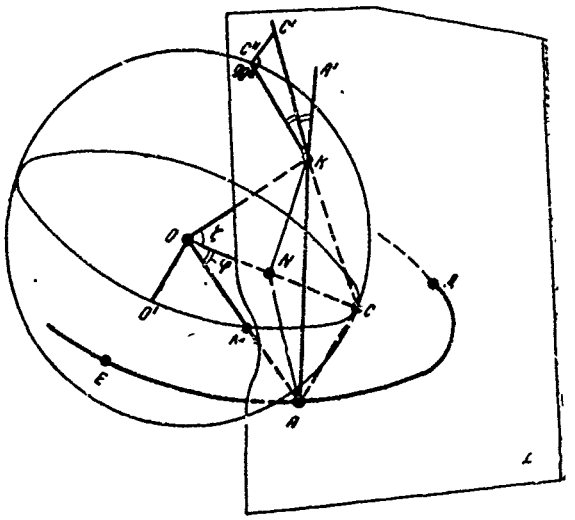

Fig. П.7.1. To the derivation of the formulas, which determine the coordinates of the interacting stations.

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Equalizing (П.7.4) and (П.7.5), we will obtain

$$\varphi = \arccos \frac{R}{(R+H)\cos\zeta}. \qquad (П.7.6)$$

Let us now move on to more general case, when angle of elevation of ground antenna to side of satellite is not equal to zero, i.e.,

$$\sphericalangle AKO = \frac{\pi}{2} + \beta = \gamma. \qquad (П.7.7)$$

here β - angle of elevation.

From Δ AKO according to the theorem of the sines

$$\frac{AO}{\sin(\sphericalangle AKO)} = \frac{OK}{\sin(\sphericalangle KAO)}.$$

Therefore

$$\frac{R+H}{\sin\gamma} = \frac{R}{\sin(\sphericalangle KAO)}.$$

Consequently, upon consideration of (П.7.7)

$$\measuredangle KAO = \arcsin\left(\frac{R}{R+H}\right)\cos\beta.$$

Since sum of angles of triangle is equal to π, and angle $\measuredangle AKO=\gamma$,

$$\measuredangle KOA = \pi - \left[\gamma + \arcsin\left(\frac{R}{R+H}\right)\cos\beta\right]. \qquad (\text{П.7.8})$$

Using for Δ AKO law of cosines, we will obtain

$$(AK)^2 = (AO)^2 + (OK)^2 - 2(AO)(OK)\cos(\measuredangle KOA)$$

with substitution (П.7.8) and the introduced earlier designations. after conversions let us find

$$(AK)^2 = (R+H)^2 + R^2 + 2(R+H)R\cos\left[\gamma + \arcsin\frac{R}{R+H}\cos\beta\right]. \qquad (\text{П.7.9})$$

Equalizing (П.7.9) and (П.7.4), we will obtain

$$\cos\varphi = \frac{-\cos\left[\gamma + \arcsin\left(\frac{R}{R+H}\right)\cos\beta\right]}{\cos\zeta}. \qquad (\text{П.7.10})$$

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Taking into account (П.7.7), and also known trigonometric relationships/ratios

$$\arcsin\alpha = -\arccos\alpha + \frac{\pi}{2},$$

$$\cos(\pi+\alpha) = -\cos\alpha,$$

we will obtain

$$\cos\varphi = \frac{\cos\left[\arccos\left(\frac{R}{R+H}\right)\cos\beta - \beta\right]}{\cos\zeta}. \qquad (\text{П.7.11})$$

However, after leading similar linings/calculations, it is

possible to obtain following relationship/ratio, which determines azimuth angle of direction from ground station to ISZ [artificial earth satellite]:

$$\operatorname{tg} A \sin \zeta = \operatorname{tg} \varphi. \qquad (\Pi.7.12)$$

Distance between terrestrial station and ISZ, designated through L, is determined by straight line AK. In accordance with (П.7.9) and (П.7.11)

$$L^2 (AK)^2 = (R+H)^2 \left[1 + \frac{K^2}{(R+H)^2} - 2\frac{R}{R+H} \cos \varphi \cos \zeta\right]. \qquad (\Pi.7.13)$$

Let us note that for stationary satellite relation

$$\frac{R}{R+H} = 0{,}151. \qquad (\Pi.7.14)$$

Appendix 8.

VALUE OF COEFFICIENT OF WEAKENING OF INTERFERENCES FOR COMMUNICATION SYSTEMS THROUGH ISZ AND FOR RRL.

Table П.8.1 gives values of KOP for RRL, calculated in [9.12] according to expression (9.5.6) with different number of telephone channels in RRL and along communication system through ISZ and with different separation of carrier frequencies. During these calculations it was accepted that in those cases, when in the communication system through ISZ there is no charging (number of transmitted signals equal to zero), it is introduced the dispersion of the energy, which carries, upon which the peak value of the deviation of frequency it is 300 kHz (see columns 2 and 3 of table).

In column of 3 tables is given value of frequency band on hf/VCh for communication system through ISZ, found from expression

$$\Delta f_{\text{ИСЗ}} = 2(\Delta f_{\text{пик}} + F_2). \qquad (\text{П.8.1})$$

Here $\Delta f_{\text{пик}}$ - peak deviation of frequency, determined from expression (7.3.7);

F_2 - frequency of upper channel in the group spectrum of the communication system through ISZ.

In columns 4 and 6 are given values of width of band of group spectrum of systems in question.

In columns 5 and 7 values of efficient deviation of frequency $\Delta f_{\text{эф.}}$ accepted for RRL and for communication system through ISZ in

accordance with graphs Fig. 7.3.6 and formulas (7.3.7) are given.

In column 8 are indicated values of separation between carrier frequencies, at which were performed calculations.

In columns 9-15 are given values $\varkappa_0$ (in dB), found for indicated value of frequency $F_{\text{к}}$, being located in group spectrum, moreover values $\varkappa_0 > 300$ dB are conditionally designated through "-0".

Table П.8.2 gives values of KOP for communication systems through ISZ, designed according to expression (9.5.6). In this case the designations in the columns are accepted the same as in Table П.8.1.

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Table П.8.1. Value κ_0 for RRL (mixing signal from the system with IZS).

(1) Число телефонных каналов		Δf ИСЗ Мгц (2)	ΔF РРЛ Мгц (2)	$\Delta f_{ср}$ РРЛ Мгц (2)	ΔF ИСЗ Мгц (2)	$\Delta f_{ср}$ ИСЗ Мгц (2)	$\Delta f_р$ Мгц (2)	$\varkappa_0$, дб, при следующих величинах $F_к$ РРЛ, Мгц (3)						
РРЛ	ИСЗ							0	0,5	1,0	2,0	4,0	5,0	8,0
1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
120	0	0,600	0,552	0,195	0,001	0,300	0	12,8	17,7					
	20	3,060			0,084	0,178	0	10,8	20,9					
	40	5,070			0,168	0,357	0	13,4	17,1					
	80	8,300			0,336	0,710	0	16,2	17,2					
							4	80,6	68,6					
	150	12,900			0,630	1,335	0	18,8	19,1					
							4	38,0	36,1					
							8	95,6	89,2					
	300	21,800			1,260	2,420	0	21,4	21,5					
							4	27,3	27,1					
							8	45,0	44,2					
							12	74,5	72,7					
	600	37,500			2,520	4,080	0	23,7	23,7					
							4	25,7	25,7					
							8	32,0	31,9					
							12	42,4	42,2					
							16	57,0	56,5					
							24	98,6	97,9					
	1200	66,000			5,040	7,030	0	26,0	26,0					
							4	26,7	26,7					
							8	28,8	28,8					
							12	32,3	32,3					
							16	37,3	37,2					
							24	51,3	51,2					
							32	71,0	70,8					
600	0	0,600	2,660	0,872	0,001	0,300	0	5,8	11,3	19,1	19,5			
							4	73,4	48,3	31,3	22,5			
	20	3,060			0,084	0,178	0	3,6	17,1	19,4	19,4			
							4	103,7	81,7	37,9	22,4			
	40	5,070			0,168	0,357	0	6,5	10,4	18,1	19,6			
							4	60,0	42,7	30,1	22,6			
	80	8,300			0,336	0,710	0	9,3	10,3	12,9	19,4			
							4	34,7	31,5	27,4	22,4			
							8	103,7	103,7	106,7	81,4			
	150	12,900			0,630	1,335	0	11,7	12,0	12,8	15,8			
							4	26,1	25,2	23,2	18,8			
							8	64,4	60,6	54,8	44,4			
							12	103,7	103,7	103,7	98,9			
	300	21,800			1,260	2,420	0	14,0	14,1	14,4	15,4			
							4	19,5	19,4	19,0	18,1			
							8	35,4	34,7	33,1	29,2			
							12	61,0	59,6	56,6	50,0			
							16	97,0	94,4	89,9	80,5			
	600	37,500			2,520	4,080	0	16,2	16,2	16,3	16,7			
							4	18,2	18,2	18,2	18,3			
							8	24,3	24,2	23,9	23,1			
							12	34,3	34,1	33,5	31,5			
							16	48,4	47,9	46,9	43,8			
							24	88,2	97,3	85,1	79,7			

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	1200	66,000			5,040	7,030	0	18,5	18,5	18,6	18,7		
							4	19,2	19,2	19,2	19,3		
							8	21,3	21,3	21,3	21,2		
							12	24,8	24,7	24,7	24,5		
							16	29,6	29,6	29,5	29,0		
							24	43,5	43,4	43,1	41,9		
							32	63,0	62,7	62,2	60,3		
960	0	0,600	4,188	1,102	0,001	0,300	0	1,5	7,3	18,1	18,9	20,3	
							4	20,3	21,4	21,9	21,9	4,5	
							8	—0	—0	—0	106,2	23,3	
	20	3,060			0,084	0,178	0	—0,7	14,5	18,9	18,9	19,6	
							4	19,6	21,8	21,9	21,9	2,3	
							8	—0	—0	—0	—0	22,€	
	40	5,070			0,168	0,357	0	2,3	6,4	16,1	18,9	20,5	
							4	20,5	21,3	21,9	21,9	5,3	
							8	—0	—0	106,2	89,1	23,5	
	80	8,300			0,336	0,710	0	5,2	6,2	9,2	17,4	21,1	
							4	21,1	21,3	21,6	20,4	8,2	
							8	92,8	80,0	66,2	44,6	24,1	
							12	—0	—0	—0	106,2	95,8	
	150	12,900			0,630	1,335	0	7,8	8,0	8,9	12,0	20,5	
							4	20,5	20,1	18,8	15,0	10,8	
							8	45,6	43,4	39,4	32,5	23,5	
							12	103,2	97,8	89,6	73,7	48,6	
	300	21,800			1,260	2,420	0	10,1	10,2	10,5	11,5	15,4	
							4	15,4	15,3	15,0	14,1	13,1	
							8	29,6	29,1	27,9	24,6	18,4	
							12	50,8	49,8	47,5	42,2	32,6	
							16	81,7	79,7	75,9	67,9	53,8	
	600	37,500			2,520	4,080	0	12,2	12,3	12,4	12,7	14,2	
							4	14,2	14,2	14,2	14,3	14,6	
							8	20,2	20,1	19,8	19,1	17,1	
							12	29,9	29,7	29,1	27,3	23,1	
							16	43,2	42,9	41,9	39,0	32,9	
							24	80,4	79,5	77,6	72,7	63,1	
	1200	66,000			5,040	7,030	0	14,5	14,6	14,6	14,7	15,2	
							4	15,2	15,2	15,3	15,3	15,7	
							8	17,3	17,3	17,3	17,3	17,2	
							12	20,7	20,7	20,7	20,4	19,7	
							16	25,6	25,5	25,4	24,9	23,4	
							24	39,3	39,2	38,9	37,8	34,7	
							32	58,5	58,3	57,7	55,9	51,2	
1200	0	0,600	5,684	1,233	0,001	0,300	0	—1,7	4,2	17,1	18,6	18,6	18,7
							4	18,6	18,6	18,6	21,0	1,3	20,1
							8	102,9	105,9	74,0	30,0	21,6	21,6
							12	—0	—0	—0	—0	105,9	74,0
	20	3,060			0,084	0,178	0	—4,0	12,1	18,6	18,6	18,6	18,6
							4	18,6	18,6	18,6	21,5	—1,0	21,6
							8	—0	—0	105,9	35,8	21,6	21,6
	40	5,070			0,168	0,357	0	—1,0	3,2	14,2	16,6	18,6	18,7
							4	18,6	18,6	18,7	20,9	2,0	17,2
							8	102,9	88,9	61,1	28,9	21,6	21,6
							12	—0	—0	—0	—0	105,9	61,1
	80	8,300			0,336	0,710	0	2,0	3,0	6,1	15,9	18,7	19,4
							4	18,7	18,7	19,0	18,2	5,0	9,1
							8	51,2	44,4	36,6	26,5	21,7	21,6
							12	—0	105,9	105,9	105,9	54,2	36,6
	150	12,900			0,630	1,335	0	4,6	4,9	5,8	9,2	18,0	20,1
							4	18,0	17,5	16,0	12,0	7,6	8,8
							8	32,5	30,4	29,4	25,5	21,0	17,0
							12	78,1	73,3	66,3	53,8	35,5	29,5
							16	102,9	102,9	105,9	105,9	81,2	55,3
	300	21,800			1,260	2,420	0	7,1	7,2	7,4	8,5	12,4	15,2
							4	12,4	12,3	12,0	11,1	10,0	10,4
							8	25,1	24,8	23,9	21,2	15,5	13,2
							12	42,0	41,3	39,7	35,6	28,1	24,7
							16	68,6	67,0	63,8	56,9	45,0	40,0
							24	102,9	102,9	102,9	105,9	105,9	98,1

	600	37,500			2,520	4,080	0	9,2	9,3	9,3	9,7	11,2	12,3	
							4	11,2	11,2	11,2	11,3	11,6	11,9	
							8	17,1	17,0	16,7	16,0	14,1	13,3	
							12	26,4	26,3	25,7	24,0	20,0	18,2	
							16	38,9	38,6	37,7	35,2	29,5	26,8	
							24	72,8	72,1	70,5	66,1	57,3	53,2	
							32	102,9	102,9	102,9	102,9	97,4	92,0	
	1200	66,000			5,040	7,030	0	11,5	11,5	11,5	11,7	12,2	12,6	
							4	12,2	12,2	12,2	12,3	12,7	12,9	
							8	14,3	14,2	14,2	14,2	14,1	14,1	
							12	17,7	17,7	17,6	17,4	16,7	16,2	
							16	22,5	22,4	22,3	21,8	20,4	19,5	
							24	36,1	36,0	35,6	34,6	31,5	29,9	
							32	54,9	54,7	54,2	52,4	47,8	45,5	
1800	0	0,600	8,204	1,059	0,001	0,300	0	—8,2	—2,2	13,9	18,2	18,2	18,2	19,5
							4	18,2	18,2	18,2	18,2	—5,2	16,9	21,2
							8	19,5	20,6	21,2	21,2	21,2	21,2	—5,2
							12	—0	—0	—0	105,5	22,5	21,2	21,2
							16	—0	—0	—0	—0	—0	—0	22,5
	20	3,060			0,084	0,178	0	—10,4	6,4	18,2	18,2	18,2	18,2	18,8
							4	18,2	18,2	18,2	18,2	—7,4	21,2	21,2
							8	18,8	21,0	21,2	21,2	21,2	21,2	—7,4
							12	—0	—0	—0	—0	21,8	21,2	21,2
							16	—0	—0	—0	—0	—0	—0	21,8
	40	5,070			0,168	0,357	0	—7,4	—3,2	9,1	18,2	18,2	18,2	19,7
							4	18,2	18,2	18,2	18,2	—4,4	12,1	21,2
							8	19,7	20,5	21,2	21,2	21,2	21,2	—4,4
							12	—0	—0	105,5	87,3	22,7	21,2	21,2
							16	—0	—0	—0	—0	—0	105,5	22,7
	80	8,300			0,336	0,710	0	—4,5	—3,4	—0,2	11,7	18,2	18,2	20,4
							4	18,2	18,2	18,2	13,8	—1,5	2,8	21,2
							8	20,4	20,5	20,9	21,2	21,2	21,1	—1,4
							12	91,4	78,9	65,1	43,7	23,4	21,4	21,2
							16	—0	—0	—0	105,5	94,4	65,1	23,4
	150	12,900			0,630	1,335	0	—1,7	—1,4	—0,5	3,1	15,0	17,9	20,7
							4	15,0	13,8	11,3	5,9	—1,3	2,5	18,0
							8	20,7	20,8	20,9	21,1	18,0	11,8	1,3
							12	44,7	42,5	38,6	31,7	23,8	22,1	18,0
							16	102,5	96,5	88,6	72,7	47,8	38,7	23,8
	300	21,800			1,260	2,420	0	0,8	0,9	1,2	2,3	6,5	9,6	19,4
							4	6,5	6,4	6,0	4,9	3,8	4,2	9,5
							8	19,4	19,1	18,4	15,7	9,5	7,1	3,8
							12	30,5	30,3	29,5	27,3	22,4	19,4	9,5
							16	50,2	49,2	47,0	42,0	33,6	30,2	22,4
							24	102,5	102,5	102,5	100,8	83,8	75,1	53,2
	600	37,500			2,520	4,080	0	3,0	3,1	3,2	3,5	5,1	6,2	11,1
							4	5,1	5,1	5,1	5,1	5,4	5,8	8,0
							8	11,1	11,0	10,7	9,9	8,0	7,1	6,0
							12	20,4	20,3	19,7	18,0	14,0	12,2	8,1
							16	31,9	31,7	31,0	28,8	23,4	20,6	14,1
							24	60,8	60,3	59,0	55,4	48,0	44,5	34,9
							32	102,5	100,8	98,5	94,1	83,3	78,1	63,8
	1200	66,000			5,040	7,030	0	5,4	5,4	5,4	5,5	6,0	6,4	8,1
							4	6,0	6,1	6,1	6,2	6,5	6,8	8,0
							8	8,1	8,1	8,1	8,1	8,0	7,9	8,0
							12	11,6	11,5	11,5	11,3	10,5	10,1	9,0
							16	16,4	16,3	16,2	15,7	14,2	13,4	11,1
							24	30,0	29,8	29,5	28,5	25,4	23,6	19,4
							32	48,5	48,3	47,8	46,0	41,6	39,3	33,0

Key: (1). Number of telephone channels. (2). MHz. (3). κ_0, dB, with following values F_K of RRL, MHz

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Table П.8.2. Value κ_0 for the systems with ISZ (interfering signal from RRL).

(1) Число телефонных каналов		Δf ИСЗ Мгц (2)	ΔF РРЛ Мгц (2)	$\Delta f_{ср}$ РРЛ Мгц (2)	ΔF ИСЗ Мгц (2)	$\Delta f_{ср}$ ИСЗ Мгц (2)	$\Delta f_р$ Мгц (2)	κ_0, дб, при следующих величинах $F_н$ ИСЗ, Мгц (3)						
РРЛ	ИСЗ							0	0,5	1,0	2,0	4,0	5,0	8 0
1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
120	20	3,060	0,552	0,195	0,084	0,178	0	34,6						
	40	5,070			0,168	0,357	0	34,2						
	80	8,300			0,336	0,710	0	33,9						
							4	98,3						
	150	12,900			0,630	1,335	0	33,8	34,1					
							4	53,0	51,1					
	300	21,800			1,260	2,420	0	32,5	32,6	32,9				
							4	38,4	38,3	37,9				
							8	56,1	55,3	53,3				
							12	85,7	83,8	80,1				
	600	37,500			2,520	4,080	0	30,3	30,3	30,4	30,8			
							4	32,4	32,4	32,4	32,4			
							8	38,6	38,5	38,3	37,4			
							12	49,0	48,8	48,1	46,1			
							16	63,6	63,2	62,0	58,7			
							24	105,3	104,5	102,0	96,3			
	1200	66,000			5,040	7,030	0	28,3	28,4	28,4	28,5	29,0	29,4	
							4	29,0	29,1	29,1	29,2	29,5	29,8	
							8	31,2	31,2	31,1	31,1	31,0	31,0	
							12	34,7	34,6	34,6	34,4	33,6	33,1	
							16	39,6	39,5	39,4	38,9	37,4	36,5	
							24	53,6	53,5	53,2	52,0	48,8	47,2	
							32	73,3	73,1	72,5	70,6	65,8	63,4	
600	20	3,060	2,660	0,872	0,084	0,178	0	34,6						
	40	5,070			0,168	0,357	0	34,8						
							4	88,2						
	80	8,300			0,336	0,710	0	34,5						
							4	59,9						
	150	12,900			0,630	1,335	0	34,2	34,4					
							4	48,6	47,7					
							8	86,9	83,1					
	300	21,800			1,260	2,420	0	32,6	32,7	33,0				
							4	38,1	38,0	37,6				
							8	54,0	53,3	51,7				
							12	79,6	78,2	75,2				

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	600	37,500			2,520	4,080	0	30,3	30,3	30,4	30,8					
							4	32,3	32,3	32,3	32,4					
							8	38,4	38,3	38,0	37,2					
							12	48,4	48,2	47,6	45,7					
							16	62,5	62,1	61,0	57,9					
							24	102,3	101,4	99,2	93,8					
	1200	66,000			5,040	7,030	0	28,3	28,3	28,4	28,5	29,0	29,4			
							4	29,0	29,0	29,0	29,1	29,5	29,8			
							8	31,1	31,1	31,1	31,0	30,9	30,9			
							12	34,6	34,5	34,5	34,3	33,5	33,1			
							16	39,4	39,4	39,3	38,8	37,3	36,4			
							24	53,3	53,2	52,9	51,7	48,6	47,0			
							32	72,8	72,5	72,0	70,1	65,3	63,0			
960	20	3,060	4,168	1,102	0,084	0,178	0	34,4								
							4	54,7								
	40	5,070			0,168	0,357	0	34,4								
							4	52,6								
	80	8,300			0,336	0,710	0	34,2								
							4	50,2								
	150	12,900			0,630	1,335	0	34,1	34,4							
							4	46,9	46,4							
							8	71,9	69,7							
	300	21,800			1,260	2,420	0	32,6	32,7	32,9						
							4	37,9	37,8	37,5						
							8	52,1	51,6	50,4						
							12	73,2	72,2	69,9						
							16	104,1	102,2	98,3						
	600	37,500			2,520	4,080	0	30,2	30,3	30,4	30,7					
							4	32,2	32,2	32,2	32,3					
							8	38,1	38,1	37,8	37,1					
							12	47,9	47,7	47,1	45,3					
							16	61,2	60,8	59,9	57,0					
							24	98,3	97,5	95,6	90,7					
	1200	66,000			5,040	7,030	0	28,2	28,2	28,3	28,4	28,9	29,3			
							4	28,9	28,9	28,9	29,0	29,4	29,7			
							8	31,0	31,0	31,0	30,9	30,8	30,8			
							12	34,4	34,4	34,3	34,1	33,4	33,0			
							16	39,2	39,2	39,1	38,6	37,1	36,3			
							24	53,0	52,9	52,6	51,4	48,4	46,7			
							32	72,2	72,0	71,4	69,5	64,9	62,6			
1200	20	3,060	5,684	1,233	0,084	0,178	0	34,1								
							4	56,7								
	40	5,070			0,168	0,357	0	34,1								
							4	53,7								
	80	8,300			0,336	0,710	0	34,0								
							4	50,7								
							8	83,2								
	150	12,900			0,630	1,335	0	34,0	34,3							
							4	47,3	46,8							
							8	61,8	60,8							

	300	21,800			1,260	2,420	0	32,6	32,7	32,9				
							4	37,9	37,8	37,5				
							8	50,6	50,3	49,4				
							12	67,5	66,8	65,2				
							16	94,1	92,5	89,3				
	600	37,500			2,520	4,080	0	30,2	30,2	30,3	30,7			
							4	32,2	32,2	32,2	32,2			
							8	38,0	38,0	37,7	37,0			
							12	47,4	47,3	46,7	45,0			
							16	59,9	59,6	56,7	56,2			
							24	93,8	93,1	91,4	87,1			
1200	1200	66,000	5,684	1,233	5,040	7,030	0	28,2	28,2	28,2	28,4	28,9	29,3	
							4	28,9	28,9	28,9	29,0	29,4	29,6	
							8	30,9	30,9	30,9	30,9	30,8	30,8	
							12	34,4	34,3	34,3	34,1	33,3	32,9	
							16	39,2	39,1	39,0	38,5	37,0	36,2	
							24	52,7	52,6	52,3	51,2	48,2	46,6	
							32	71,6	71,4	70,9	69,1	64,5	62,2	
1800	20	3,060	8,204	1,059	0,084	0,178	0	33,8						
							4	62,4						
							8	63,0						
	40	5,070			0,168	0,357	0	33,8						
							4	59,4						
							8	60,9						
	80	8,300			0,336	0,710	0	33,7						
							4	56,4						
							8	58,5						
	150	12,900			0,630	1,335	0	33,7	34,0					
							4	50,5	49,3					
							8	56,2	56,2					
							12	80,2	78,0					
	300	21.800			1,260	2,420	0	32,4	32,5	32,8				
							4	38,1	38,0	37,6				
							8	51,0	50,7	49,9				
							12	62,1	61,8	61,1				
							16	81,8	80,8	78,6				
	600	37,500			2,520	4,080	0	30,1	30,2	30,3	30,6			
							4	32,2	32,2	32,2	32,2			
							8	38,1	38,1	37,8	37,0			
							12	47,5	47,3	46,8	45,1			
							16	59,0	58,8	58,1	55,9			
							24	87,9	87,4	86,1	82,5			
	200	86,000			5,040	7,030	0	28,1	28,2	28,2	28,3	28,8	29,2	
							4	28,8	28,8	28,9	29,0	29,3	29,6	
							8	30,9	30,9	30,9	30,8	30,8	30,7	
							12	34,4	34,3	34,3	34,1	33,3	32,9	
							16	39,2	39,1	39,0	38,5	37,0	36,2	
							24	52,7	52,6	52,3	51,2	48,2	46,6	
							32	71,3	71,1	70,6	68,8	64,4	62,1	

Key: (1). Number of telephone channels. (2). MHz. (3). κ_0, dB, with following values of ISZ, MHz.

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No Typing.

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