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FOREIGN TECHNOLOGY DIVISION

BROADBAND RADIO COMMUNICATION

by

A. M. Semenov, A. A. Sikarev

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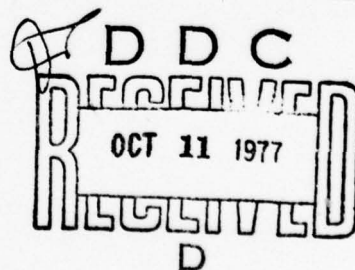
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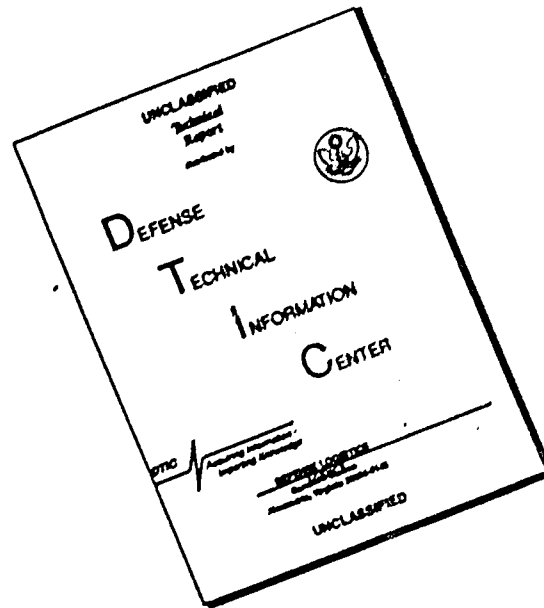
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BROADBAND RADIO COMMUNICATION

By: A. M. Semenov, A. A. Sikarev

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TABLE OF CONTENTS

U. S. BOARD ON GEOGRAPHIC NAMES TRANSLITERATION SYSTEM

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.
The use of diacritical marks is preferred, but such marks
may be omitted when expediency dictates.

GREEK ALPHABET

Alpha	A	α	**α**	Nu	N	ν		
Beta	B	β		Xi	Ξ	ξ		
Gamma	Γ	γ		Omicron	O	o		
Delta	Δ	δ		Pi	Π	π		
Epsilon	E	ε	**ϵ**	Rho	P	ρ	**ϱ**	
Zeta	Z	ζ		Sigma	Σ	σ	**ς**	
Eta	H	η		Tau	T	τ		
Theta	Θ	θ	**ϑ**	Upsilon	Τ	υ		
Iota	I	ι		Phi	Φ	**φ**	φ	
Kappa	K	**ϰ**	κ	**ϗ**	Chi	X	χ	
Lambda	Λ	λ		Psi	Ψ	ψ		
Mu	M	μ		Omega	Ω	ω		

RUSSIAN AND ENGLISH TRIGONOMETRIC FUNCTIONS

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

———

rot	curl
lg	log

GRAPHICS DISCLAIMER

All figures, graphics, tables, equations, etc.
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BROADBAND RADIO COMMUNICATION.

A. M. Semenov, A. A. Sikarev.

Pages 1-278.

BROADBAND RADIO COMMUNICATION.

A. M. Semenov, A. A. Sikarev.

Orders of the Red Banner of Labor the military publishing house of
the Ministry/Department of Defense of the USSR Moscow - 1970.

Page 2.

Broadband radio communication. Voenizdat, 1970. 280 s., 10.000
copy 63 kopeckss.

Broadband radio communication of the series valuable properties:
by large reticence and by interference shielding (in comparison with
usual radio communication), the possibility of the readout, etc.

In the book, written using the open Soviet and foreign
materials, are examined the fundamental principles of construction
and special feature/peculiarity of broadband communicating systems,
are given examples of the realization of such systems, are shown the
prospects for their further development and use in radio
communication.

The book is intended to the cadets of the military schools,
students of the military academies and officers of the troops of
communication/connection. It can be also used by the wide circle of
the readers, who are interested in the problems of the contemporary
technique of communication/connection.

Page 3.

Preface.

In passed decade noticeably increased the interest of the
specialists, working in the range of radio communication, in the
so-called broadband methods of the transmission of

report/communications on radio channels. Is explained this by the
fact that the broadband radio communication in comparison with usual
possesses a series of the valuable (especially for a military radio
communication) properties: the high correctness of the transmission
of information in channels with multiple-pronged propagation and in
the handled sections of range, by larger stability with respect to
spot jammings and somewhat by larger reticence.

In recent years in Soviet and foreign periodicals appeared the
large number of works, dedicated to the in-depth investigations of
separate aspects and to the description of the
concrete/specific/actual specimen/samples of the systems of broadband
communication/connection. However, the wide circles of the
specialists of radio communication are still little familiar with the
fundamental operating principles of such systems. This is explained
by the absence of those generalizing and at the same time the
available to wide circle readers of the works, in which the
principles of the construction of broadband systems would be examined
from unity of opinion with the detailed analysis of the physical
essence of phenomena, and also by the fact that for the construction
of broadband communicating systems are utilized some specific
principles, which are based on the last/latter achievements of
statistical radio engineering.

The proposed book must to a certain degree complete the indicated gap/spacing. In this case in it is enveloped only the basic group of the questions, on understanding of which depends the possibility of further deeper research on broadband communication/connection.

Page 4.

In presenting the material the authors approached possibly the simpler and more available interpretation of very complex processes and principles, lying at the base of the work of the broadband transmission systems of discrete information. Into the book are introduced some common questions, which relate to the statistical theory of communication/connections, without understanding the which the work of broadband systems cannot be correctly understood.

Introduction, chapter 1 and 5 are written by A. M. Semenov; *Chapter* gl. 2, 3, 4 - by A. A. Sikarev; chapter 6 and conclusion are developed by

the authors together.

The authors express their gratitude to all comrades, who gave a series of useful advice in the process of the work on the book.

All wishes and observations about the book one should guide to: Moscow, K-160. Voenizdat.

Page 5.

Introduction.

Radio communication has at present exclusively important value in all branches of the economic and cultural life of our country.

Stable control of troops under conditions of modern combat is also unthinkable without the wide use of a radio communication. Important place and value for a control as troops has short-wave

radio communication, and also the radio communication, which uses tropospheric or ionospheric scattering and reflection from meteor trails. By the most important and most valuable property of these forms of radio communication is the possibility of the establishment of the direct/straight communication/connection between the moved away from each other correspondents at the limited power of radio transmitter.

At the same time the continuous and whole being accelerated increase in the number of radio stations, which work in skip band, led to the fact that as a result of the large number of simultaneously working radio stations very strongly increased the level of interferences, and the problem of the distribution of spectrum and rigid regulation of its use it became difficult solved.

Under these conditions the solution of the problem of providing a stable radio communication on skip band became necessary to search for not only by the path of the search of measures for a decrease in the interstation interferences by the methods of the ordering of the organization of radio communication service, but also by means of the search of such methods of radio communications, which would provide durable relationship under conditions of unavoidable powerful

interferences. In other words, besides the problem of fight with
interferences arose the problem of providing a radio communication
under conditions unavoidable and during that sufficient powerful
interferences.

Page 6.

To the essential factors, which complicates radio communication
and which lower its stability by skip band, and also when using
tropospheric (or ionospheric) scattering or reflection from meteor
trails, one should relate also the phenomenon of multiple-pronged
radiowave propagation, when emitted by transmitter signal comes into
the point of reception/procedure on the different paths, which are
characterized each by their phase and amplitude characteristics. This
leads to the fact that the signal at the point of reception/procedure
is the vector sum of the signals, the amplitude and phase
relationship/ratios between which they depend on their individual
characteristics of propagation. Since these characteristics
continuously change according to random law, also the power of signal
at the input of receiver continuously changes; these changes are
different at the different frequencies, even closely distant each
other, which leads not only to the fluctuations of level, but also to

the disturbance/breakdown of the relationship/ratios of the frequency
components in the composition of signal, i.e., to its distortion.

The tendency possible more effective to overcome the noted above
difficulties of obtaining stable radio communication led to the
development of the fundamentally new methods of using a frequency
spectrum - the methods of the transmission of information with the
aid of broadband signals. Information theories completely justifies
the advisability of applying such signals under conditions of
powerful interferences, frequency "hunger" and of multiple-pronged
propagation. More that, the broadband systems turn out to be
completely justified also when it is necessary to provide
communication/connection under conditions of the effect of man-made
interferences from enemy (damping) which, undoubtedly is for a
military radio communication completely necessary.

The important special feature/peculiarity of the systems of
broadband radio communication is also the fact that during their
practical use turns out to be possible under specific conditions of
news correspondent's stable reception/procedure even when with the
reception of such signals to usual narrow-band receiver their level
below average interference level, i.e., when signals are found "under

noise".

Page 7.

use in the broadband radiolink systems of the signals of complex form
(noise-like) impedes also the extraction of information from signal,
if are unknown the data on its structure. This also is quite
significant for a military radio communication.

Broadband radio communication can ensure the high authenticity
of reception with the very small fluctuations of the probability of
errors, while in narrow-band systems sometimes (in the absence of
interference on the given frequency) the authenticity of
communication/connection will be very high, and sometimes very low.
This is caused by the fact that in the band of frequencies of the
narrow-band signal the spectral density of interferences fluctuates
strongly, and in the band of frequencies of the broadband signal it
is little affected. With the expansion of band occurs the averaging
of the operating in this band interferences.

All advantages of broadband systems are caused by the fact that in them widely are utilized the principles of the statistical theory of communication/connection, which make it possible most complete to realize the optimum conditions of the reception of signals.

The methods of broadband radio communication, as a rule, are designed for the transmission of discrete report/communications. However, it is necessary to keep in mind that the technician of the transmission of discrete report/communications have already have long it exceeded the limits of the transmission only of text (telegraphy) and it composes at present one of the most important component/links of the data-transmission systems for a remote control and in other ranges, but the most promising transmission systems of continuous report/communications (for example, telephone signals) they are based also on their transformation into discrete by the so-called quantization. Thus, broadband communication/connection can be applied for many forms of the radio communication: telegraphy, data transmissions, telephony and other methods of the transmission of information.

The first works, dedicated to broadband systems, were works of Feno (1952) and A. A. Charkevich (1957). Later both in foreign and in Soviet literature appeared the very large number of works, in detail examining the different sides of the problem of broadband communication/connection.

Broadband radio communication in its properties and the methods of technical realization differs significantly from the usual traditional methods of radio communication.

Page 8.

Most important its differences are use of signals with the frequency band, considerably wider, than the band of the transmitted report/communication, and the methods of selection, based on the application/use of signals of various forms on that which transmit and matched with the waveform of filters on receiving ends.

Broadband radiolink systems completely consistent in principle with narrow-band, i.e., on just one section of range can

simultaneously work those and others, without exerting serious
interferences with each other. At the same time it is necessary to
keep in mind that near the transmitter, which radiates broadband
signals, the reception of the moved away correspondents, who work by
narrow band, is substantially hinder/hampered in an entire emission
band of transmitter.

The fundamental principles of this form of communication and
path of their possible realization are examined in L. M. Fink's
fundamental work "theory of the transmission of discrete
report/communications" [36], N. L. Tyuermal's works [33], A. G. Zyuko
[17] and a series of others. In journal "foreign electronics", No 9
for 1965 was published survey/coverage of the most important foreign
works in this range [27]; to this survey/coverage was applied vast
bibliography.

This book has as a goal to acquaint the wide circle of the
readers with the most important properties, special
feature/peculiarities and the advantages of broadband
communication/connection, and also to give the representation of some
concrete/specific/actual systems of this communication/connection and
the paths of their realization. For larger convenience in the use of

the material of the book possibly by the wider circle of the readers
the authors considered advisable to construct the book as follows.

In the first chapter are set forth the fundamental principles of
broadband communication/connection, are examined its fundamental
special feature/peculiarities and advantages, are given
survey/coverage and classification of the mostd widely use systems of
broadband communication/connection. Research on this chapter does not
require any special knowledge from the field of mathematics and
statistical radio engineering and at the same time can be sufficient
in order to clarify the fundamental principles and the ideas,
embedded into such systems.

Page 9.

In the second chapter are examined some most important questions
of the statistical theory of radio communication, substantially
necessary for more a fundamental understanding of the fundamental
principles of the construction of broadband radiolink systems.
Specifically, is given short noise characteristic with radio
reception, is given the concept of the methods of the quantitative

evaluation of the properties of these interferences as random
processes, are examined the most important positions of the theory of
the potential interference rejection, which lay as the basis of the
construction of broadband systems, are described the operating
principles of fundamental and at the same time the specific
cell/elements of equipment for these systems.

The third chapter is dedicated to the more detailed examination
of the broad class of the so-called mutually correlated broadband
systems, which make it possible to the best degree to realize all
advantages of the broadband methods of communication/connection, but
which possess during their practical realization a comparative
complexity. In this chapter are examined the principles of the
realization of such systems, their freedom from interference, the
block diagrams and the work of the most important elements. A typical
example of the mutually correlated system of broadband
communication/connection is the system "Rake" whose description
appeared into 1958 [44].

In the fourth chapter is given the information about the
autocorrelation systems, which are simplest, but not making it
possible to obtain such high results from the freedom from

interference, which can be reached in mutually correlated systems.

The fifth chapter is dedicated to the systems, which most
frequently are called discrete-address. Such systems receive at
present especially wide acceptance in the range of VHF for a
radiotelephone circuit, since they allow in certain frequency band to
simultaneously work to the large number of transmitters. In this case
to any subscriber is represented the possibility of call and work
with any other subscriber of grid/network so simply, as this is made
with the aid of the automatic telephone station of wire
communication. An example of the discrete-address system, intended
specially for an army radio communication, is the system whose
description appeared in the literature in 1961 [46]. In this system
can be realized the communication/connection between the large number
of subscribers, utilizing a common/general/total for all
communication/connections frequency band. Each subscriber can
immediately cause other.

Page 10.

The span of range here is substantially more than for one narrow-band

channel, but is considerably less than the necessary frequency band for the realization of the system with frequency division multiplex or any to another, in which to each subscriber it would be abstract/removed its frequency band.

In the last/latter chapter of the book are placed the materials, which make it possible to conduct a comparative evaluation of different broadband systems according to their interference shielding and their comparison with narrow-band systems.

Entire material in the book is presented in connection with the use of broadband signals in radiolink systems with the maximum isolation/liberation of the physical essence of phenomena and the interpretation of the most important results, obtained recently.

MT/ST-77-0122

Page 11.

Chapter 1.

GENERAL INFORMATION ABOUT BROADBAND RADIO Communication SYSTEMS.

§1.1. Basic concepts. Base of signal.

Until recently in the development of radio communication ruled
the tendency of an all possible reduction in the spectrum, occupied
in ether/ester by transmitter during the transmission of
report/communication, to maximum possibilities. This was explained
first of all by the fact that than already the spectrum of signal,
the more number of radio stations it can be placed in the assigned
section of frequency band, providing the absence of interferences.
The tendency to throttle/taper the frequency band contributed to the
achievement of the large successes in the range of frequency fixing,
to the creation of highly efficient selective systems and to the wide
introduction of the most effective in relation band of frequencies of
single-band modulation.

The high degrees of frequency fixing of radio stations and the
narrow frequency bands, occupied by signals, made it possible at the
same time clear to regulate the use of a frequency range.

In recent years the majority of the ranges, utilized for a radio
communication, and first of all skip band, turned out to be so
handled that the clear regulation of frequencies turned out to be

completely impossible and with interferences it became necessary to consider as the unavoidable phenomenon.

The tendency to ensure durable relationship under conditions of interferences and to eliminate the effect on the stability of the communication/connection of the phenomenon of multiple-pronged propagation they led to onset and development of opposite tendency - toward the use of complex broadband signals and toward failure within certain limits from the traditional method of the selection of radio station in frequency.

Page 12.

In broadband discrete communicating systems unlike the communicating systems, which use the simple signals, when each realization of signal (for example, release and pressure in binary systems) is the cut of harmonic oscillation, which differs in terms of amplitude, frequency, by the initial phase or several of these parameters, the cell/element of signal is not the cut of harmonic oscillation, and it has more complex form.

To utilize for the first time complex noise signals for the
transmission of information proposed A. A. Charkevich [37]. He
indicated the possibility of using a noise as carrying
oscillation/vibration (carrier). By a change in the intensity of
noise it is possible to carry out modulation, similar by usual
amplitude, while by a change in the cut-off frequencies of the noise
spectrum - frequency-noise modulation.

Figure 1.1.1 gives the block diagram of the formation of
noise-like signal, proposed to F. Lange [21]. One realization of
signal here is formed by the sum of the initial noise z (t) and the
same noise, shifted for a period τ_1 with the aid of delay line, but
another realization - by the sum of the initial noise and the same
noise, delayed for a period τ_2.

Another example of the formation of noise-like signal is given
in the diagram of Fig. 1.1.2. Here of the input of delay line, which
has the large number of removal/outlets, for the transmission of each
cell/element of signal is supplied narrow pulse.

Output broadband signal is obtained by means of the addition of
the group of the short delayed pulses, removed in the removal/outlets
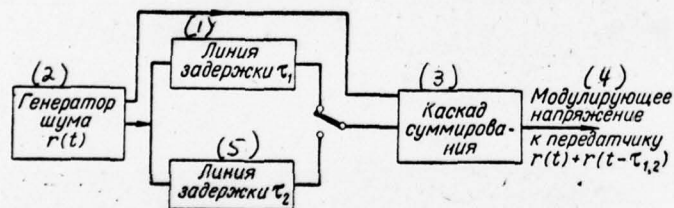of delay line.

Fig. 1.1.1.

Key: (1). Delay line τ_1. (2). Noise generator r (t). (3).

Cascade/stage of addition. (4). Modulating stress to transmitter.

(5) Delay line T_2.

To each realization (for example, to pressure and release)
corresponds its group of removal/outlets.

The more skeletal diagram of the formation of serrated signal is
given in Fig. 1.1.3. Here to the input of the group of narrow-band
filters is supplied momentum/impulse/pulse sufficient short duration
(δ- momentum/impulse/pulse). After each filter is included
amplifier-attenuator, which determines the transmission factor of
each component (β_k),   and delay line, also with the individually
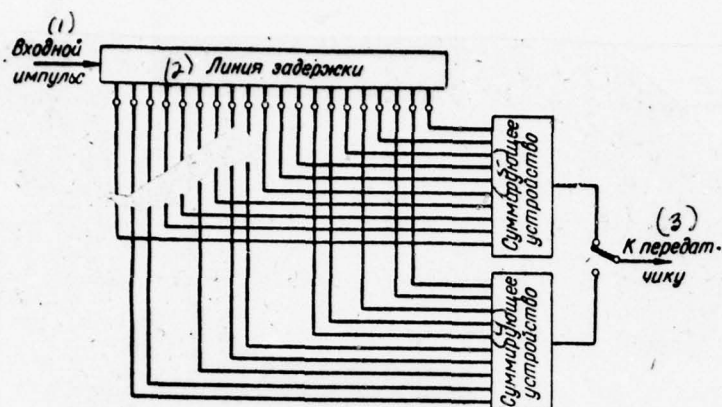selected for each component delay time (τ_k).

Fig. 1.1.2.

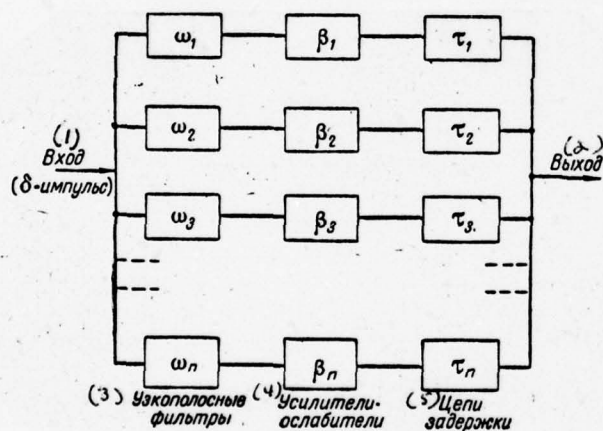Key: (1). Input pulse. (2). Delay line. (3). To transmitter. (4).
Adder. (5). Adder.

Fig. 1.1.3.

Key: (1). Input (δ-momentum/impulse/pulse). (2). Output/yield. (3).
Narrow-band filters. (4). Amplifier-attenuator. (5). Delay circuits.

Page 14.

By a change in the number of filters (n), of frequencies of these filters (ω_n), values β_k and τ_k it is possible over wide limits to vary the form of the form/shaped serrated signal. It is not difficult to see that the earlier examined diagram (Fig. 1.1.2) is a special case of this more common diagram.

Let us note that by a change in values $\beta_k, \tau_k, \omega_n$ in the determined law can be reached the further "complication" of waveform. As one of special cases of broadband signal can serve, for example, the signal with the linearly changing frequency, which found use in radar. It is possible to give the very large number also of other examples of broadband signals.

At the present time, specifically, finds wide acceptance the method of the formation/education of broadband signals with the aid

of the carrier modulation frequency by binary pseudorandom sequences.
This sequence (Fig. 1.1.4) is formed by the video pulses of the
rectangular form of two forms (positive and negative, or otherwise
"1" and "0". The duration of these momentum/impulse/pulses is taken
identical (t_i), a the character of alternation it is determined by
the selected law of coding. These sequences were called the name
pseudorandom, since to outside observer the law of alteration was
unknown.

The formation of this pseudorandom sequence virtually most is
simple to carry out with the aid of the so-called shift registers,
which are devices for the memorization of multidigit binary number.

Fig. 1.1.4.
Key: (1) Period of Sequence.

Page 15.

As an example Fig. 1.1.5 gives the block diagram of three-stage

register, which includes three Origger circuits, clock pulse

generator and summator on module/modulus two. Each Origger circuit

has two steady states (" 0" or "1"). The shearing rate of information

(momentum/impulse/pulse) in register corresponds to the repetition

frequency of clock momentum/impulse/pulses. The summation over

module/modulus two is determined by the following table of logical

addition:

$$1+1=0$$
$$0+0=0$$
$$1+0=1$$
$$0+1=1$$

Let us assume that in the initial state "unit" it was recorded

in the first cell, i.e., in register was recorded combination 100. By

the first shift pulse from clock pulse generator the state of the

first and second cascade/stages is shift/sheared respectively in the

second and third cascade/stages. Being summarized by module/modulus

two, the information of the second and third cascade/stages it enters

the input of the first cascade/stage. Is obtained combination 010. In

the following stage to the first cell will enter the result of the

addition of the states of the second and third cells, and the state

of the first and second cells will pass to the second and third. Will

be obtained combination 101. Subsequently they will follow

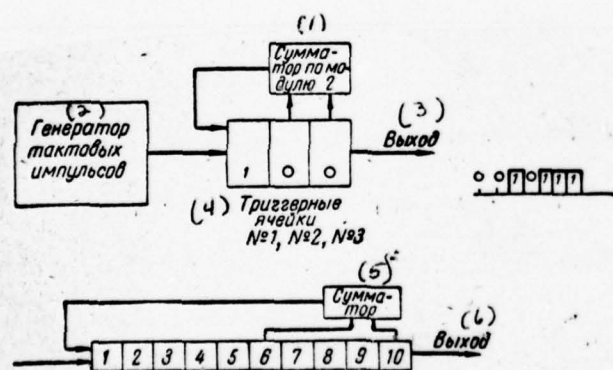combination 110, 111, 011, 001 and 100, whereupon cycle is repeated.

Fig. 1.1.5.

Fig. 1.1.5

Key: (1). Summator on module/modulus. (2). Clock pulse generator. (3). Output/yield. (4). Origger circuits. (5). Summator. (6). Output/yield. Page 16.

Output voltage (output code) is the sequence of the states of the last/latter (the third) cell and in our case takes the form of seven-symbol group 0010111 (Fig. 1.1.5). The durations of all momentum/impulse/pulses t_i are identical and are determined by the parameters of the diagram of Origger circuit. The number of cells can be increased, but to summator is not compulsory to connect the output/yields of the last/latter cells. In the same figure is shown the register of ten Origger circuits and with feedback on 6-s cell. Switching the cascade/stage, to which is included the feedback, changes output combination and is utilized usually for the exchange of the output sequence of momentum/impulse/pulses - code. The maximum number of the bits of code is determined from formula $m = 2^n - 1$, where n - the number of Origger circuits. Let us emphasize that this formula characterizes the maximum length of code, which is obtained with the determined connection of cells to summator. From the aforesaid it is clear that with the aid of a comparatively small number of cells in register it is possible to obtain the sufficiently prolonged cycle of work. For example, with m = 10 period will compose 1023 units.

The formed in a described manner sequence of video pulses is utilized for the formation of high-frequency broadband signal either by means of its direct supply to the balanced modulator, to another input of which they are supplied the fluctuation of carrier frequency or by means of supply to band-pass filter (Fig. 1.1.6).
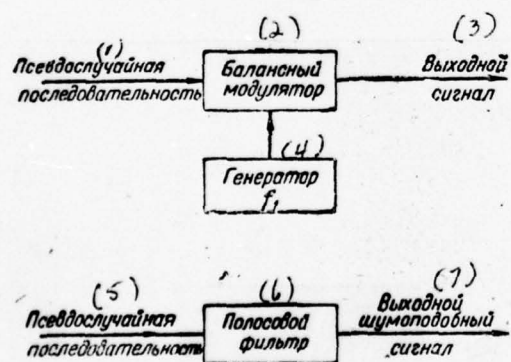
Fig. 1.1.6.

Fig. 1.1.6.

Key: (1). Pseudorandom sequence. (2). Balanced modulator. (3). Output
signal. (4). Generator. (5). Pseudorandom sequence. (6). Band filter.
(7). Output noise-like signal. Page 17.

In the first case at the output/yield of balanced modulator are
obtained the fluctuations of carrier frequency with phase jumps
according to the law of the sequence of momentum/impulse/pulses "0"
and "1" at the output/yield of register, i.e., the so-called
phase-code-keyed signals. Such signals find a use in discrete-address
or in mutually correlated systems.

In the second case, when the pseudorandom sequence of video
pulses enters the input of band-pass filter with band F and medium
frequency f_1, filter passes the only harmonic components, which lie
at band F, and as a result the output potential of filter obtains
noise-like character.

Mathematically serrated signal is conveniently presented in the
form Fourier series in the interval of the duration of signal (T)

$$z_r(t) = \sum_{k=0}^{\infty} (a_{rk} \cos k\omega_0 t + b_{rk} \sin k\omega_0 t),$$
$$0 \leqslant t < T,$$

where $\omega_0 = 2\pi/T$; k is a number of the component of signal (k = 0, 1, 2, 3, ...); r - the number of realization (for binary system r = 1, 2).

In the majority of the practical cases broadband signals for the radio communication are taken such, in which the only finite number of Fourier coefficients in the given formula differs from zero, but the number of realizations is equal to two (pressure, release). In these cases the expression for a serrated signal is reduced to the form

$$z_r(t) = \sum_{k=k_1}^{k_2} (a_{rk} \cos k\omega_0 t + b_{rk} \sin k\omega_0 t) =$$

$$= \sum_{k=k_1}^{k_2} A_{rk} \cos(k\omega_0 t + \varphi_{rk}), \quad 0 \leqslant t < T, \tag{1.1.1}$$

where $A_{rk} = \sqrt{a_{rk}^2 + b_{rk}^2}$, $\varphi_{rk} = -\arctan \dfrac{b_{rk}}{a_{rk}}$.

With this signal of component frequency they range from $\dfrac{\omega_0}{2\pi} k_1 = \dfrac{k_1}{T}$ to $\dfrac{\omega_0}{2\pi} k_2 = \dfrac{k_2}{T}$, where k_1, $(k_1 + 1)$, ... k_2 are integers.
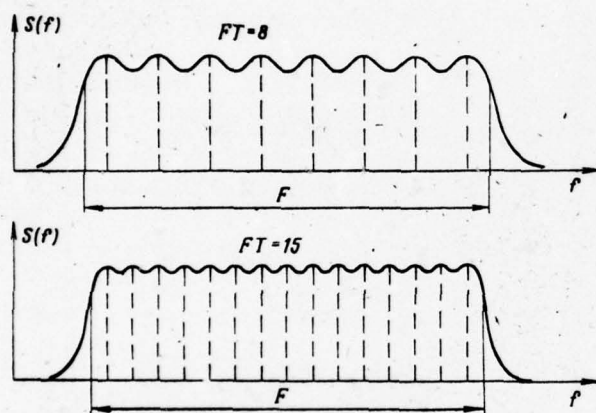
Page 18.

The frequency band, in which are contained the discrete

components of signal and the fundamental part of it of energy

spectrum, is determined in this case by the expression

$$F = \frac{k_2 - k_1 + 1}{T}.$$

The product of this value and the duration of the cell/element

of signal B = FT is accepted to call the base of signal. The value of

the base of signal characterizes the ratio of the width of the

spectrum of signal (F), that depends on the method of signal

conditioning (form of coding), to the width of the spectrum of the

report/communication, determined by the speed of transmission of

information, i.e., by the value of the reverse/inverse duration of

the cell/element of signal (1/T). For the signals, called broadband,

characteristic is the value of base, considerably greater than one.

Virtually values B reach 100-1000 and more.

Figure 1.1.7 shows energy spectrum S (f) of the signal, which is

expressed by formula (1.1.1), for two values of base. Than more FT,

those all large part of the energy of signal turns out to be that

concentrated in band F. For better/best understanding the question

concerning the base of signal, let us turn to examples.

Fig. 1.1.7

Page 19.

Let the report/communication is be the alternating current
(pressure) and noncurrent (release) premise/impulses by duration T
(period of manipulation 2T, the fundamental frequency of manipulation
1/2T). Such report/communications can be presented in the form of the
sum of the large number of sinusoidal fluctuations whose frequencies
are multiple to the fundamental frequency of manipulation. For the
satisfactory reproduction of report/communication with reception it
suffices to transmit only its low-frequency components in the band of
frequencies $F \approx 3/2T$, called the spectrum of report/communication.
During the off-on modulation of transmitter this report/communication
the frequency band, occupied by the transmitter, will be two times
wider than the spectrum of report/communication, i.e., it will
comprise 3/T. With single-band modulation it is equal to 3/2T, i.e.,
to the band of the report/communication itself.

Thus, the base of signal in the first case is equal to two, and
in the second case - unity. The values of the bases, close to unity,
are characteristic for all narrow-band systems. For broadband systems
as already it was said, values B reach several hundreds and even
thousand.

Will give the estimate/evaluation of the frequency band,
occupied by transmitter at the different speeds of transmission of
discrete information and the different values of base. If we assign
the speed of transmission of discrete information 50 bauds (T = 20
ms), then for using a signal with B = 1000 will be required the band
of frequencies $F_{(50)} = \frac{B}{T} = \frac{10^3}{20 \cdot 10^{-3}} = 50$ kHz. At speed 1200 bauds and the
same value of base will be required the band of frequencies $F_{(1200)} = 1,2$
MHz, which in skip band cannot be carried out virtually. If necessary
to provide such the high rates of the work of value B they are
selected smaller (for example, 100).

From the aforesaid it is clear that the larger values B can be
provided for at the lower speed of work. Figure 1.1.8 graphically
shows the interdependence of the rate of the work of the band of
frequencies (F) and of base.

Page 20.

As we shall see from the further paragraphs of this chapter, the
major advantages of the use of broadband signals are caused by the
precisely large value of their base. Hence it follows that if there
are limitations in the question of the expansion of the band of
signal (but they always appear as a result of the technical
difficulties of designing of broadband receivers, transmitters,
antennas and due to a difference in the conditions of radiowave
propagation within limits of very broad band), then the advisability
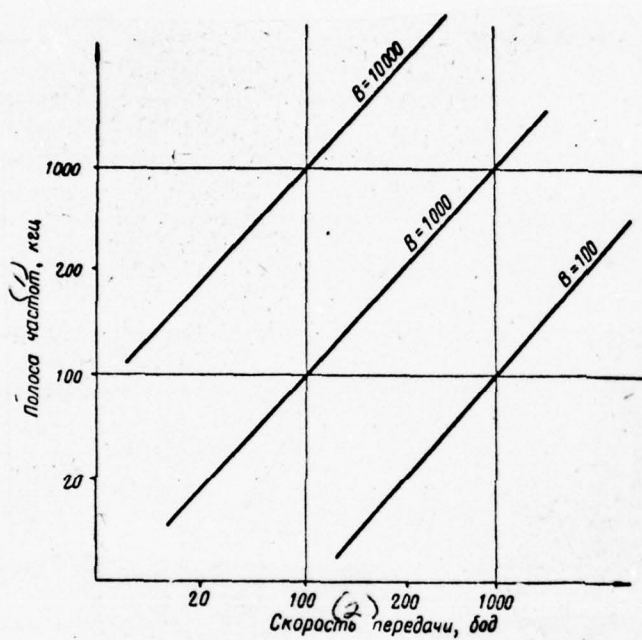of passage to broadband systems considerably more at the lower speed
of work.

Fig. 1.18.

Fig. 1.1.8.

Key: (1). Band of frequencies, kHz. (2). Speed of transmission, bauds. Page 21.

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§1.2. On the possibility of the simultaneous work of a series of broadband systems in just one section of range and the joint operation of broadband and narrow-band systems.

At first glance it can be shown that the passage to the use of broadband systems for purposes of communication/connection decreases the number of radio stations, which can simultaneously, also, with that without interferences work in the assigned section of range. On the matter itself this not entirely so, since it turns out to be possible in one and the same the frequency band to simultaneously work to a comparatively large number of transmitters. Their selection at receiving end will be in this case realized not in frequency, but in form of signal. The possible difference in the waveforms can be very large. It is the greater, the greater the base of signal. Consequently, the more the base of signal, i.e., the frequency band, occupied by transmitter at the given operating speed, the more number of transmitters it can simultaneously work in one and the same the

frequency band. In this case, of course, is assumed the

application/use at the receiving end of so-called optimum processing

signal with the aid of matched filters or correlation diagram.

Let us explain first in a simple example of the possibility of

the formation of the broadband signals of various forms and their

selection at receiving end. Let be realized the telegraph work at the

definite speed, by which the duration of the cell/element of

report/communication is equal to T. We convert each such cell/element

of report/communication with the aid of the diagram, shown in Fig.

1.1.2, into the series the following in the determined sequence more

narrow pulses of length τ. The law governing the following of these
pulses for each realization of the cell/element of report/communication
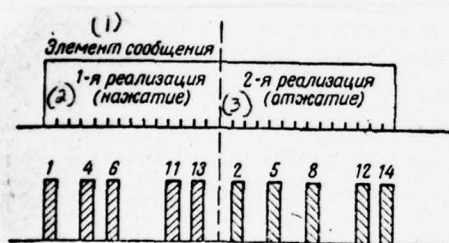(pressure, release) let us accept different (Fig. 1.2.1).

Fig. 1.2.1.

Key: (1). Communication element. (2). 1-st realization (pressing).
(3). 2-nd realization (release). Page 22.

If we high-frequency oscillations key by the obtained narrow

pulses, then the frequency band will be in T/τ times wider than with

modulation of high-frequency oscillation directly the

momentum/impulse/pulses of communication by duration T. For another

correspondent, who works in the same frequency band, let us select

other sequences of narrow pulses both for the release and for a

pressure. In order that the correspondents it would be possible to

distinguish, we utilize on the receiving end as matched filters of

delay line in removal/outlets (Fig. 1.2.2). In this matched filter

the momentum/impulse/pulse, which entered to input, passes from the

beginning of line to each removal/outlet for the accurately specific

time.

Removal/outlets from delay line for the isolation/liberation of

each realization of signal are placed so that when the first of the

momentum/impulse/pulses of signal appears at the last/latter

removal/outlet of delay line, the second momentum/impulse/pulse will

arise on penultimate, the following - on the third from end/lead,

etc. Since all removal/outlets are connected in parallel, all these

momentum/impulse/pulses store/add up (it is realized roll) at the

torque/moment of the termination of the cell/element of signal
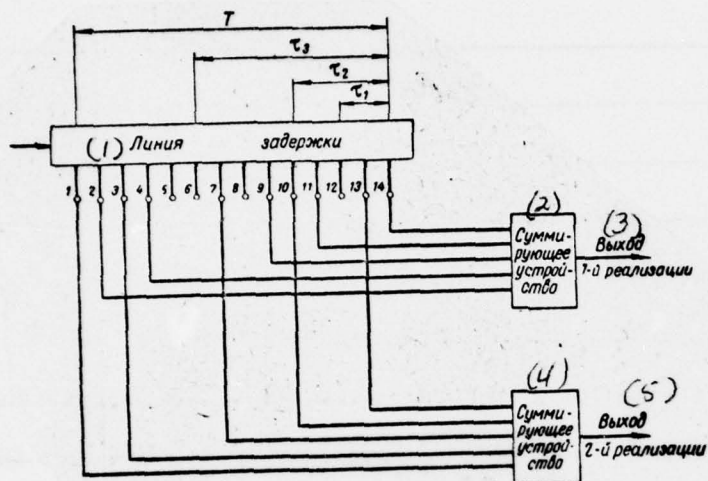
(premise/impulse by duration T).

Фиг. 1.2.2.

Fig. 1.2.2.

Key: (1). Delay line. (2). Adding device. (3). Output/yield of 1-1 realizations. (4). Adder. (5). Output/yield of 2-1 realizations.

If of the input of the datum of delay line enter the momentum/impulse/pulses, transmitted in another sequence (momentum/impulse/pulses of another realization or another correspondent's momentum/impulse/pulses), then on the removal/outlets of line will appear momentum/impulse/pulses into the noncoincident torque/moments of time and stress it will turn out to be small. Rolls it will not occur. The appearing in this case stress is accepted to call stress from system interferences. It is understandable that for data acquisition correspondent in binary communicating system it is necessary to have two matched filters, for another correspondent's reception - two other filters, etc.

The shorter will be the momentum/impulse/pulses i the more respectively it will be their number for time T, i.e., the more will be the frequency band or the more the base of signal, fact those more differing one from another combinations can be used for the

formation/education of different radio channels in one and the same the frequency band. Is clear also the fact that, the greater the transmitters it will to simultaneously work in one and the same the frequency band, the higher will be the interference level, created by other transmitters at the output/yield of receiver.

In order that interferences among two signals z_1 (t) and z_2 (t), simultaneously entering the input of matched filter, would be possibly smaller, must be made the condition of the orthogonality of these signals under range from 0 from T. Mathematically the condition cf orthogonality is record/written as follows:

$$\int_0^T z_1(t)\, z_2(t) = 0. \qquad (1.2.1)$$

As the simplest example of orthogonal signals can serve, for example, the signals of the form:

$$z_1(t) = \begin{cases} A\cos(\omega t + \varphi), & \text{when } 0 \leqslant t < \frac{T}{2}; \\ 0 & , \text{when } \frac{T}{2} \leqslant t < T; \end{cases}$$

$$z_2(t) = \begin{cases} 0 & , \text{when } 0 \leqslant t < \frac{T}{2}; \\ A\cos(\omega t + \varphi), & \text{with } \frac{T}{2} \leqslant t < T. \end{cases}$$

Page 24.

During the formation/education of signals out of the large
number of momentum/impulse/pulses for the example, shown in Fig.
1. 1.2 or 1.2.1, the condition of orthogonality is satisfied, if the
momentum/impulse/pulses of each realization are supplied to the
noncoincident moments of time. Orthogonal they will be two signals z_1
(t) and z_2 (t), also, when they are represented in the form of the
nonoverlapping Fourier series in range from 0 to T. The
nonoverlapping Fourier series here called two such series, that if in
one of them $a_k \neq 0$, that in other for the same index $a_k = 0$;   the
analogous relationship/ratios must be made and for coefficients b_k.

In the general case the signals can be orthogonal and answer condition (1.2.1) and then, when their spectra overlap.

Let us allow now, which in the band of frequencies F is by n of different orthogonal signals. one of these signals useful, and the others with respect to it mixing. The power of all signals for simplicity of reasonings we consider identical and equal to P_c. With the sufficiently large number of interfering signals their sum can be considered as fluctuating interference with total power $P_c(n-1)$ or spectral density $\frac{P_o(n-1)}{F}$.

From the theory of communication/connection it is known that the authenticity of reception during optimum processing signal is determined by the ratio of the energy of signal to the spectral power of fluctuating interference, i.e., by value

$$h^2 = \frac{P_c T}{(n-1)\frac{P_o}{F}} \approx \frac{FT}{n}.$$

If requirements for the authenticity of reception are satisfied at certain value $h^2 = h^2_{\text{тр}}$ that permissible number of simultaneously working in one and the same band frequencies of the radio stations will be

$$n \approx \frac{FT}{h^2_{\text{тр}}}.$$

From this expression directly it follows that the number of simultaneously working stations in this frequency band with the determined assigned authenticity of reception is proportional to the base of signal.

Page 25.

Hence escape/ensues the possibility in principle of operational
provisions in the defined band of frequencies of the same number of
radio stations as when using narrow-band systems. Actually, in the
case of single-band modulation for the duration of the cell/element
of signal T the band of report/communication $F \approx 1/T$, the base of
signal FT = 1 and in the band of frequencies Δf can be placed by $n \approx$
$\Delta f/F$ transmitters. For the same operating speed the base of broadband
signal in band Δf is obtained $\Delta f/F$ times more and, consequently, also
number of transmitters in band it will be $n \approx \Delta f/F$; in this case
there is in form an identical value $h_{\text{тр}}^2$. This comparison, of course,
approximated, but it makes it possible to understand that the
essential losses in the use of a range passage to broadband signals
with themselves will not bear.

By examining the question concerning the use of a frequency
range during the application/use of broadband and narrow-band
systems, it is necessary also to give the considerations, caused by
the fact that in actual conditions never all stations of this range
work simultaneously. Taking into account the indicated fact in
broadband systems considerably simpler to increase the number of

radio stations in the assigned frequency band, than in narrow-band systems. Let us explain this.

When using widespread methods of selection in frequency and narrow-band systems to each routing is secured the determined frequency channel. This frequency channel cannot be (at least, if we do not apply the complex systems of the automatic search for free channels) engaged by another correspondent, even if it at the given torque/moment is free, since compulsorily after each communication line must be preserved the possibility to begin work at any time. The unemployment of this channel to no extent does not improve communication/connection conditions for other channels. When using broadband systems, if at each given moment of **time of the total** number n of radio stations it works not more than $n' = \eta n$ (η - **the** diversity factor of work), the number of simultaneously working stations **in this frequency band** can be increased $\frac{1}{\eta}$ **times.** Set/assuming $\eta = 0.1-0.2$, we obtain supplementary **very con**siderable possibilities on an increase in the effectiveness of the use of a frequency range without the application/use of fairly complicated actions for the provision for a search for the empty channels and the retuning of equipment in these empty at the given torque/moment channels. Let us focus attention also to the fact that in broadband systems a decrease in the amount of simultaneously working stations

automatically improves communication/connection conditions for other

channels. This advantage of the work with broadband signals becomes

especially noticeable at the low values η, i.e., when the majority of

the stations of this range works short-term. Specifically, this

situation frequently occurs under conditions of military radio

communication.

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Page 25.

§ 1.3. Broadband systems under conditions of multiple-pronged radiowave propagation.

The majority of radio channels, especially short- and ultrashort-wave, is characterized by the multiple-pronged propagation during which the signal from transmitter comes to the place of the reception/procedure by several paths, whereupon for each of these

paths signal experience/tests different attenuations and different time lag. If during single-ray propagation the dependence between that which is taken and that transmitted by signals is determined by the expression

$$x(t) = \mu z(t - \Delta t), \qquad (1.3.1)$$

in which value μ and Δt they characterize attenuation and signal lag in the process of its propagation from transmitter to receiver, then during multiple-pronged propagation this dependence takes the form

$$x(t) = \sum_{i=1}^{n} \mu_i z(t - \Delta t_i), \qquad (1.3.2)$$

where n - the number of incoming ray/beams.

In this expression μ_i and Δt_i they are determined by attenuation and the time lag of each separate ray/beam. Values μ_i

and Δt_i, as a rule, continuously change in time, although for the larger part of the real signals and at the in practice utilized speeds of transmission of information their change in time is sufficient slowly as compared with the duration of transmission of the cell/element of signal.

Page 27.

During multiple-pronged propagation at the input of receptor, therefore, operates the sum of separate oscillation/vibrations with the changing according to random laws phases and amplitudes. This leads to the phenomenon of interference, and also, therefore, to signal fading. The phenomena signal fading are expressed especially vividly, when path differences of ray/beams are commensurable or multiple to the half of wavelength.

They distinguish several types of signal fading: common/general/total or flat, selective, slow and rapid. With flat fadings the incoming signal differs from that transmitted by the random, but approximately identical for all frequency components of signal values of the transmission factor and phase displacement. In

selective fadings of each frequency component of signal corresponds
their **transmission factor and** their phase displacement. Slow fadings
occur, if μ_i and Δt_i **they are** characterized approximately by
identical values for the extent/elongation of the reception/procedure
of several cell/elements of the signal, following one after another.
For rapid fadings is characteristic **the absence** of interdependence
(correlation) between values μ_i and Δt_i **for a** series of the
consecutively transmitted cell/elements **of signal.**

Let us examine in somewhat more detail the special
feature/peculiarity of common/general/total and selective fadings.

Let in accordance with expression (1.1.1) transmit the signal

$$z(t) = \sum_{k=k_1}^{k_2} A_k \cos(k\omega_0 t + \varphi_k).$$

Then the received signal

$$x(t) = \sum_{i=1}^{n} \mu_i \sum_{k=k_1}^{k_2} A_k \cos[k\omega_0(t-t_{pi}) + \varphi_k] =$$

$$= \sum_{i=1}^{n} \mu_i \sum_{k=k_1}^{k_2} A_k \cos[k\omega_0(t-\overline{t_p}) + \psi_{ik} + \varphi_k]. \tag{1.3.3}$$

where $\bar{t}_p$ is the mean propagation time for all ray/beams: $\psi_{ih} = 2\pi \frac{k}{T} \Delta t_i$; Δt_i - the deflection of the propagation time of the i ray/beam from the mean propagation time $\bar{t}_p$.

Page 28.

Values ψ_{ih} for any ray/beam will range from $2\pi \frac{k_1}{T} \Delta t_i$ to $2\pi \frac{k_2}{T} \Delta t_i$; if we

$$|\Delta t_i| \ll \frac{1}{F} = \frac{T}{k_2 - k_1 + 1}, \qquad (1.3.4)$$

that they will differ one from another not more than by $2\pi F \Delta t_i \ll 2\pi$.
That means with the observance of condition (1.3.4) it is possible to
count values ψ_{ik} approximately identical for all values of k, i.e.,
for all frequency components of signal. Thus, condition (1.3.4)
causes the flat character of fading. Specifically, let us note that
common/general/total fadings occur both with ionospheric (Fig.
1.3.1a) and during tropospheric (Fig. 1.3.1d) radiowave propagation,
provided path differences of the adopted ray/beams were much less
than value 1/F. Selective fadings occur when condition (1.3.4) is not
satisfied, i.e., path differences of ray/beams Δt_i are commensurable
or exceed value 1/F. For selective fadings of the value of phases
are different for different k. Then received signal x (t) is
represented in the form

$$x(t) = \sum_{k=k_1}^{k_2} A_k \sum_{i=1}^{n} \mu_i \cos\left[k\omega_0\left(t - \overline{t_D}\right) + \psi_{ik} + \varphi_k\right] =$$
$$= \sum_{k=k_1}^{k_2} \mu_k A_k \cos\left[k\omega_0\left(t - \overline{t_D}\right) + \varphi_k + \Theta_k\right]. \tag{1.3.5}$$

where

$$\mu_k = \sqrt{\left[\sum_{l=1}^{n} \mu_l \cos \psi_{ik}\right]^2 + \left[\sum_{l=1}^{n} \mu_l \sin \psi_{ik}\right]^2};$$

$$\Theta_k = -\arctan \frac{\sum_{l=1}^{n} \mu_l \sin \psi_{ik}}{\sum_{l=1}^{n} \mu_l \cos \psi_{ik}}.$$

Page 29.

From this expression it follows that with **selective fadings** of the fluctuation of amplitudes $\mu_k A_k$ and of phases $(\varphi_k + \Theta_k)$ they depend on the frequency (index **k**) of **the harmonic** components of **signal**. Usually such fadings appear when to receptor come the ray/beams, which reflected from the different ionospheric layers (Fig. 1.3.1b)

or of the spaces of the troposphere, and also the undergone multiple
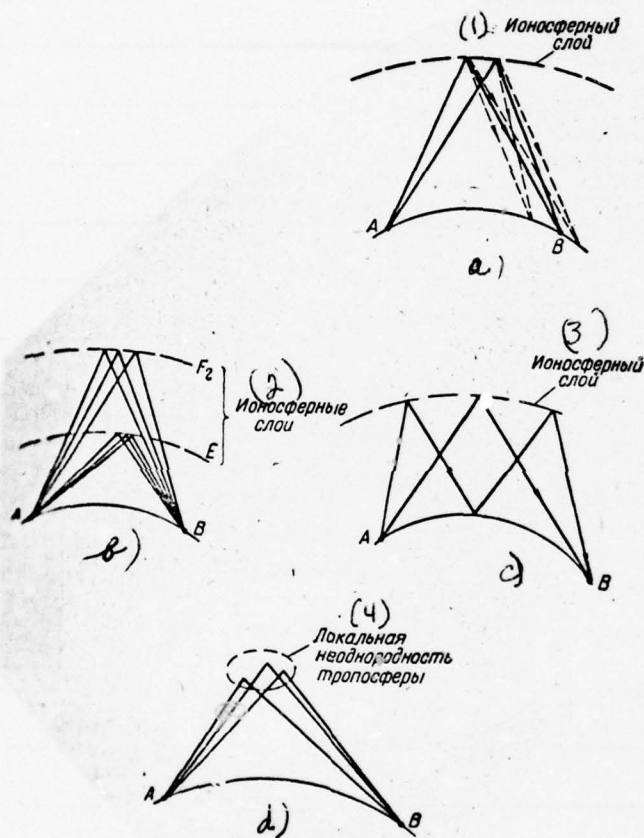reflections (Fig. 1.3.1c).

Fig. 1.3.1.

Fig. 1.3.1.

Key: (1). Ionospheric layer. (2). Ionospheric layers. (3).
Ionospheric layer. (4). Local heterogeneity of the troposphere.

In this case, as a rule, each come ray/beam is the beam of elementary
ray/beams and therefore is also subjected to fadings, which bear,
however, common character [1].

FOOTNOTE [1]. In the literature very frequently instead of the term
"channel with selective fadings" are used the terms "multiple-pronged
channel, "channel with multiple-pronged propagation". Thereby is
emphasized the fact that the channels with selective fadings are
characterized by the arrival of signal cn several paths with
considerable path differences. Subsequently this terminology widely
is utilized in the present work. ENDFOOTNOTE.

When values Δt_i are commensurable with the duration of the
cell/element of signal T, the phenomenon of multiple-pronged
propagation causes not only signal fading, but also it leads also to

the imposition of the adjacent cell/elements of signal on each other.
This phenomenon is accepted to call the phenomenon of echo. In Fig.
1.3.2 is illustrated the picture of the overlap of the cell/elements
of signal with its arrival on several paths.

For narrow-band radiolink systems, when the base of signal $FT \approx$
1, values 1/F and T prove to be one order, therefore, of the
phenomenon of selective fadings and echo are observed simultaneously.
For the broadband systems, in which $FT \gg 1$, are possible the cases,
when Δt_i of the same order as 1/F, but is considerably less than T;
in these cases will be the only selective fadings.

Both common/general/total and selective fadings cause the
fluctuations of the level of received signal with the average period
from several minutes to fractions of a second [13]. However, they are
not the sole reason for a change in the intensity of received
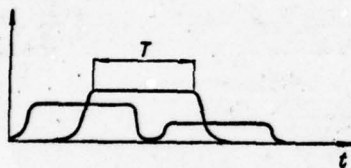signals.

Fig. 1.3.2.

Page 31.

In the information circuits are observed also the relatively slower

changes (hour, diurnal) in the level of received signals, caused not

by interference processes in receiving antenna, but by effect on the

transmission factor of the channel of such reasons as change of the

magnitude of absorption in the ionosphere, a change in the

temperature conditions of the troposphere, etc.

Finally, in the case of using signals with the wide spectrum the important factor, which affects received signal, is also the dispersity of the medium, which participates in radiowave propagation (ionosphere, the troposphere), that leads to the fact that the coefficient of reflection or scattering finds to be dependent on frequency, i.e., different for the different frequency components of signal, even for a "single-ray" channel. Let us note that virtually during ionospheric propagation the dispersive phenomena cause the noticeable differences in transmission factor for frequencies, which differ by dozen kilohertz (for a skip band), while the values of this coefficient because of the interference of ray/beams turn out to be sometimes different for frequencies, which differ only by hundred hertz. If transmits sufficiently narrow pulse, then as a result of dispersion in the ionosphere the momentum/impulse/pulse of separate ray/beam is eroded, and multiple-pronged propagation leads to the fact that instead of one momentum/impulse/pulse are accepted several that and it is clarified in Fig. by 1.3.3.
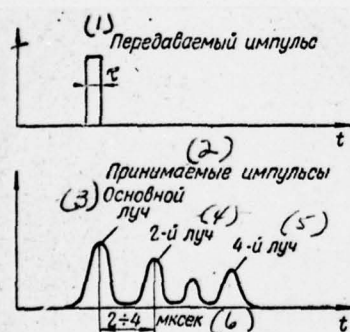
Fig. 1.3.3.

Key: (1). Transmitted momentum/impulse/pulse. (2). Received pulses.

(3). Fundamental ray/beam. (4). 2-nd ray/beams. (5). 4-th ray/beams.

(6). ms. Page 32.

The interval between the incoming ray/beams can reach by 2-4 ms, and the number of ray/beams, especially for the extended short-wave lines, to reach to four - six.

The phenomena of fading, echo and the effect of the dispersity of the medium of propagation in the practice of radio communication lead to a considerable decrease in freedom from interference or authenticity of reception/procedure (with this capacity) they are considered as negative factors. When using multiple-pronged channels always they attempted to remove the effect of "parasitic" ray/beams in order to approach itself work conditions of single-ray channel. This path, however, was not sufficiently rational, since each of the incoming ray/beams contains the information about the transmitted report/communication and their use it could improve in principle the conditions of reception/procedure.

One of the important advantages of the use of broadband signals is a possibility of using a phenomenon of multiple-pronged propagation for an increase in the stability of reception/procedure

or with the assigned authenticity of reception/procedure for a
decrease in the power of radio transmitting equipment. This
possibility is open/disclosed as a result of the fact that the
broadband systems can ensure the separate reception of the signals,
which come in to the place of reception variously, and therefore they
make it possible to utilize an energy of several most intense
ray/beams. Let us explain this in a simple example.

Let, for example, the signals of the form, given in Fig. by
1.2.1, enter the matched filter of Fig. 1.2.2. Input voltage of
reference system, i.e., on the output/yield of matched filter, with
single-ray reception will have a shape of pulse as duration τ, i.e.,
1/F, the appearing into torque/moment termination of the cell/element
of signal by duration T. Here F is the band of signal, which is
determined by the duration of narrow pulses τ.

Let us allow now, that into the point of reception the signal
comes by several paths with the different values of transit time from
transmitter to receiver $t_{p1}, t_{p2}, t_{p3}, \ldots, t_{pn}$. In this case to each ray/beam
at the output/yield of the matched filter will correspond its
momentum/impulse/pulse with the maximum of the voltage/stress at the
torque/moment of time $t_{pl} + T$ and of the same duration 1/F.

Page 33.

Not difficult to understand that if the difference in time of the
arrival of ray/beams is less than the duration of these
momentum/impulse/pulses (τ), then the separate reception of ray/beams
turns out to be impossible. The separate reception of the signals,
which come in on different paths, is feasible only if value τ less
than the minimum delay time between ray/beams, i.e., the possibility
separate the reception of ray/beams increases with decrease τ or,
otherwise, with the expansion of the band of signal frequencies.
Furthermore, the maximum delay time among ray/beams must be less than
the duration of the cell/element of signal T.

In narrow-band systems the pulse duration at the input of
reference system is approximately equal to the duration of the
cell/element of signal T, and consequently, the examined above
conditions, necessary for the separation of ray/beams, cannot be
satisfied. Under broadband systems the given conditions due to
inequality $\tau \ll T$ are satisfied easily.

Let us note that the pulse character of the signal, in example
of which we examined the possibility of isolation/liberation in
receiver of one of the incoming ray/beams, it is not necessary. This
possibility is provided in all cases, when the base of signal FT >> 1
and signal has approximately uniform spectral density in the band of
frequencies F.

§1.4. Classification of broadband radiolink systems.

Concept "broadband radiolink systems" includes at present the
large number of versions of such systems, which differ one from
another from in form of the utilized signals, method of their
formation, the method of reception and to a series of other
sign/criteria. Respectively the fields of application of broadband
communicating systems also are very diverse. This causes the need for
the determined classification.

The representation of the possible classification of broadband

systems gives the table, given in Fig. by 1.4.1.

The large group of broadband systems form the so-called mutually correlated active and passive systems. In mutually correlated active systems the signal is recorded in receptor by means of comparison with the standards, formed on the same principle, as at the transmitting end/lead.
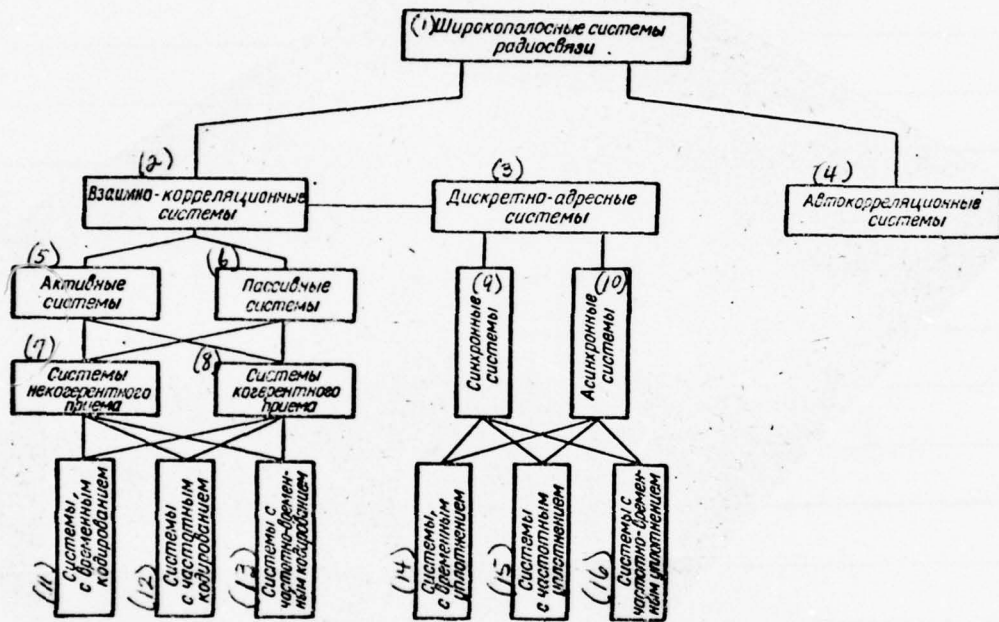
Fig. 1.4.1.

Fig. 1.4.1.

Key: (1). Broadband radiolink systems. (2). Mutually correlated systems. (3). discrete-address systems. (4). Autocorrelation systems. (5). Active systems. (6). Passive systems. (7). Systems of incoherent reception. (8). Systems of coherent reception. (9). Synchronous systems. (10). Asynchronous systems. (11). Systems, with time/temporary coding. (12). Systems with frequency coding. (13). Systems with frequency-time coding. (14). Systems, with time-division multiplex. (15). Systems with frequency division multiplex. (16). Systems with frequency-time packing/seal.

Page 35.

Systems of such type make it possible to most completely realize potential interference rejection. Multiple-beams in them does not limit the speed of transmission. At the same time it is necessary to indicate some factors, which impede and limiting their practical realization. The greatest difficulties are caused by the need for precise synchronization of reference signals, developed in receptor, with those which are taken (accuracy of synchronization must be order 1/F for an incoherent reception and still above in the case coherent). For these systems is characteristic also relatively long in comparison with other systems time of netting, that lowers the reticence of work and impedes the work with short performances, and

the relative complexity of the practical solutions by the use of multiple beam characteristics, which leads to an increase in the overall sizes of equipment, the costs and the complication of its operation.

By mutually correlated passive systems they understand such, whose information about received signals is laid in the characteristics of the matched filters of receptor. Such systems also make it possible to realize potential interference rejection, and multiple-beam characteristics in them it does not limit the speed of transmission of information. In principle these systems for the realization of the optimum conditions of reception require synchronization with the accuracy of order 1/F. To synchronize is necessary the torque/moment of the reading of output output potential of matched filter. However, these systems, after allowing small energy loss, it is possible to forego synchronization to generally or considerably lower requirement for its accuracy. The realization of such systems leads also to sufficiently bulky equipment/devices and is conjugate/combined with the determined technical difficulties. In more detail the operating principles of mutually correlated systems and path of their realization are examined in chapter 3. An example of mutually correlated system is the system "Rake" [44].

Both active and passive mutually correlated systems can be
divided to two classes: with coherent and with incoherent by the
receptions of signals.

Page 36.

As it was shown into §1.3, in the general case the adopted broadband
signal it is represented in the form

$$x(t) = \sum_{i=1}^{n} \mu_i \sum_{k=k_1}^{k_1} A_k \cos \left[k\omega_0 \left(t - \overline{t_p} \right) + \psi_{ik} + \varphi_k \right]. \qquad (1.4.1)$$

For a coherent reception at receiving end must be known the
amplitudes (relative values) of all components of signal A_k, the
initial phases φ_k, the phase shifts of each component ψ_{ik} and the
mean propagation time for all ray/beams $\overline{t}_p$. With incoherent reception
must be known all the same values, except the initial phases. The
initial phase of the incoming signal in this case either is unknown

or it is not utilized and, inasmuch as in the initial phase of received signal also is contained certain information about the transmitted report/communication, utilized with coherent reception, the probability of error for the methods of incoherent reception proves to be larger.

The theoretical examination of a comparative freedom from interference of systems of coherent and incoherent reception shows that at the high values of the authenticity of reception, determined by the probability of the error of order $10^{-3}-10^{-4}$, the energy loss of incoherent systems does not exceed 1 dB. At the same time the practical realization of the diagrams of coherent reception is conjugate/combined with the considerable technical difficulties, which moreover are aggravated by instability in time of the phase responses of the medium of propagation. In connection with this in certain cases are more advisable the systems of incoherent reception.

Another group of broadband systems form the systems autocorrelation whose freedom from interference is worse than mutually correlated. However, they differ significantly from the latter in terms of considerably larger simplicity of realization and in terms of respectively higher reliability and the compactness of

equipment. Very is substantial also the fact that the autocorrelation

systems little subjected to the ill effect of multiple-beam

characteristics and dispersity of the ionosphere and, as a rule, do

not require the high frequency stability of radio link.

Autocorrelation systems at the same time have that advantage that in

them as carrier it is convenient to utilize cuts of white noise,

i.e., to utilize in the complete sense of word "noise-like" signals.

Page 37.

Let us note that in all mutually correlated systems are utilized the

signals, which rather one should call/namenoise-like, since in

reality they are completely regular and the precisely precise

knowledge of their structure makes it possible to successfully

isolate them from noises in receptor.

The comparison of the freedom from interference of the systems

of autocorrelation reception with the systems of incoherent reception

shows that their energy loss turns out to be equal approximately

$\sqrt{FT}$, i.e., by sufficiently considerable. Their use therefore turns

out to be advisable only if the advantages, noted above, have more

important value. By an example of autocorrelation system is the
system with so-called korrel4qionno-time/temporary modulation,
proposed in 1959 in GDR [21, 49], examine/considered in detail in
chapter 4.

One Additional Group of broadband radiolink systems form the
so-called discrete-address systems, which are in essence mutually
correlated and those found use application/use for an army
radiotelephone circuit in VHF range. In these systems in transmitter
each informational cell/element is coded by the dialing/set of
signals, intended only for a communication/connection with the
determined subscriber. Receptor isolates the only signals, intended
to this correspondent, and it does not accept others, although they
are emitted in the same frequency band. As signals here usually is
utilized the dialing/set of short radio pulses. In this case can be
used in principle any form of the pulse modulation: APM_A, DPM_A pulse [АИМ] [ПИМ]
position modulation$_N$, pulse-frequency modulation$_A$ KIM_A, delta [ФИМ] [ЧИМ] [КИМ]
modulation, etc.

Discrete-address systems are of two types - synchronous and
asynchronous. In synchronous systems are applied orthogonal signals.
The transmission of information from one subscriber here is strictly

synchronized. Synchronous systems can be both with time/temporary and with frequency division multiplex. During the time-division multiplex between orthogonal signals is provided the constant temporary displacement and is allow/assumed the overlap of the spectra.

Page 38.

During frequency division multiplex the spectra of signals must not overlap, but the orthogonality of signals must be provided in any temporary situations of signal.

In asynchronous discrete-address systems usually are applied nonorthogonal signals. Let us note that during the operation of the discrete-address radio communication systems of the large number of subscribers is observed the phenomenon of self-regulation. This is caused by the fact that with an increase in the number of subscribers increases the level of interferences and, therefore, deteriorates the quality of communication/connection, in consequence of which the subscribers turn out to be those which were forced to speak slower or to cease work. In this case for the remaining subscribers the quality of communication/connection is improved. Those, who do not need

urgent communication/connection, automatically will wait the more light/lung conditions of communication/connection. Systems of such type are examined in greater detail in chapter 5.

§1.5. On the reticence of broadband communicating systems.

In a number of cases to communicating systems is presented the requirement for reticence. An especially important value this requirement acquires for the systems of army radio communication.

The general requirement for the reticence of radio communication falls into three independent requirements: first, the provision for a reticence of the very fact of the work of radio link (transmitter); in the second place, the provision for a reticence of the fact of the presence in this signal of information; thirdly, the provision for a reticence of information itself.

All these three requirements broadband radiolink systems satisfy, as a rule, somewhat larger degree than usual, narrow-band. As has already been spoken above, in broadband systems are applied

the serrated signals with base FT >> 1. It proves to be that when using such signals the reception of discrete information can be realized in the power of signal, per unit of the frequency band smaller than the spectral power of the fluctuating interferences in the utilized band of frequencies, i.e., seemingly go under noise. In other words, it turns out to be possible to conduct the reception, when signal to usual narrow-band receiver simply is not audible. Let us explain this in somewhat more detail.

Page 39.

With the piece-by-piece incoherent reception of discrete information the probability of error for a binary system with active pause (in the case of channel with the constant parameters) is determined by the expression

$$p = \frac{1}{2} e^{-\frac{h^2}{2}}, \qquad (1.5.1)$$

where $h^2 = \frac{P_c T}{v^2}$ - the ratio of the energy of signal $(P_c T)$ to the

spectral density of fluctuating interference (v^2).

Consequently, under conditions of the effect of fluctuating interferences the probability of the error unambiguously is determined by value h^2, which characterizes the excess of the signal above interference.

At the same time

$$h^2 = \frac{P_0 T}{v^2} = \frac{P_0}{P_\pi} FT, \qquad (1.5.2)$$

where P_π is power of fluctuating interference. From this expression it follows that in any relation $\frac{P_0}{P_\pi}$ it is possible to fit this value FT (base of signal), which will ensure the required for obtaining the assigned authenticity of reception value h^2. For example, for providing a probability of the errors of order 10^{-4}–10^{-5} in the system of incoherent reception in accordance with expression (1.5.1) desired value $h^2_{\tau p} \geqslant 20$. With the base of system more than twenty relation $\frac{P_0}{P_\pi}$ is retained less than unity. For the confident

reception under conditions of a small relationship/ratio $\frac{P_0}{P_\pi}$ will be required an increase in the base of signal, which under practical conditions is achieved either by the expansion of band (F), or by a decrease in the velocity of the transmission of information (by increase T).

The reticence of the presence in the signal of information for broadband communicating systems is provided (of course, only to a certain extent) first of all by the noise-like quality of signal, obtained as a result fairly complicated and almost in each case of the specific coding.

Page 40.

With reception to usual narrow-band receiver and to the broadband receiver, in which is not utilized the information about the method of signal conditioning of this broadband transmitter, its signal will be received as usual noise, whereupon transition from one realization to another (pressure, release,) will not be noticeable. Recall again that this noise can be in a number of cases even lower the level of usual fluctuating interferences. The greatest reticence of the fact

of the presence of information in noise signal, apparently, can be reached in the systems, where as carrier are utilized the cuttings off of white noise.

The reticence of information itself in broadband systems is reached by sufficient complexity of the coding of signal and by comparative simplicity of the exchange cf coding. For information recovery is necessary the completely determined coherent transformation of signal, which it is unknown in the point/item of interception, and the parameters of this transformation can change according to the concealed/latent before enemy program with sufficient speed and complexity. It is natural that the extraction of information even from the fixed signal in such cases requires to have fairly complicated analyzing equipment, especially if one considers that at the large values of base FT the number of possible for use waveforms turns out to be very large.

From the aforesaid it follows that the broadband radiolink systems, as a rule, are self-contained in the sense that the signals, intended to this correspondent, remain unavailable for other.

Page 41.

Chapter 2

RADIO JAMMINGS. Optimum methods of radio reception.

§2.1. General noise characteristic of radio communication.

one of the most important requirements, presented to the systems
of military radio communication, is the correctness of the
information, transmitted from information source to recipient. Under
practical conditions fulfilling this requirement unavoidably they
block the various kinds of the interferences, which are caused by the
following factors:

- by the outside interferences, which enter the input of
receptor from the communication channel;

- by the internally-produced noise, which appear in the quite
receptor;

- by the radio-signal distortions, connected directly with the
passage of signal along channel.

Let us examine each of these factors in somewhat more detail.

Outside interferences appear as a result of the different
natural electromagnetic processes, which take place in the
atmosphere, the ionosphere and outer space (atmospherics, cosmic
noises, etc). Furthermore, they are created various kinds of
electrical devices (the so-called man-made interferences) and
numerous foreign radio stations. finally, outside interferences
(especially on the lines of military radio communication) can be
called the special jamming transmitters, used at the point of enemy
for the disorganization of the work of radio aids.

Internally-produced noise of receptor appear as a result of the
chaotic thermal electron motion and ions in the cell/elements of the
receiver itself.

Page 42.

The fundamental of these noise sources are electron tubes, the
semiconductor devices, the resistor/resistances and other
cell/elements of diagram.

Both outside interferences and internally-produced noise of

receptor are superimposed on signal and distort it. The

characteristic feature of these forms of interferences is the fact

that they are independent of signal and occur even when signal on

output of receiver is absent. On the basis of this property the

outside interferences and internally-produced noise were called the

name additive interferences. According to its properties the mostd

widely use additive interferences can be broken into three basic
groups: fluctuating, concentrated (sinusoidal), and pulse.
    Fluctuating interference in the general case is chaotic,

irregular change in time of the stress or current in any electrical

circuit. In telephone communication the fluctuating interference is

in the form of the characteristic noise, audible in telephones.

Therefore it frequently they call also by noise interference or

simply by noise.

For the illustration of noise interference Fig. 2.1.1 shows the

form of fluctuating stress ("noise path/track") from the output/yield

of the receptor, observed in the screen of electron oscillograph.

Figure 2.1.2a shows not distorted by the integrating effect of screen

the structure of te voltage of the same source, removed during one

period of scanning/sweep, while Fig. 2.1.2b, c gives the structures

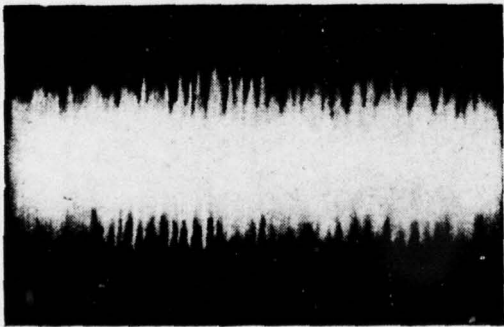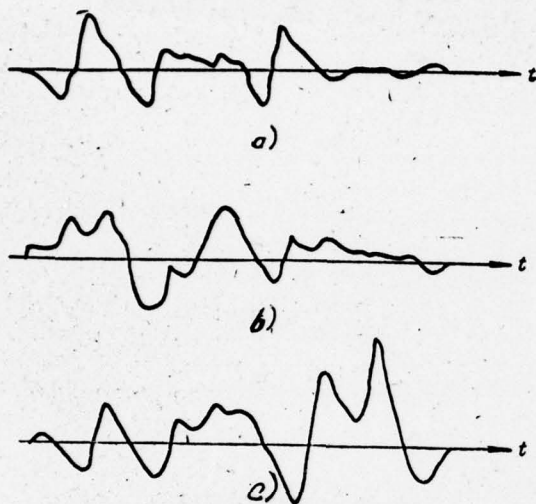of the stresses at other moments of time.

Fig. 2.1.1.

Fig. 2.1.2.

Page 43.

Internally-produced noise of receptor - a typical example
fluctuating interferences are many forms of atmospherics, cosmic
noises, etc. Fluctuating interference exists at the output/yield of
receptor continuously, and its spectrum virtually fills entire band
of frequencies of the receiver.

The concentrated interferences were called their name on the
strength of the fact that basic part of the power of such
interferences was concentrated in the separate relatively small
sections of frequency band, as a rule, smaller than the passband of
receiver Δf_{np}. Usually basic part of the power of the concentrated
interference is arrange/located in the frequency band, commensurable
with value 1/T, where T is a duration of the cell/element of signal
during the transmission of discrete report/communication.

page 44.

Such interferences have relatively prolonged in time character and

are sinusoidal high-frequency oscillaticns, modulated from one or
several parameters (amplitude, frequency, phase).

For an example Fig. 2.1.3 shows the temporary forms and the
spectra of the concentrated interferences in the form of not
modulated and modulated in amplitude HF oscillations.

The concentrated interferences are created by the signals of
extraneous radio stations, for example by the signals of
radiotelegraph, broadcast, television radio stations, and also by the
emission/radiations of the high-frequency oscillators of different
designation/purpose (industrial, medicinal) etc. A change of the
parameters of the concentrated interference in the place of reception
depends on the conditions of signal conditioning of the sources of
interferences, on the conditions of the propagation of these signals
and, as a rule, it has random character.

It should be noted that to the action of the concentrated
interferences are especially subjected the channels of communication
in the ranges of long, average and short waves. This fact is the
consequence of the conditions of radiowave propagation under the

indicated ranges, in which the emission/radiations of transmitters

create the noticeable strengths of field at considerable distances.

However, with the large number of concentrated interferences, which

is observed in these ranges sufficiently frequently, they store/add

up and form in the place of reception the interference, which differs

little from noise interference. At the same time in a number of cases

to the input of receptor can enter the separate concentrated

interferences, which sharply are isolated against

common/general/total noise background and they have power,

commensurable with the power of useful signal or which exceed it.

Under pulse interference it is accepted to understand this

regular either chaotic sequence of the mixing

momentum/impulse/pulses, by which for time of the duration of the

cell/element of signal $\mathcal{T}$ for the input of receiving of device enters

one or the small number of momentum/impulse/pulses [1].

FOOTNOTE [1]. If for time of the cell/element of signal for the input

of receiver enters the very large number of mixing

momentum/impulse/pulses, then the effect of this interference

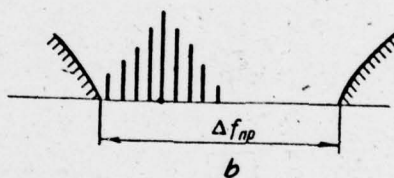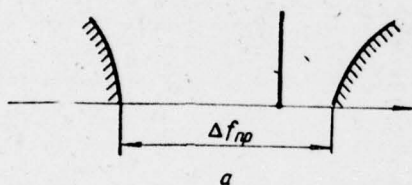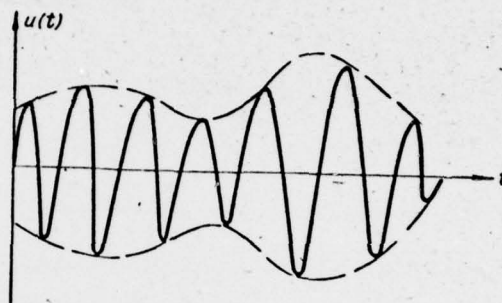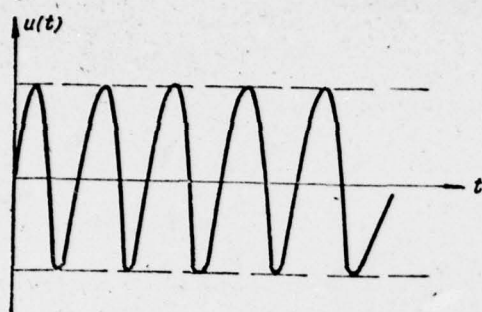virtually in no way differs from the fluctuating. ENDFOOTNOTE. Page

45.

Fig. 2.1.3.

Page 46.

The mixing momentum/impulse/pulse is any interference, which has the duration τ, considerably less than duration of the cell/element of signal (Fig. 2.1.4). The real mixing momentum/impulse/pulses have a duration of order $10^{-5}-10^{-8}$ s. Despite the fact that the pulse interference operates very short time, it can substantially lower the correctness of the transmitted information, since interference spectrum fills entire band of frequencies of the receiver and its energy can be very considerable.

The pulse interferences include many atmospherics (for example, lightning discharges) and interferences of the industrial origin: interference from the devices of the engine ignition of internal combustion, gas discharges, interferences from the electric power lines, etc.

Another very widespread factor of a decrease in the correctness of the transmitted information are signal distortions because of random changes in the state of the communication channel. Such distortions cause random modulation of signal on amplitude, frequency

or phase and develop themselves only during the passage of signal along the real link of communication/connection. They were called the name multiplicative interferences. Np
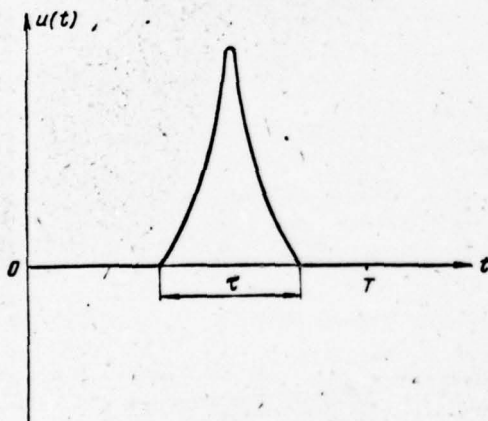
Fig. 2.1.4.

One of the most popular types of multiplicative interferences, that are inherent in the majority of radio channels, is the effect of signal fading at the input of receptor, in other words, the effect of the continuous and irregular fluctuations of signal level at the point of reception. Depending on the character of these fluctuations differentiate common/general/total (flat) and selective, rapid and slow fadings [24, 36]. The determinations and the general characteristic of the indicated types of fadings were given in §1.3. The physical cause for fadings most frequently is multiple-beam characteristics [13].

The examined types of additive and multiplicative interferences cause the random in time character of received signals, which previously cannot be predicted. Because of this descends the correctness of the transmitted information in any communicating system. At the same time so on use for the transmission of the report/communications broadband signal is a series of the specific special feature/peculiarities. For example, such communicating systems provide higher noise-resistance under the influence of the single concentrated noise, in channels with multiple-beam

characteristics, etc. To account for these peculiarities of broadband
signals and affecting them noise is necessary probabilistic, static
approach. Fundamental punty about quantitative noise characteristic
in this approach are given in the following paragraphs of the present
chapter.

§2.2. Radio jammings as random process. Concept of random processes
and their fundamental characteristics.

Fluctuating, concentrated, pulse interferences, and also signal
distortions in the communication channels are according to the
terminology of the probability theory of different kind random (or
stochastic) processes. Under random process it is accepted to
understand this function of time, which in the course of experiment
can accept one or another concrete/specific/actual form, but is
unknown previously, as which precisely [1].

FOOTNOTE [1]. In practice can be encountered the random processes,
which depend not on time, but from any other argument. for example, a
change in the temperature of air in the different layers of the
atmosphere depends randomly on the height/altitude of layer, etc.

ENDFOOTNOTE.

Let us explain this determination with the aid of following an example.

Page 48.

Let us assume that there is the large number N of completely identical receivers (the "ensemble" of the receivers), which work simultaneously in identical conditions. Let at the output/yield of each of them be observed the random process ε (t) (Fig. 2.2.1). The concrete/specific/actual form (oscillogram or photograph), which acquires random process at the output/yield of any of the receivers, i.e., in the result of one experiment, it is called the realization of random process.
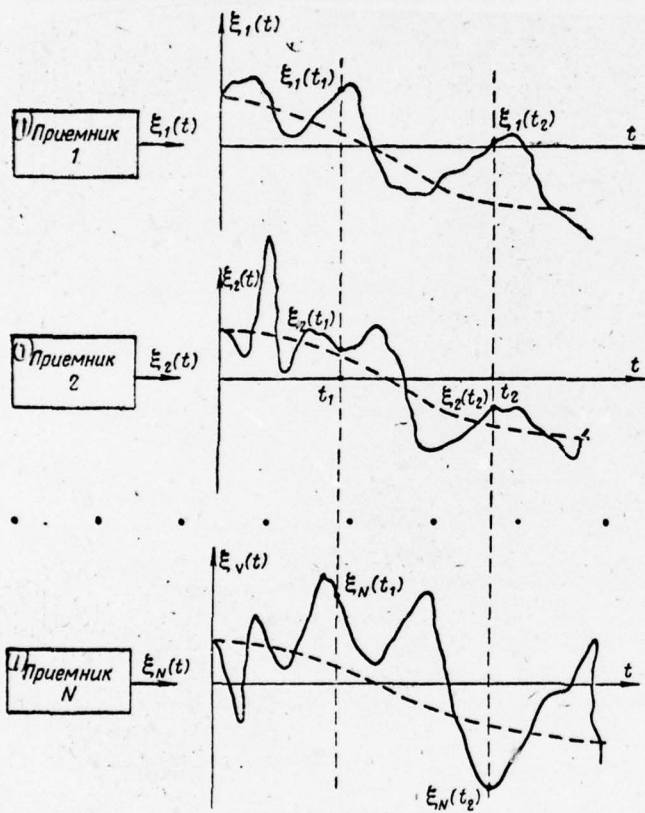
Fig. 2.2.1.
Key: (1) Receiver.

The totality of all realizations $\xi_1(t), \xi_2(t), \ldots, \xi_N(t)$ forms set or the

ensemble of the realizations of random process. Each realization $\xi_i(t)$,

i = 1, 2, ..., N is certain concrete/specific/actual function of

time. However, realizations at the output/yield of each of the

identical receivers will be different, i.e., the values of argument t

do not determine the unambiguously appropriate values of random

process. In this is an essential difference in the random process as

functions of time of the determined (regular) function.

It is natural that the random process is determined by entire

totality of the possible realizations. The number of latter in

ensemble is how conveniently large. In our example it was limited by

value N only for the clearness of presentation.

Let us fix now the moment of time, for example, $t = t_1$ (Fig.

2.2.1), and let us produce for this torque/moment the reading of the

values of random process at the output/yield of each of the

receivers. Then we obtain many random value $\xi_1(t_1)$, $\xi_2(t_1)$, ..., $\xi_N(t_1)$, which correspond to the realization of random process at this moment of time. Totality $\xi_1(t_1)$, $\xi_2(t_1)$, ..., $\xi_N(t_1)$ forms random variable, i.e., value, which as a result of experiment can accept certain value, whereupon is previously unknown, which precisely. This random variable is conventionally designated as the section of random process for the point in time $t = t_1$. If we fix another moment of time $t = t_2$ (Fig. 2.2.1), then we will obtain another random variable $\xi_1(t_2)$, $\xi_2(t_2)$, ..., $\xi_N(t_2)$ and, etc. In the general case to the different moments of time will correspond different random variables [1].

FOOTNOTE [1]. Since the number of realizations of process ξ (t) how is conveniently great, these random variables are continuous, i.e., each of them has a countless multitude in of the values, which continuously fill certain interval of the values. ENDFOOTNOTE.

In other words, any random process could be considered as random variable, which depends on time (or any other argument).

Thus, random process combines in itself and the feature of the random value and feature of the function: with the fixed/recorded

torque/moment in time it is converted into random variable, and as a result of one experiment - into certain concrete/specific/actual function of time.

Page 50.

Arises the question concerning how to quantitatively rate/estimate one or another random process. Answer/response to this question is not entirely evident. Actually, the simple observation of realization $\xi_i(t)$ at the output/yield of any of N of receivers (Fig. 2.2.1) during the determined interval of time still nothing says about which it will be it on the other interval of time, or it does not make it possible to determine the possible realizations at the output/yield of other identical receivers. In exactly the same manner the simple observation of a series of the values of random process at certain fixed/recorded moment of time still nothing says about other possible values in this section, or about the values of random process in sections for other moments of time. Answer/response to the placed question can be yes only when using the probability theory.

Random process ξ (t) can be described quantitatively, if we for

each cross section of process at any moment of time indicate not only

its possible values, but also describe these values from the point of

view of the probability of their appearance, or indicate the

probabilistic communication/connections between the values from the

different sections of process. In order to explain this confirmation,

let us return again to the examined above example.

Figure 2.2.2 depicts the ensemble of the realizations of random

process ε (t) at the output/yield of system from N of identical

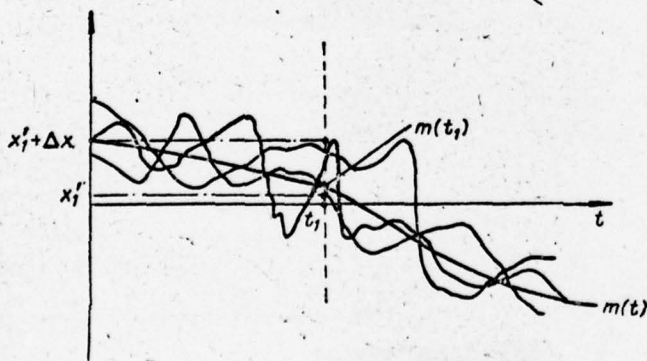receivers. Let us fix it at point in time $t = t_1$.

Fig. 2.2.2.

From the obtained value part $\xi_1(t_1), \xi_2(t_1), \ldots, \xi_N(t_1)$ let us isolate those n_1 of the values, which are included in very low (infinitesimal) range from certain value x_1 to $x_1' + \Delta x$.

Let us examine relation n_1/N. With the unlimited increase in the number of realizations N this relation determines probability that the random process $\xi(t)$ with $t = t_1$ (random variable $\xi(t)$) will render/show within the limits of interval $(x_1' \div x_1' + \Delta x)$. The indicated probability is proportional to the value of interval Δx and depends on the moment of time t_1 as on the parameter. Mathematically this is record/written as follows:

$$P\{x_1' < \xi(t_1) < x_1' + \Delta x\} = W(x_1'; t_1)\Delta x. \quad (2.2.1)$$

In this to formula the expression $W(x_1'; t_1)$ relates certain selected by us possible value of random variable $\xi(t_1)$ with the probability of its appearance. Let x_1 be any possible value of the random quantity $\xi(t_1)$. Analogous with

relationship/ratio (2.21) it is possible to determine the hit probability of random value into infinitesimal range in the vicinity of value x_1. then expression $W(x_1; t_1)$ will functionally relate any possible value x_1 continuous random value $\xi(t_1)$ with the probability of its appearance, i.e., it determines the law of the probability distribution of this random variable. Day lily $W(x_1; t_1)$ is called

one-dimensional probability density variable x_1 or one-dimensional probability density of random process [1].

FOOTNOTE [1]. In a word "one-dimensional" is emphasized that fact that the probability density $W(x_1; t_1)$ is determined for one any moment of time t_1. ENDFOOTNOTE.

Probability density $W(x_1; t_1)$ it is represented usually, as this will be shown somewhat later, either in the form analytical or in the form graphic dependence of it. One-dimensional probability density gives the total characteristic of the behavior of random process at any fixed/recorded moment of time. It serves as base for the probabilistic description of random process. However, this characteristic is not exhausting for entire process as a whole, since it does not furnish information on about that, are how tightly interconnected the values of process at the different moments of time, for example t_1 and $t_2 > t_1$, as they affect these values one for another and, etc.

Page 52.

In other words, one-dimensional density nothing speaks about the
dynamic loudspeaker of the development of process in time. These
information gives two-dimensional density. By discussing analogously,
it is possible to introduce into examination three-dimensional,
four-dimensional and the, etc of probability density. These
multidimensional densities form the peculiar staircase, raising on
which researcher it obtains consecutively increasing volume of
information about behavior of random process. One should again
emphasize that the probability densities give the comprehensive
quantitative characteristic of random process.

In practice not always appears the need for the use of
multidimensional probability densities. In a number of cases for the
description of random process sufficiently it is to be restricted to
the one-dimensional of probability. Furthermore, in a series of
practical problems, for example in the analysis of the
transformations of the spectra of signals and interferences in any
radio engineering circuit, the need for characterizing the process
the density function of probability, but sufficiently is to indicate
some numerical ratios, to a certain degree the characteristic most
essential features of densities. These quantitative

relationship/ratios are called numerical characteristics of random process. From numerical characteristics let us let us point out to those having very large applied value mathematical expectation (average value), dispersion and correlation function.

Mathematical expectation (average value) of the process

$$m(t_1) = \overline{\xi(t_1)} = \int\limits_{-\infty}^{\infty} x_1 \mathbb{W}(x_1;\ t_1)\,dx_1. \qquad (2.2.2)$$

Mathematical expectation is certain average value (average level), relative to which in this cross section of process change the values of random variable $\xi(t_1)$. In general, mathematical expectation depends on the timing t_1, i.e., it is certain nonrandom function of time.

Page 53.

Figure by 2.2.2 fine/thin lines shows the realizations of random process, and greasy/fatty isolated its mathematical expectation m (t).

Dispersion of the process

$$\sigma^2(t_1) = \overline{[\xi(t_1) - \overline{\xi(t_1)}]^2} =$$
$$= \int_{-\infty}^{\infty}(x_1 - \overline{x_1})^2 W(x_1; t_1)\,dx_1. \qquad (2.2.3)$$

Dispersion is the average value of the standard deviation of the random variable in the section of process at $t = t_1$ from its average value in this same section, i.e., it characterizes the degree of

scattering the separate values of random variable $\xi(t_1)$ relative to its average value. The dispersion of process also is the nonaccidental function of time.

Correlation function

$$R(t_1;\ t_2) = \overline{\left[\xi(t_1) - \overline{\xi(t_1)}\right]\ \left[\xi(t_2) - \overline{\xi(t_2)}\right]}. \qquad (2.2.4)$$

The correlation function characterizes on the average the degree of the communication/connection between the values of random process in sections for certain torque/moment of time t_1 and of another torque/moment in time t_2. It also is the nonaccidental function, which depends in the general case from two parameters t_1 and t_2. In detail correlation function and its properties will be examined in §2.3.

Let us note that the mathematical expectation, the dispersion and correlation function, determined by relationship/ratios (2.2.2) - (2.2.4), are numerical chaaracteristics of mean for the ensemble of

realizations (or simply ensemble average), since they are obtained by means of averaging for the assigned moments of time on entire multitude of realizations of random process ξ (t). The feature above symbols in the indicated formulas indicates the sign of averaging on many realizations.

Random processes is accepted to divide into transient and stationary. Transient random process is characteristic fact that its probabilistic characteristics depend on time, i.e., transient process has the determined tendency of development in time. the presence of stability indicates the fact that the random process occur/flow/lasts in time statistically uniform. Regarding the stationary process, at whose all the probability densities do not depend on the zero time reference.

Page 54.

Consequently, for such processes one-dimensional probability density does not depend upon where on the axis of time is selected the moment of time t_1, i.e., $W(x_1; t_1) = W(x)$. The numerical characteristics of stationary processes also do not depend on the zero time reference. Specifically, the mathematical expectation and the dispersion of this process are in this case some constant values m and σ^2. In turn, correlation function depends only on interval $\tau = t_2 - t_1$, but not on values themselves t_2, t_1, i.e., $R(t_2; t_1) = R(\tau)$.

Stationary random processes frequently are encountered in practice [1].

FOOTNOTE [1]. Sometimes for random processes they are limited only to requirement for independence from the zero time reference of the average value, dispersion and correlation function. Such processes are called stationary and wide sense [22]. ENDFOOTNOTE.

As their example can serve fluctuating noises at the output/yield of receptor, the noises of resistor/resistances, of electron tubes, etc.

The concept of stability largely facilitates the investigation of the random processes, which describe the properties, for example, of telegraph (including broadband) signals in the communication channels with the fluctuating and other interferences.

as we saw, the densities of probability and the numerical

characteristics of such processes are described simpler than in the
case of transient processes. Besides the fact that it is not less
importantly, for stationary processes is considerably simpler than
the experimental determination of different numerical
characteristics. Actually, the determination of mathematical
expectation, dispersion and correlation function from formulas
(2.2.2), (2.2.3) and (2.2.4) requires knowledge sufficient larger
amount of realizations of random process are related with cumbersome
calculations.

The presence of stability as the statistical uniformity of the
behavior of process in time runs against thought about the fact that,
apparently, for this process the only one realization for a
sufficiently long interval of observation possesses the properties,
characteristic for entire process as a whole, and it allows it will
determine a series of its characteristics.

Page 55.

Having only one realization of process, possible it would be
sufficient simply to find its mathematical expectation, utilizing the

formula

$$m = \lim_{T \to \infty} \frac{1}{T} \int_0^T \xi_1(t)\, dt, \qquad (2.2.5)$$

where T - the duration of the interval of observation; $\xi_1(t)$ - certain realization of process in this interval.

Actually this expression characterizes the constant component of process. So, if $\xi(t)$ there is certain stress of complex form, then m - the constant component of this stress (Fig. 2.2.3).
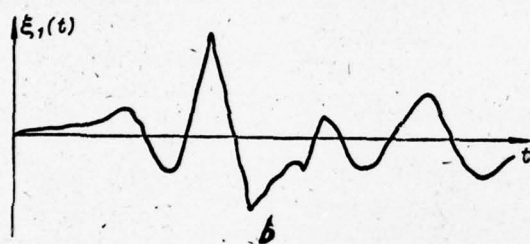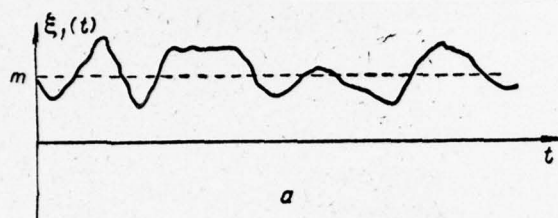
Fig. 2.2.3.

~~Page~~ Page 56.

The dispersion is determined by the relationship/ratio

$$\sigma^2 = \lim_{T \to \infty} \frac{1}{T} \int_0^T [\xi_1(t) - m]^2 \, dt \qquad (2.2.6)$$

and characterizes the power of the variable component (fluctuations) of process [1].

FOOTNOTE [1]. It is necessary, however, to keep in mind that, although the dispersion characterizes the power of the fluctuations of process, these values they have different dimensionalities.
Electrical power is expressed in watts, and dispersion - in volts or amperes squared. Dispersion is numerically equal to power, if random process, which is stress or current, operates during resistor/resistance into 1 ohm. ENDFOOTNOTE.

In turn, the correlation function

$$R(\tau) = \lim_{T \to \infty} \frac{1}{T} \int_0^T [\xi_1(t) - m][\xi_1(t + \tau) - m] dt. \quad (2.2.7)$$

Values m, σ^2, R (τ), found with the aid of relations (2.2.5) -
(2.2.7) by means of averaging in sufficiently large time interval,
are called time average. Question is comprised only in that,
correspond to ensemble average values, strictly defined by formulas
(2.2.2) - (2.2.4). It proves to be that among stationary processes
are many processes, which possess the remarkable property of
ergodicity. The essence of this property is comprised in the fact
that for such processes any probabilistic characteristic, obtained by
means of averaging on the ensemble of realizations with probability,
how is convenient to close to unity, is equal to the analogous
characteristic, obtained from single unique realization of process by
means of averaging for sufficiently large time interval.

Thus, for ergodic processes time average m, σ^2 and R (τ) are equal to analogous characteristics, obtaned by averaging p to the ensemble of realization. This property very widely is utilized in theory and practice of radiolink systems, where the main role play the stationary processes, which possess the property of ergodicity. Therefore further will be examined the only such processes.

§2.3. Correlation function of stationary process.

The values of random process at the different moments of time are mutuallyed-depend. This means that its some values cannot arbitrarily pass to others. In other words, the preceding/previous values of process to a certain degree predetermine the nearest subsequent values.

Page 57.

Physically   the interdependence of the values of process at the different moments of time can be explained by the fact that each random process operates in real system, or the electrical circuit,

which possesses inertial properties. The presence of inertness limits the rates of change in the process. These communication/connections between the values of random process at the different moments of time bear probabilistic character. For their quantitative estimate/evaluation on the average is utilized the numerical characteristic of process, called correlation function.

In the present paragraph fundamental the attention will be allotted to the correlation function of the stationary process ξ (t), which possesses ergodic property, i.e., to such functions, with which we encounter during the study of the properties of broadband signals in noisy channels.

The interference of the various kinds of interferences is determined by the power of the variable component (by fluctuation) of process. The constant component (average value) of process usually nothing adds to the interference of interference and in concrete/specific/actual diagrams always can be easily filtered out. Therefore subsequently, if this is not specified especially, the mathematical expectation of process ξ (t) is eliminated from the examination by the assumption that m = 0. Then correlation function (2.2.4) assumes the form

$$R(\tau) = \overline{\xi(t)\,\xi(t+\tau)}. \qquad (2.3.1)$$

Correlation function (2.3.1) possesses the following properties.

1. It is the even function of argument τ, i.e.,

$$R(\tau) = R(-\tau). \qquad (2.3.2)$$

This property is the consequence of the stability of process ξ (t).
Actually, for a stationary process correlation function does not
depend on the selection of the zero time reference t, ~~but it depends
on the selection of the zero time reference t,~~ but it depends only on
interval τ between countdowns. Therefore the values of correlation
function for the moments of time t, t + τ and t, t - τ (see Fig.
2.3.1) take the form

$$R(\tau) = \overline{\xi(t)\,\xi(t+\tau)} = \overline{\xi(t)\,\xi(t-\tau)} = R(-\tau).$$

Page 58.

2. With $\tau = 0$ value of correlation function is equal to the dispersion of process, i.e.,

$$R(0) = \overline{\xi^2(t)} = \sigma^2. \qquad (2.3.3)$$

3. The absolute value of correlation function at any values τ cannot exceed its value with $\tau = 0$, i.e., [1]

$$|R(\tau)| \leqslant \sigma^2. \qquad (2.3.4)$$

FOOTNOTE [1]. The proof of this confirmation the reader can find, for example, in [8.22]. ENDFOOTNOTE.

4. For the stationary process, which possesses ergodic property, is correct the relationship/ratio

$$\lim_{\tau \to \infty} R(\tau) = 0. \qquad (2.3.5)$$

This result means that with an increase in the interval τ among two sections of process the dependence among the values of process in these cross sections determined by relationship/ratio (2.3.1), ever more weakens and it vanishes. For the values of range τ from 0 to sufficiently large τ the character of a change in the correlation

function can be different. With this R (τ) it can take both the positive and negative values, included in absolute value between σ^2 and zero.

Figure 2.3.2 gives the curve/graphs of two correlation functions of random processes, very being frequently encountered in practice.
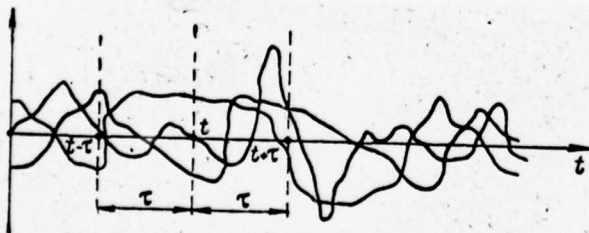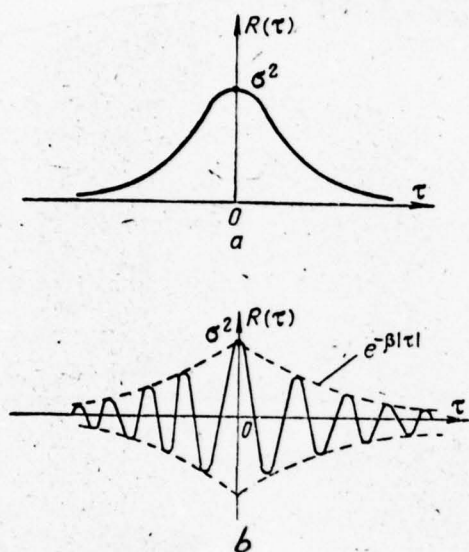

Fig. 2.3.1.

Fig. 2.3.2.

Page 59.

The first of these functions characterizes the correlation properties of fluctuating noises at the output/yield of the idealized oscillatory duct with the resonance curve, accurately repeating the curve of Fig. 2.3.2a. The second correlation function is inherent in noises at the output/yield of real parallel oscillatory duct. Both correlation functions satisfy the given above conditions. However, the character of their change with an increase τ is various.

The correlation function of Fig. 2.3.2a is described by the formula

$$R(\tau) = \sigma^2 e^{-\alpha\tau}, \qquad (2.3.6)$$

where α - certain constant value. as can be seen from figure, with an increase τ it monotonically vanishes. This character of behavior R (τ) is inherent in the random processes, which are relatively slowly changed in time.

Page 60.

The correlation function of Fig. 2.3.2b is determined by the relationship/ratio

$$R(\tau) = \sigma^2 e^{-\beta |\tau|} \cos \omega_0 \tau, \qquad (2.3.7)$$

where β is certain constant; ω_0 - change frequency correlation. From the figure one can see that it with an increase τ vanishes not monotonically, but consecutively reversing the sign. The negative values $R(\tau)$ attesting to the fact that in the appropriate interval τ most probable the different in sign values random process. This character of the behavior of correlation function is inherent in the rapidly changing in time processes. The faster this change, the more ω_0.

For the majority of random processes, which describe the distortions of signal and radio interference, virtually always it is possible to indicate this finite time interval τ_K, with excess of which it is possible to count $R(\tau) = 0$. This interval is called time of correlation, and the value of random process with $\tau > \tau_K$ they are called those which were not correlated.

By time of correlation is understood this time interval τ_K, after which correlation function it attenuates to certain negligible value, for example, to 1o/o or of several percentages from maximum. The value of time of correlation gives the representation of that, at which on the average values τ it is possible to set/assume not correlated the values of random process.

In practice very frequently are encountered the processes, the probability densities of which are described by the normal law of distribution (see §2.6). For these processes the lack of correlation of the values determines simultaneously and their static (probabilistic) independence.

Since the process $\xi(t)$ possesses ergodic property, its correlation function $R(\tau)$ can be determined by one realization, by utilizing the relationship/ratio

$$R(\tau) = \lim_{T \to \infty} \frac{1}{T} \int_0^T \xi(t)\xi(t+\tau)\, dt. \qquad (2.3.8)$$

Formulae (2.3.1) and (2.3.8) represent correlation function for the value of just one process at the different moments of time.

Page 61.

Emphasizing this fact, in such cases they say that R (τ) is aelf-correlation function. In the general case correlation function can characterize the degree of the communication/connection of two different random processes (for example, the stress of the sum of signal and interferences with the stress of heterodyne in the schematic of the converter of receiver etc.). In this case they speak about mutually correlated function (crosscorrelation function).

Let ξ (t) and $\eta(t)$ be two stationary process with zero $R_{\xi,\eta}$ (τ), mathematical expectations. Then their mutually correlated function as time average, it is determined by the relationship/ratio

$$R_{\xi,\eta}(\tau) = \lim_{T \to \infty} \frac{1}{T} \int_0^T \xi(t)\eta(t \pm \tau)\,dt. \qquad (2.3.9)$$

Unlike autocorrelation function, the crosscorrelation function is not even, i.e.,

$$R_{\xi,\eta}(\tau) \neq R_{\xi,\eta}(-\tau). \qquad (2.3.10)$$

If processes ξ (t) and $\eta(t)$ are independent (and the average value at least in one of them equal to zero), then $R_{\xi,\eta} = 0$ for all values τ.

The proof of these properties mutual-correlation function the reader will be able to find, for example, in works [21, 22]. Concepts auto- and the mutually correlated function as the parameters of signal find quite wide application in the transmission systems of

discrete information. They are fundamental for the understanding of the operating principles of broadband communicating systems.

Formulas (2.3.8) and (2.3.9) indicate the possibility of the simple experimental determination of correlation functions. in this case there is no need for that in order that T - time of the observation of the realization of process would be infinite. The error in the determination of correlation function because of finite time will be negligible, if T considerably exceeds time of correlation τ_K, i.e. if it is made condition [21]

$$T \gg \tau_k. \qquad (2.3.11)$$

Page 62.

Then it is possible to write the following formulas for correlation function:

$$R(\tau) = \frac{1}{T} \int_0^T \xi(t)\,\xi(t \pm \tau)\,dt;$$

$$R_{\xi,\eta}(\tau) = \frac{1}{T} \int_0^T \xi(t)\,\dot\eta(t \pm \tau)\,dt. \qquad\qquad (2.3.12)$$

Attempting to emphasize that fact that R (τ) and $R_{\xi,\eta}(\tau)$ in these expressions are determined in the finite interval of observation, they speak about short-term auto- and mutually correlated functions. The experimental computation of these functions is realized by the special devices, which obtained the name of correlators, or correlometers.

Figure 2.3.3a depicts the functional diagram of the simplest correlator, which works in accordance with (2.3.12). It includes the following fundamental elements: delay unit (delay line), multiplier,

the integrator and display.

Let us examine the operating principle of correlator during the computation of the short-term self-correlation function of process ξ (t). In this case the switch Π_1 is establish/installed in position "1". Let also the switch Π_2 be established/installed in certain i-th position that corresponds to certain fixed/recorded value τ_i (i = 0, 1, 2, ..., m - the number of positions of switch). Then the input of multiplier enter signals (realizations of process) ξ (t) and $\xi(t-\tau_i)$, displaced (delayed) one relative to another into the lines of delay for a period τ_i (Fig. 2.3.3b). In multiplier ξ (t) and $\xi(t-\tau_i)$ they are multiplied either with the aid of electron-tube (transistor) diagrams, or with the aid of the devices, which possess effect of multiplication, for example by the Hall effect [21].

Voltage $\xi(t) \cdot \xi(t - \tau_i)$ enters the integrator with time of integration T, at output/yield of which is formed the value of autocorrelation function $R(\tau_i)$ (Fig. 2.3.3c). In concrete/specific/actual diagrams the integration can be realized, for example, with the aid of chain/network RC, the low-pass filter or special electronic circuits. The result of integration enters the display unit, as which there can be either measuring meter or oscillograph, etc. Page 63.
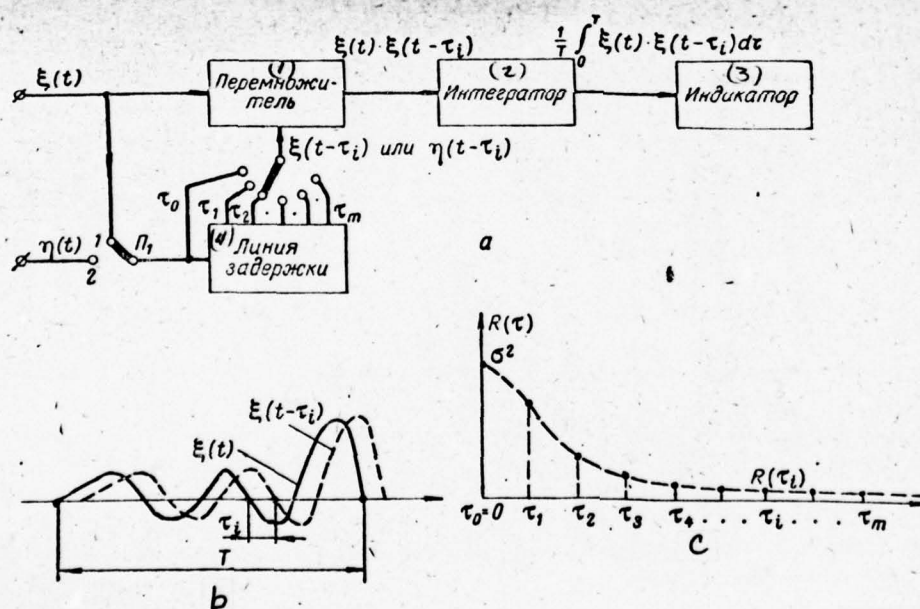
Fig. 2.3.3.

Fig. 2.3.3.

Key: (1). Multiplier. (2). Integrator. (3). Indicator. (4). Delay
line. Page 64.

By changing consecutively the positions of switch Π_2 from τ_0 to τ_m
and by giving each time of computation, it is possible to obtain the
sufficiently complete representation of the nature of the
autocorrelation function of process. Of course, with this the values
of delay time τ_i must be selected in accordance with the character
of a change in the correlation function: for the correlation function
of the form, presented in Fig. 2.3.2a, it is possible to take them
thinner than for the function of Fig. 2.3.2b.

During the determination of mutually correlated function (switch
Π_1 is established/installed in position "2") correlator works
analogously.

The given functional diagram for computation auto- or to
mutually correlated function is placed as the basis of the operating
principle of the correlators, utilized in broadband communicating

systems (see §3.1 and 4.1).

§2.4. Spectrum of stationary process.

Besides the probability density, and also of mathematical expectation, dispersion, correlation the functions and of other numerical characteristics, the properties of stationary random process to the certain degree can be described also by its frequency spectrum. As is known, the concept of the spectrum widely is utilized in radio engineering for the analysis of the determined (regular, nonaccidental) processes.

Usually oscillating process in any electrical circuit is represented in the form of the sum of the harmonic oscillations of different frequencies. Then the function, which characterizes the amplitude distribution of harmonics according to different frequencies, is called the spectrum of oscillating process. The spectrum determines the internal structure of process, it indicates that, the fluctuations of which frequencies predominate in the process being investigated.

Spectral representation can be applied also for the analysis of random processes. However, in this case is an essential difference from the spectral representation of the determined process. If we present any realization of just one process in the form of the sum of harmonic oscillations, then the amplitudes of harmonics will be the random variables, which depend on the selected realization.
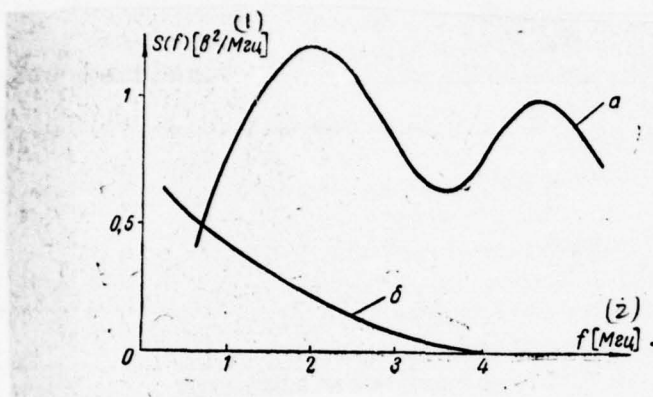
Page 65.

It is obvious that in this case the concept of the spectrum of amplitudes becomes meaningless, since it is unknown, which of the realizations one should give preference. But for a random process it is possible to introduce this concept of the spectrum, which considers the limitedness of the energy "resource/lifetimes" of process and it is connected with average power of its fluctuations (by dispersion of process σ^2). This concept is the spectrum of average power, or the energy spectrum. It characterizes the distribution of average power of the fluctuations of the harmonic components of process, in other words, distribution according to the different sections of the frequency range of average power

(dispersion) of process. This position is can illustrate Fig. 2.4.1., in which are shown the noise spectra, generated by thyratron (a) and by neon tube/lamp (b).

Quantitatively the spectrum of random process described by the spectral density of average power or by simple spectral density. The spectral density S (ω) is determined by relation containing in infinitesimal frequency interval (from ω to ω + dω) of average power of process dσ² to the width of this interval dω

$$S(\omega) = \frac{d\sigma^2}{d\omega}. \qquad (2.4.1)$$

Fig. 2.4.1.

Key: (1). [V²/MHz]. (2). [MHz].

Page 66.

The sense of this determination is illustrated by Fig. 2.4.2, in
which the shaded elementary section conditionally corresponds
infinitesimal value of average power $d\sigma^2$. Spectral density represents
the "height/altitude" of this section, i.e., power density for
certain frequency ω. As the function of frequency spectral density
characterizes power distribution according to the spectrum. It is
obvious that the total power of random process is equal to the sum of
the power, included in all elementary sections within the limits of

the occupied frequency band, which in the general case can stretch from 0 to ∞. Then the power of the process

$$\sigma^2 = \int_0^\infty S(\omega)\, d\omega. \qquad (2.4.2)$$

To evaluate the value of the frequency band, occupied by the random process, is introduced the concept of the width of energy spectrum, determined by relationship/ratio [22]

$$\Delta\omega = \frac{\int_0^\infty S(\omega)\, d\omega}{S(\omega_0)}, \qquad (2.4.3)$$

where $S(\omega_0)$ - the value of spectral density at certain characteristic frequency ω_0 (appropriate usually to the maximum of spectral density).
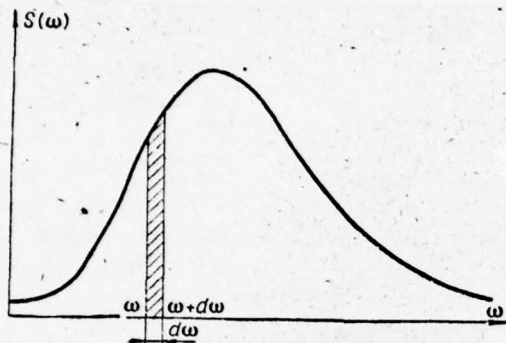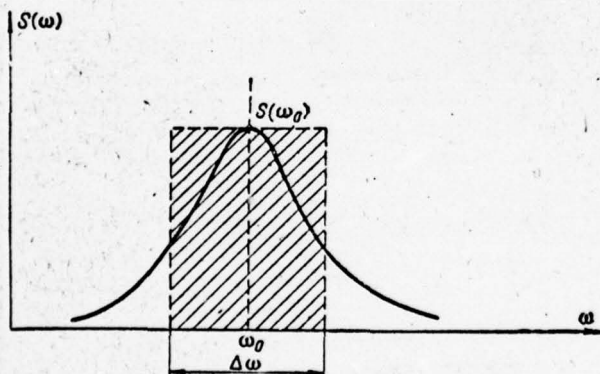
Fig. 2.4.2.

Fig. 2.4.3. Page 67.

As can be seen from formula (2.4.3), the width of the spectrum $\Delta\omega$ is equal to the foundation of rectangle with height/altitude S (ω_0), that has the same area as the area, included between the curve of S (ω) and the axis of abscissas (Fig. 2.4.3) [1].

FOOTNOTE [1]. Sometimes by the width of the spectrum $\Delta\omega$ is understood the interval between some extreme frequencies, at which the spectral density is reduced to the determined level, for example to half from the maximum value. The computation of the width of the spectrum at the level of half from the maximum value of spectral density and due to formula (2.4.3) brings in the majority of cases to virtually very close results. ENDFOOTNOTE.

In §2.1 it was noted that the radio interference possess different energy spectra. Their spectra depend, in the first place, on the character of the source of interferences and, in the second place, on that, through which circuits of selection passes the interference.

Sources can generate interferences with so-called discrete (lined) spectra, whose harmonic frequencies is multiple to certain fundamental frequency. Such spectra possess, for example, some concentrated interferences (see Fig. 2.1.3) or pulse interferences in the form of regular sequence of mixing momentum/impulse/pulses.

The sources of fluctuating interferences generate interferences with the continuous (continuous) spectrum, frequency components of which are not multiple one another (see Fig. 2.4.1). Such spectra I possess fluctuating noises of receptor, various kinds atmospherics, noises, created electronic tubes, and so forth [13, 18].

Page 68.

Depending on the character of spectral density as functions of frequency distinguish interferences with the nonuniform and uniform spectra. The nonuniform spectrum possess, for example, atmospherics, intensity which it decreases with an increase in the frequency, the single concentrated interferences, the fluctuating noises, generated by some electron-ion equipment/devices. Figure 2.4.1 depicts nonuniform noise spectra, generated by thyratron and neon tube/lamp.

Some sources generate interferences with the uniform spectrum, i.e.,
with constant in certain sufficiently broad band spectral density.
The uniform spectrum possess the noises, which appear in electron
tubes due to the presence of shot effect, the noises of
resistor/resistances, etc. For example, the spectral noise density of
resistor/resistances remains virtually constant up to frequencies
$10^{13}-10^{14}$ Hz.

In a number of cases with sufficient accuracy it is possible to
assume that the fluctuating interference, which operates on the input
of receptor, possesses the uniform spectrum in unlimitedly wide
frequency band (Fig. 2.4.4). The random process, which approximates
fluctuating interference with this spectrum, is called "white" noise
by analogy with the white world/light, which has uniform and
continuous spectrum in the visible part [1].

FOOTNOTE [1]. Unlike "white" noise fluctuating interference with
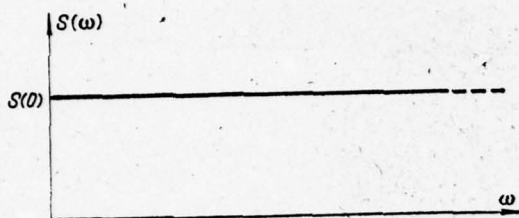nonuniform spectral density is called sometimes "painted" noise [39].
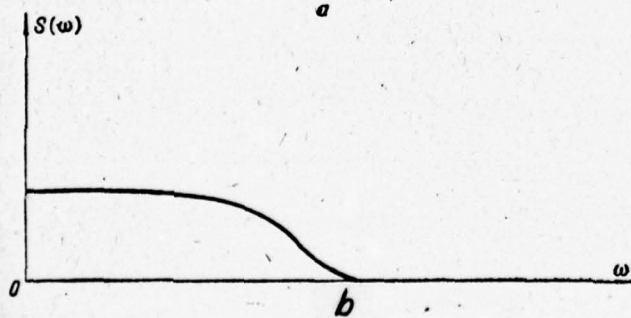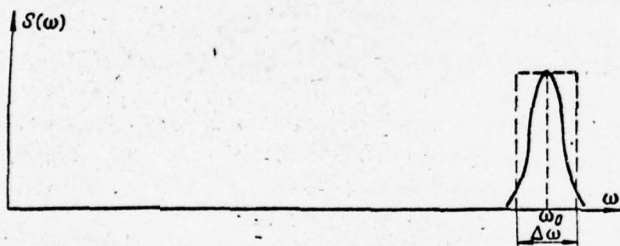ENDFOOTNOTE.

Fig. 2.4.4.

Fig. 2.4.5.

Page 69.

The concept of "white" noise is the very complete mathematical idealization, used when within the limits of the passband of receiver the spectral density of the affecting real noise can be considered approximately constant.

Further, as a result of the passage through the selecting electrical circuits the spectra of real interferences turn out to be to a certain degree those which were limited. In this case are distinguished narrow-band and broadband random processes [22]. Random process is called narrow-band, if its spectrum is concentrated in essence in the relatively narrow frequency band about certain frequency ω_0. Mathematically the condition of narrow-band characteristic is represented in the form

$$\Delta\omega \ll \omega_0, \qquad (2.4.4)$$

where $\Delta\omega$ - the width of the spectrum.

At symmetrical curve spectral density the frequency ω_0 is the medium frequency of the spectrum. The condition of narrow-band

characteristic means that the width of the spectrum of process must be much less than the medium frequency of the spectrum (Fig. 2.4.5a).

Page 70.

The random process is broadband, if condition (2.4.4) is not satisfied (Fig. 2.4.5b). Let us note that one ought not to mix the concept of the broad-band character of random process with the concept of the broad-band character of signal. When they tell about broadband signal in radiolink system, thereby emphasize that fact that the spectrum of this signal is much wider than the spectrum of the transmitted report/communication, i.e., the base of system FT >> 1. However, broadband signals just as all signals, used in radio communication are narrow-band in the sense of determination (2.4.4).

§2.5. Communication/connection between the correlation function and the spectrum of stationary of process.

From the viewpoint of the properties of stationary random process the spectrum characterizes the rate of change in its

instantaneous values. Figure 2.5.1 depicts to the realization of the
processes, in spectra of which are contained the different
components, the spectrum of the process of Fig. 2.5.1b containing
more high-frequency components, than the spectrum of the process of
Fig. 2.5.1a. As can be seen from figures, the instantaneous values of
process with the more high-frequency spectrum are characterized by
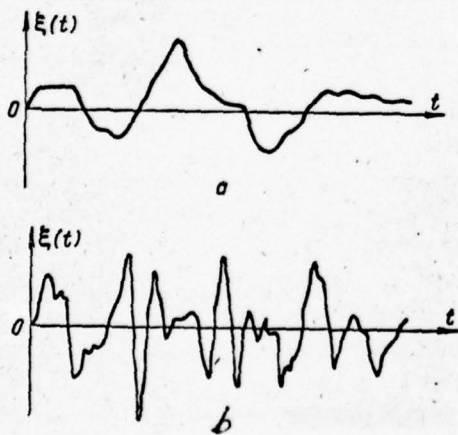the more frequent transitions through zero and by more rapid changes
in value. Np

Fig. 2.5.1.

~~Page 71.~~ Page 71.

On the other hand, the rate of change in the instantaneous values of process is determined by its correlation function. It is obvious that the correlation function and the spectrum are must be mutually-dependent. This interconnection is establish/installed by the following two integral relationships:

$$R(\tau) = \int\limits_{0}^{\infty} S(\omega) \cos \omega\tau d\omega; \qquad (2.5.1)$$

$$S(\omega) = \frac{2}{\pi} \int\limits_{0}^{\infty} R(\tau) \cos \omega\tau d\tau. \qquad (2.5.2)$$

Formulas (2.5.1) and (2.5.2) were simultaneously obtained by Soviet scholarly a. Ya. by Khinchin and American scholar N. Weiner and frequently they are called by khinchina - Weiner's relationships ¹.

FOOTNOTE [1]. Let us note that relationship/ratio (2.5.2) is more rigid than (2.4.1), by the determination of spectral density. ENDFOOTNOTE.

These relationship/ratios have great practical value. If is known the correlation function of process R (τ), then with the aid of formula (2.5.2) can be found its spectrum S (ω). If is known the spectrum of process S (ω), then, by substituting its value in (2.5.1), it is possible to determine the correlation function R (τ). Furthermore, khinchina - Weiner's relationship/ratios establish/install the interconnection between time of correlation τ_K and width of the spectrum $\Delta\omega$: the wider the spectrum of process, the lesser time of correlation, and vice versa.

Let us examine as an example the interdependence of the correlation function and spectrum for the limited on band fluctuating noise. Let of this noise spectrum be limited by frequencies ω_1 and ω_2 (Fig. 2.5.2a), i.e., it has a width $\Delta\omega = \omega_2 - \omega_1$. Let also the spectral noise density be constant. Then it is equal to $S_0 = \sigma^2/\Delta\omega$. Let us designate by $\omega_0 = \omega_2 + \omega_1/2$ the medium frequency of noise

spectrum and assume that it is considerably greater than the width of
the spectrum ($\Delta\omega << \omega_0$), i.e., the noise is narro-band. Then,
substituting in formula (2.5.1) expression for spectral density S_0,
and also taking into account that the integration limits for ω are
included between extreme frequencies ω_1 and ω_2, we will obtain as a
result of the integration

$$R(\tau) = \int_{\omega_1}^{\omega_2} \frac{\sigma^2}{\Delta\omega} \cos \omega\tau \cdot d\alpha = \sigma^2 \frac{\sin \frac{\Delta\omega\tau}{2}}{\frac{\Delta\omega\tau}{2}} \cos \omega_0\tau. \quad (2.5.3)$$

Page 72.

The curve/graph of this correlation function is represented in

Fig. 2.5.2b. From relationship/ratio (2.5.3) it is evident that the
correlation function takes the form of harmonic oscillation with
frequency ω_0, to the equal medium frequency of noise spectrum, and
with the changing according to the law $\frac{\sin\Delta\omega\tau/2}{\Delta\tau/2}$ amplitude. Factor $\frac{\sin\Delta\omega\tau/2}{\Delta\tau/2}$
is envelope correlation function which changes considerably slower
than "high-frequency filling" of $\cos\omega_0\tau$. The character of a change
in the envelope is determined by the form of the spectrum and depends
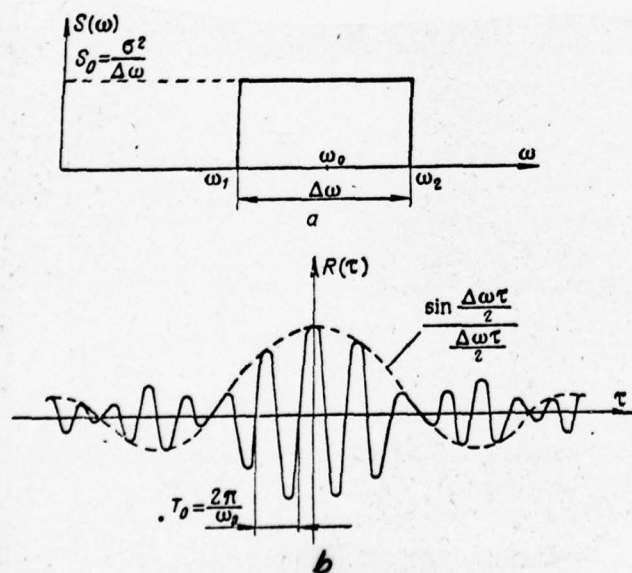on the width of the spectrum of $\Delta\omega$.

Fig. 2.52.

~~Sign Action~~ Page 73.

The product of time of correlation for the width of the spectrum composes for the case in question value

$$\tau_k \Delta\omega = \pi. \qquad (2.5.4)$$

It should be noted that the correlation function in the form of "high-frequency filling" cps $\omega_0\tau$ with the slowly being changed envelope is typical for the narrow-band processes, which have the symmetrical relative to certain medium frequency spectrum of arbitrary form. Difference from the correlation function of Fig. 2.5.2b will is only in the character of a change in the envelope, connected with the form of the spectrum.

Stationary processes with the continuous spectrum do not exhaust all processes of such type. Are possible the stationary processes which have discrete spectra. They include the processes, which describe a series of the radio interference (concentrated, pulse).

Discrete spectra possess the utilized for a transmission information the signals of form (1.1.1). The presence of discreteness in the spectra of such processes testifies to their periodicity. The concept of correlation function is added to such processes. However, unlike the examined earlier cases, when R (τ) vanishes with τ ⟶ ∞, the correlation function of periodic stationary process itself is the periodic function of argument τ. For this explanation let us examine some examples.

Let us examine the correlation function of the random process

$$\xi(t) = \sum_{k=k_1}^{k_2} A_k \cos(\omega_k t + \varphi_k) =$$
$$= \sum_{k=k_1}^{k_2} (a_k \cos \omega_k t + b_k \sin \omega_k t), \qquad (2.5.5)$$

where A_k and φ_k - some random the amplitude and the phase of the K-th harmonic component of process, connected with a_k and b_k the relationship/ratios

$$A_k = \sqrt{a_k^2 + b_k^2}, \ \varphi_k = -\arctan \frac{b_k}{a_k};$$

$$\omega_k = k\omega_0 = k\frac{2\pi}{T}$$

$(k_2 - k_1 + 1)$ - the number of the harmonic components of process; $\wedge$ is constant angular frequency of K-th component (t - period).

Page 74.

With $\overline{a_k} = \overline{b_k} = 0$, $\overline{a_k^2} = \overline{b_k^2} = \sigma_k^2$, $\overline{a_i b_j} = \overline{a_i a_j} = \overline{b_i b_j} = 0$ for any i and j, included between k_1 and k_2, the process ε (t) is stationary in the broad sense, but its correlation function is equal to [22]

$$R(\tau) = \sum_{k=k_1}^{k_2} \sigma_k^2 \cos \omega_k \tau. \qquad (2.5.6)$$

Hence it is apparent that R (τ) has a repetition period T.

The expansion of form (2.5.5) frequently is utilized in practice for the representation of stationary (in the broad sense) processes. When the dispersions of harmonic components are identical ($\sigma_\kappa^2 = \sigma_0^2$), correlation function (2.5.6) assumes the form

$$R(\tau) = \sigma_0^2 \sum_{k=k_1}^{k_2} \cos k\omega_0\tau = \sigma^2 \frac{\sin \frac{\Delta\omega\tau}{2}}{FT \sin \frac{\omega_0\tau}{2}} \cos\omega_{0\mathrm{p}}\tau, \quad (2.5.7)$$

where $\Delta\omega = (k_2 - k_1 + 1) \omega_0 = FT \omega_0$ - the width of the band of process ξ (t); $\omega_{0\mathrm{p}} = \frac{k_2 + k_1}{2} \omega_0$ - the medium frequency of the spectrum of process; $\sigma^2 = FT\sigma_0^2$ - the dispersion of process. From (2.5.7) it follows that in the case of the uniform spectrum of process its correlation function is high-frequency oscillation with frequency $\omega_{c\rho}$, to the equal medium frequency of the spectrum, and by envelope, being slowly changed according to the law $\sigma^2 \frac{\sin \Delta\omega\tau/2}{FT \sin \omega_0\tau/2}$.

The exemplary/approximate form of correlation function (2.5.7) for FT >> 1 is shown in Fig. 2.5.3. Time of correlation in the case in question let us estimate approximately in terms of that value τ, at which diffraction correlation function for the first time it becomes zero. We have

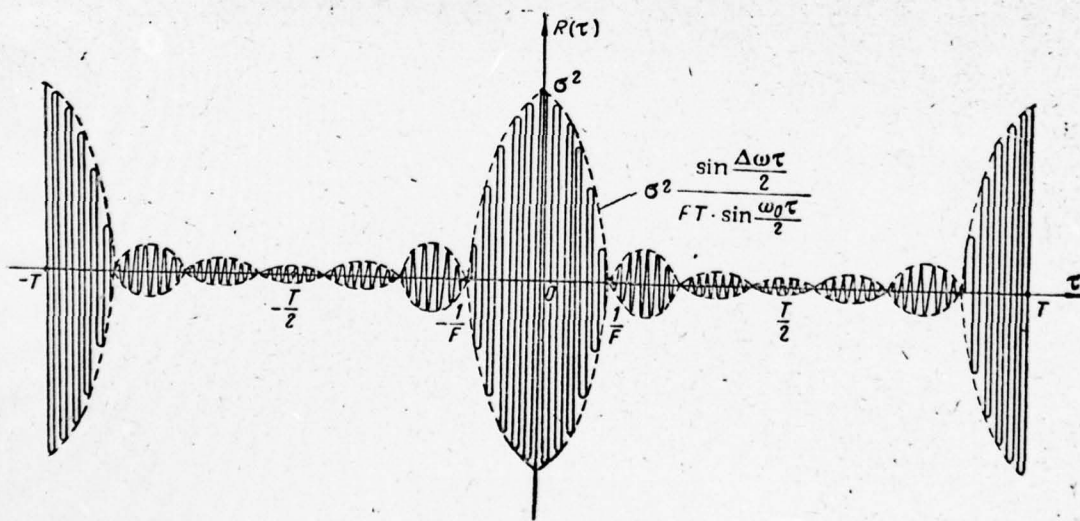$$\sigma^2 \frac{\sin \frac{\Delta\omega\tau_K}{2}}{FT \sin \frac{\omega_0\tau_K}{2}} = 0. \qquad (2.5.8)$$

Fig. 2.5.3.

Fig. 2.5.3. Page 76.

With FT >> 1 denominator of this expression changes considerably slower than the numerator, i.e., the position of the first zero in (2.5.8) is determined zero numerator. Then to relationship/ratio (2.5.8) is equivalent the equality

$$\sin \frac{\Delta \omega \tau_\kappa}{2} = 0$$

or

$$\frac{\Delta \omega \tau_\kappa}{2} = \pm \pi.$$

Hence it follows that time of the correlation

$$\tau_{\kappa} = \pm \frac{2\pi}{\Delta\omega} = \pm \frac{1}{F}. \qquad (2.5.9)$$

Consequently, time of correlation is inversely proportional to the width of the spectrum of process.

From Fig. 2.5.3. it is evident that the envelope of correlation function, besides the fundamental maximum with $\tau = 0$, has the supplementary maximums whose positions with FT >> 1 are determined by the values τ, which satisfy the condition

$$\frac{FT}{2}\omega_0\tau = \left(k + \frac{1}{2}\right)\pi; \; k = 1, 2, 3...$$

greatest of the supplementary maximums, corresponding k = 1,
i.e., τ = 3/2F, composes value

$$\left| \frac{\sigma^2}{FT \sin\left(\frac{3\omega_0}{2F}\right)} \right| \approx \frac{\sigma^2}{FT \frac{3\omega_0}{2F}} = 0.2\sigma^2,$$

considerably smaller than the fundamental maximum. Remaining
supplementary maximums have another value.

Let us note that the described character of a change in the
correlation function is retained in general terms, also, for a
periodic narrow-band stationary process with the nonuniform spectrum.
However, in this case the nonuniformity of the spectrum manifests
itself an increase in the time of correlation [6].

The concept of correlation function can be used not only to
stationary periodic processes, but also to utilize with respect to
determined periodic processes [22].

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Pages 77-122

§2.6. Normal random process. Envelope and the phase of narrow-band normal process.

The concept of normal random process plays the important role in practice. This is caused by that fact that the large number of sources creates interferences with the so-called normal law of probability density. As widely confirmed by experiment, this law, in

particular, it satisfies the probability distribution of the overwhelming majority of fluctuating interferences (noises of electron tubes, resistor/resistances and, etc). Specifically, in connection with these interferences most completely developed at present the basic condition/positions of the statistical theory of communication/connection. The one-dimensional probability density of stationary normal random process is determined analytically by following expression [8]:

$$W(x) = \frac{1}{\sqrt{2\pi}\sigma} e^{-\frac{(x-m)^2}{2\sigma^2}}, \qquad (2.6.1)$$

where m and σ^2 are a mathematical expectation and the dispersion of random process.

Graphically this dependence is represented in Fig. 2.6.1a. As can be seen from figure, probability density (2.6.1) takes the bell-shaped form, symmetrical relative to the mathematical expectation m. The maximum value a probability density has with x = m, and with an increase in the divergence to one side or the other it decreases, strive for zero when x ⟶ ∞. The latter means that

theoretically, although with small probability, are possible any in value overshoots in realization normal random process. The value of the mathematical expectation m characterizes the position of distribution curve on the axis of abscissas. During a change in value m the density curve of probability is displaced along the axis of abscissas, without changing form (Fig. 2.6.1b). The dispersion σ^2 characterizes not position, but only the very form of the density curve of probability. The more value σ^2, that more washed away it becomes density curve probability. (Fig. 2.6.1c) [1].

FOOTNOTE [1]. In the probability theory it is proven, that the area, included between the density curve of probability and by the axis of abscissas, does not depend on the law of distribution and is always equal to unity. ENDFOOTNOTE.

Page 78.

By returning to datum to §2.2 The determination of one-dimensional probability density (2.2.1), it is possible to explain its sense in the case in question as follows. Let us turn to Fig. 2.6.2, in which are combined the curve/graphs of the realization

of normal random process with zero mathematical expectation $(m = 0)$

and its one-dimensional probability density. Coincidence is conducted

so that the values of random process in both cases would be
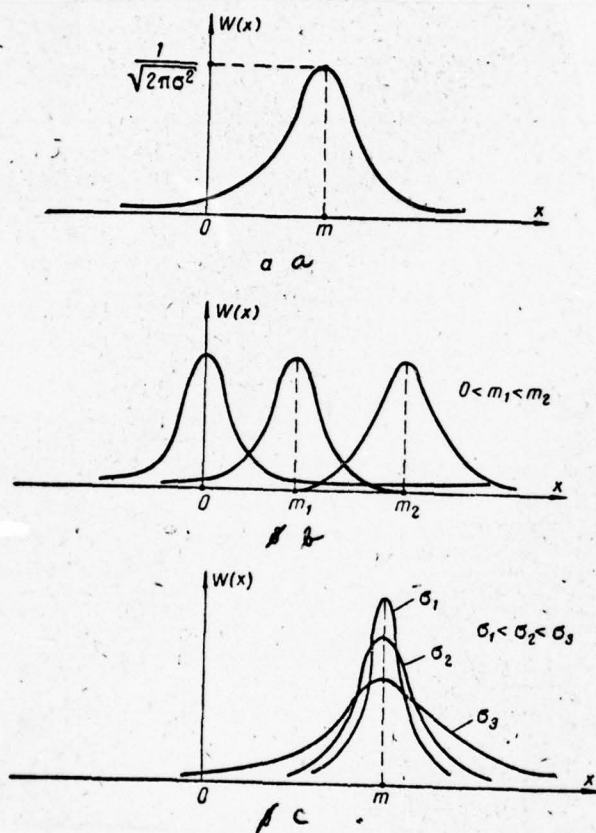
plot/deposited along the axes of ordinates.

Fig. 2.6.1.

~~Fig. 2.6.1~~. Page 79.

Let us isolate certain possible value of random process x_1 and infinitesimal interval Δx relative to it. Then the probability density W (x_1) characterizes probability of the appearances of this value of random process (shaded region for values from x_1 to x_1 + Δx).

If we isolate certain another value of random process x_2 (more than x_1) with infinitesimal interval Δx in its vicinity, then the probability density W (x_2) determines probability of the appearances already of this value. Since probability density with $x = x_1$ larger than with $x = x_2$, also the probability of the appearance of the first value of random process is more than the second. Thus, probability density relates any possible value of random process with the probability of its appearance. The greater the probability density for certain/some value, the more frequent the random process it will to take this value. The examined interdependence between the probability density and the random process turns out to be valid not only with normal, but also under any other law for a density of probability.

Normal stationary processes differ one another in terms of the form of correlation function (energy spectrum). As it follows from expression (2.6.1), the one-dimensional probability density of this process does not depend on the form of its correlation function (spectrum). Because of this normal processes with different correlation functions (spectra), but with identical as dispersions σ^2 will have identical densities probability.
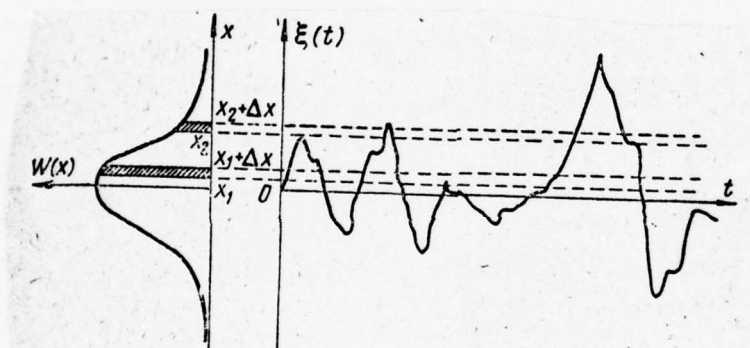
Fig. 2.6.2.

Fig. 2.6.2. Page 80.

The indicated difference between normal stationary processes develops itself in examination two-dimensional and the higher probability

densities of each of them.

Normal random processes possess a series of remarkable
properties. Specifically, for such processes from stability in the
broad sense follows their strict stability. Furthermore, if two
random processes are not correlated and normal, then they are
independent.

In the probability theory also it is proven (for example, see
[8]), that any linear combination of normal random variables has also
the normal law of density probability. Because of this during the
passage of the normal random process through any linear electrical
circuit (filter, integrator etc.) the law of its distribution does
not change, i.e., it remains normal. During passage it through
nonlinear electrical circuit (detector, limiter etc.) the law of
distribution changes and can differ significantly from the normal.

The detailed information about the conversion of normal random
processes by linear and nonlinear electrical cell/elements the reader
will find in work [22] and others. Given information relative to
normal process and its one-dimensional density probability is

sufficient for the understanding of the most important special
feature/peculiarities of the broadband systems, subjected to the
effect of fluctuating interferences.

In practice frequently are encountered the cases, when as a
result of the passage through these or other linear selecting
circuits stationary random process is converted into narrow-band with
the symmetrical relative to certain medium frequency spectrum. This,
as a rule, it occurs in the circuits of the high and intermediate
frequencies of radio receivers and other equipment/devices.

If we turn to the time diagram of this process (Fig. 2.6.3),
then it is evident that it is the "fluctuations" of comparatively
high frequency ω_{cp} with the slowly changing according to certain
random law amplitude. Outwardly narrow-band process is very similar
to the sinusoidal modulated in amplitude oscillation.

Page 81.

However, unlike the amplitude-modulated oscillation/vibration here

the amplitude of oscillations is modulated according to random law.
Furthermore, occurs slow change in the random law of the phase of
oscillations, which leads to a change in their frequency [1].

FOOTNOTE [1]. The rate of change in the envelope and phase is
determined by value $T = 2\pi/\Delta\omega$, where $\Delta\omega$ is width of the spectrum of
process. If the process is narrow-band, i.e., is fulfilled condition
(2.4.4), then envelope and phase virtually do not change during
period $T_{cp} = \dfrac{2\pi}{\omega_{cp}} \ll T$.   ENDFOOTNOTE.

This character of a change in the narrow-band stationary process
makes it possible to present it analytically in the form of the
harmonic signal, randomly modulated on amplitude and the phase:

$$\xi(t) = V(t)\cos[\omega_{op}t + \varphi(t)], \qquad (2.6.2)$$

where $V(t)$ and $\varphi(t)$ - being slowly changed in comparison with $\cos\omega_{cp}t$
random amplitude and phase;

ω_{cp} - the medium frequency of the spectrum of process.

Function of V (t) is called the envelope of narrow-band process
(see Fig. 2.6.3), a ϕ) t) - by the phase of process. The concepts of
envelope and phase of narrow-band process highly useful. In a number
of cases they facilitate the understanding of the physical essence of
the phenomena, which occur under the influence on the receptor of
interferences, and also they facilitate mathematical analysis. For
example, during detection of the various kinds of the interferences
often sufficiently it is to be restricted to the analysis only of
envelope of the process, which is isolated at the output/yield of
detector, and its remaining parameters can be disregarded.
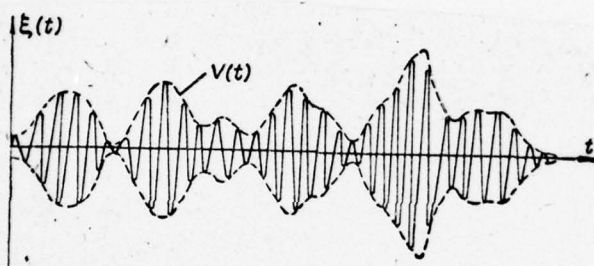
Fig. 2.6.3.

~~Fig. 2.6.3.~~ Page 82.

Let us note that, although the spectrum of process ξ (t) is
found in vicinity relative to the high frequency ω_{cp}, the spectrum of
the envelope V (t), defined through (2.6.2), it lie/rests at the
range of the low frequencies, adjacent at the beginning of
coordinates, analogous with that, as it takes place for the spectrum

of the envelope of the amplitude-modulated oscillations.

Further, if we in expression (2.6.2) indicate probability laws, to which satisfy the envelope V (t) and phase ϕ (t), then thereby process ϵ (t) will be completely determined. Of course, in this case examine/considered separately V (t) and ϕ (t) possess each their inherent laws of distribution of probabilities. Specifically, if ϵ (t) is normal stationary random process with zero mathematical expectation and dispersion σ^2, then the one-dimensional probability density of envelope satisfies the so-called Rayleigh law [22]:

$$W(V) = \begin{cases} \frac{V}{\sigma^2} \, e^{-\frac{V^2}{2\sigma^2}} \; ^{(I)}\text{при } V > 0; \\ 0 \qquad ^{(I)}\text{при } V \leqslant 0, \end{cases} \qquad (2.6.3)$$

$^{(I)}$ with

a the probability density of phase it satisfies uniform law in the interval of values from 0 to 2π, i.e.,

$$W(\varphi) = \begin{cases} \frac{1}{2\pi} & \text{(1) для } 0 \leqslant \varphi \leqslant 2\pi; \\ 0 & \text{(2) в остальных случаях.} \end{cases} \qquad (2.6.4)$$

(1) for

(2) in the remaining cases.

The graph/diagrams of dependences (2.6.3) and (2.6.4) are represented in Fig. 2.6.4.

As can be seen from figure, the probability density of envelope vanishes with V, that vanishes and for infinity. The character of the density curve of probability depends on value σ^2. Than is more σ^2, those "sealed" curve. The maximum of probability density is equal to

$$\frac{1}{\sqrt{e\,\sigma^2}},$$

where to e, i.e., Naperian base (e = 2.7182). It is obtained always with V = σ, i.e., among the different possible values of envelope (amplitudes) most probable the values equal to to value σ.

Page 83.

Other values are less probable and the possibility of their appearance is characterized by the appropriate values of probability density. As concerns the probability density of phase (Fig. 2.6.4b), during the observation of random process any possible values of it are equally probable in range from 0 to 2π.

§2.7. Concept of statistical testing hypotheses and the optimum methods of radio reception.

During the transmission of discrete information the subject of transmission of report/communication (text, word, the data of combat situation etc.) are converted in the coding equipment/device of transmitter into the discrete sequence of code symbols. In binary systems, while subsequently will be examined the only such systems, code sequences are formed as the different combinations of two code symbols, which let us designate by letters Y_1 and Y_2.
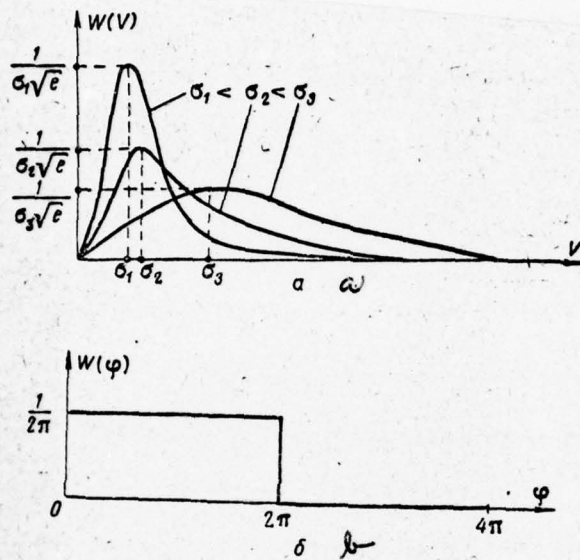
Fig. 2.6.4.

~~Page Ortho~~. Page 84.

As an example of such code symbols can serve, for example, the premise/impulses of "pressure" and "releases" (" 1" and "0") in telegraph work. For a transmission along the communication channel the code combination is converted in the modulator of transmitter into signal. In this case to each cell/element of signal is placed in one-to-one correspondence certain code symbol.

Let us consider that in binary system to the cell/elements of signal or, for the sake of simplicity radi, to signals z_1 (t) and z_2 (t) correspond the code symbols Y_1 and Y_2 and vice versa. In the receptor of communicating system at the taken signals $z_r(t)$, r = 1; 2, are restored code symbols Y_r, r = 1; 2, then is created code combination and in the final analysis the transmitted report/communication.

Subsequently, by leaving aside the questions of coding and decoding, let us assume that the problem of investigation at the output/yield of receptor is making a decision about which of the signals $z_r(t)$ (symbols Y_r) it was transmitted in the course of the

determined time interval. This approach is typical for the study of

the freedom from interference of the transmission of discrete

information with piece-by-piece reception/procedure. It is obvious

that if the interferences in the communication channel were not or

they would have the determined character, then always it would be

possible to unambiguously restore/reduce the transmitted signal $z_r(t)$

in receptor and no problem here it would appear. However as it was

shown above, interferences in the communication channel always bear

that which is nonpredicted, i.e., to a certain degree random,

character. Therefore as would not be accepted the solution to which

of the signals $z_r(t)$ it was transmitted, always possible error, i.e.,

observer it can state that is accepted signal z_1 (t), when in

actuality was transmitted z_2 (t) and vice versa.

Thus, observer must carry out a selection between two possible

confirmations: H_1 is accepted signal z_1 (t) and H_2 - is accepted

signal z_2 (t). These two confirmations are called hypotheses. In this

case must be selected that from the possible hypotheses, that with

the greatest probability corresponds to reality. Since greatest how

disposes of the observer in the relation to interferences, their this

statistical (probabilistic) description of of as random processes

(see §2.2), selection of one of the possible hypotheses was called

the name testing statistical hypotheses.

Page 85.

Of course, during statistical testing hypotheses observer must
utilize such a strategy (this method of treatment/working received
signals and interferences in receptor) in order that the solution
would be accepted with the greatest possible success or (that the
same) in order that losses due to making erroneous decisions would be
smallest possible. But this will depend on the character of signals,
interferences, and also on the determination of the concept of
success.

In order to explain the aforesaid, let us turn for example. Let
in binary communicating system for the transmission of discrete
information be utilized signals z_1 (t) and z_2 (t) whose structure
(amplitude, frequency, the initial phase) is known in all parts. Are
known also time of the arrival of signal and statistical properties
of the noises which enter together with signal for the input of
receptor. Let W (x/z_1) and W (x/z_2) - the probability density of the
possible values of the sum of signal and interferences at the input

of receptor respectively in hypotheses H_1 (is accepted signal z_1 (t)
and H_2 (is accepted signal z_2 (t)). In the case of probability
density W in question (x/z_2) they do not depend on the unknown
parameters of signal. Therefore hypotheses H_1 and H_2 are called
simple. We will also describe also to the expected signals $z_r(t)$ some
expected or a priori (pretest) probability of their appearances,
which can be determined, for example, by means of analysis sufficient
long code sequences, utilized in communicating system. Let to signal
z_1 (t) correspond the a priori probability q. Then to signal z_2 (t)
it corresponds to probability (1-q). Observer's problem lies in the
fact that, in order by the measurement of value x in receptor for
time of the duration of the cell/element of signal t to determine
with the greatest possible success, which of the signals $z_r(t)$ is
accepted. As strategy it can take the division of the interval of all
possible values x of the adopted sum of signal and interferences for
two ranges (Fig. 2.7.1a): G_1 (values x range from $-\infty$ to certain x_0)
and G_2 (values x range from x_0 to ∞). If value x is included in range
G_1, then observer solves, that is valid hypothesis H_1. With respect
is accepted hypothesis H_2, when x is located in range G_2.

Page 86.

Figure 2.7.1a shows that no matter how was selected threshold value x_0, is always feasible the case, when the solution is accepted incorrectly. Actually, if was transmitted signal z_1 (t), then the probability of error (acceptance of hypothesis H_2) is determined by the area, limited by ordinate at point x = x_0, the density curve of probability W (x/z_1) and by the axis of abscissas ($x_0 < x < \infty$), and is numerically equal to

$$p_1 = \int\limits_{x_0}^{\infty} W\left(\frac{x}{z_1}\right) dx. \qquad (2.7.1)$$

Respectively during the transmission of signal z_2 (t) the probability of error (acceptance of hypothesis H_1) is equal to

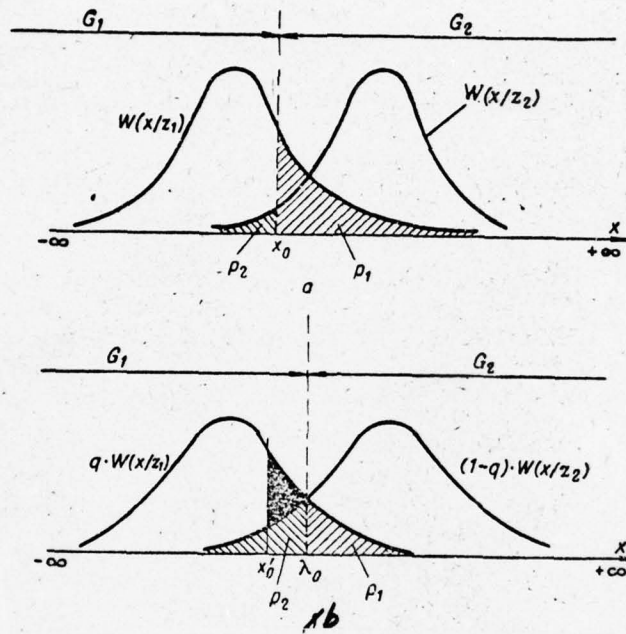$$p_2 = \int\limits_{-\infty}^{x_0} W\left(\frac{x}{z_2}\right) dx. \qquad (2.7.2)$$

Fig. 2.7.1.

~~Digested~~ Page 87.

in order that losses due to making erroneous decisions would be smallest possible, observer must select in an appropriate manner threshold value x_0. The latter depends on the cost for it the probabilities of errors p_1 and p_2.

The contemporary transmission systems of discrete information must very accurately restore the transmitted report/communications. In this case it can happen so that the error in the reception/procedure at least of one transmitted symbol considerably will lower the value of entire taken report/communication. From this viewpoint any error with reception/procedure is undesirable. Therefore it is reasonable to set/assume the relative values of the probabilities of errors identical and equal, for example, to unity. Then the total probability of the erroneous reception of the cell/element of signal is determined by the relationship/ratio

$$p = qp_1 + (1-q)p_2. \qquad (2.7.3)$$

It is necessary to select this value λ_0 from all possible x_0, which would minimize probability of p. For determining this threshold value it is necessary to differentiate (2.7.3) in terms of x_0 and to equate derivative zero.

By executing these operations, we will obtain

$$\lambda_0 = \frac{W\left(\frac{x_0}{z_2}\right)}{W\left(\frac{x_0}{z_1}\right)} = \frac{q}{1-q}, \qquad (2.7.4)$$

where λ_0 - the optimum threshold level, depending on the a priori

probabilities of the transmission of signals z_1 (t) and z_2 (t).

Figure 2.7.1b depicts graphically to function qW (x/z_1) and $(1-q)$ W (x/z_2) - the probability densities of the adopted sum of signal and interferences in transmission z_1 (t) and z_2 (t), "suspended" (multiplied) in accordance with the a priori probabilities of these signals. The position of the optimum threshold λ_0, defined by formula (2.7.4), must be by such, as this shown in figure. Then complete probability of the erroneous reception of the cell/element of signal is equal to the shaded range and smallest possible. If threshold level is selected not in accordance with expression (2.7.4), then complete probability of error will increase. Actually, let be selected the threshold $x_0' < \lambda_0$. Then probability of error during the transmission of signal z_1 (t) increases and will be equal to $q\int_{x_0'}^{\infty} W(x/z_1)dx$, a with the transmission of signal z_2 (t) - it decreases will be equal to $(1-q)\int_{-\infty}^{x_0'} W(x/z_2)dx$.

Page 88.

However, the composite probability of error, equal to the sum of these probabilities, will increase to the value of the blackened

area. Analogously complete probability cf error will increase, if $x_0' > \lambda_0$.

Formula (2.7.4) has large value, since it indicates such a strategy, by which the losses due to making erroneous decisions will be smallest possible. For this in receptor during treatment/working the adopted sum of signals and interferences must be calculated the relation

$$\lambda(x) = \frac{W\left(\frac{x}{z_2}\right)}{W\left(\frac{x}{z_1}\right)}, \qquad (2.7.5)$$

called likelihood ratio. Then value $\lambda(x)$ must be compared with threshold $\lambda_0 = q/1-q$. If $\lambda(x) > \lambda_0$, then is valid hypothesis H_2, but if $\lambda(x) < \lambda_0$, then is valid hypothesis H_1. Emphasizing that fact that into the "responsibility" of receiver in this case enters not only the computation of likelihood ratio, but also the delivery of the solution to which of the signals is accepted, they speak about the decisive schematic of receptor. It is natural that the

constructed in accordance with rule (2.7.5) decisive schematic of receptor minimizes complete probability of the error and in this sense is optimum.

The criterion of the minimum composite probability of error very frequently is called Kotelnikov's criterion or by ideal observer's criterion [20]. This criterion extremely widely is utilized during the development of the transmission systems of discrete information for the investigation of the questions concerning the structure of the optimum decisive schematic of receptor and its freedom from interference in the different interference situations, which appear in the communication channels.

~~end section~~.

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                                     191

Page 89.

In this case the composite probability of the error in the optimum

decisive schematic, based on Kotelnikov's criterion, depends only on

the character of interferences in the communication channel. The

freedom from interference of this schematic is called potential. By

comparing the freedom from interference of real receptors with

potential, it is possible to determine the possibilities in principle

of an increase in the freedom from interference in this or another

method of reception.

Subsequently we will assume that a priori the probability of the transmissions of signals z_1 (t) and z_2 (t) are identical and equal to [1].

$$q = 1 - q = \frac{1}{2}. \qquad (2.7.6)$$

FOOTNOTE [1]. In work [36] it is proven, that in any rationally constructed transmission system of discrete information the symbols of code sequences are approximately equiprobable. ENDFOOTNOTE.

It is not difficult to find, as must be constructed in this case the decisive schematic of the optimum according to Kotelnikov receiver. This receiver must calculate during treatment/working the adopted sum of signals and interferences likelihood ratio λ (x), determined by relationship/ratio (2.75). Moreover the optimum threshold level on the basis of formula (2.7.4) is equal to

$$\lambda_0 = 1. \qquad (2.7.7)$$

The condition of recording signals z_1 (t) and z_2 (t) takes the

following form:

$$\text{if } \lambda (x) > 1, \text{ then is accepted } z_2 (t)$$

$$\text{(is valid hypothesis } H_2)$$

$$\text{if } \lambda (x) < 1, \text{ then is accepted } z_1 (t) \qquad (2.7.8)$$

$$\text{(is valid hypothesis } H_1).$$

Very frequently the construction of the decisive schematic substantially is simplified, if in processing of the adopted sum of signal and interferences is calculated not likelihood ratio, but certain by the corresponding shape the selected monotonic function of $\lambda (x)$. In channels with normal interferences as this function most conveniently it is to utilize functions of form $\ln \lambda (x)$. In this case, taking into account that $\ln \lambda_0 = 0$, we will obtain the following equivalent (2.8.8) rule of the solution:

$$\text{if } \ln \lambda (x) > 0, \text{ then is accepted } z_2 (t);$$

$$\text{if } \ln \lambda (x) < 0, \text{ then is accepted } z_1 (t). \Bigg\} (2.7.9)$$

Page 90.

Let us examine important examples of the use of relationship/ratio
(2.7.9) for the construction of the optimum according to Kotelnikov
decisive schematic of receptor.

Optimum decisive schematic in channel with the constant
parameters and the normal fluctuating interference (coherent
reception). Let in the binary transmission system of discrete
information be utilized the signals $z_r(t)$, whose structure is
determined by relationship/ratio (1.1.1) and is completely known with
reception up to the phase of their high-frequency filling. The
communicating system, in which during processing received signal is
utilized the a priori information about the phase of high-frequency
filling, was called the name the systems of coherent reception. The
durations of both versions of signal we set/assume by identical and
equal T. Let also in the communication channel operate the additive
fluctuating interference ξ (t) in the form of normal white noise, and
the parameters of the communication channel: transmission factor μ

and propagation time are known with reception and are constant. Let us accept for the zero time reference the torque/moment of the beginning of the reception of the transmitted cell/element of signal. Then the received signal

$$x(t) = \mu z_r(t) + \xi(t), 0 \leqslant t < T. \qquad (2.7.10)$$

The very important and widespread in practice case is the use in the binary transmission systems of the discrete information of signals with identical energy (energies):

$$P_1 = P_2 = P_0; P_r = \frac{\mu^2}{T} \int\limits_0^T z_r^2(t) dt. \qquad (2.7.11)$$

Such systems were called the name systems with active pause. Let us determine, which must be in this case the structure of the optimum decisive receiver circuit and such the probability of the error of piece-by-piece reception in this schematic.

Likelihood ratio in the case in question takes form [20, 36]

$$\frac{v^2}{T} \ln \lambda(x) = X_2 - X_1, \qquad (2.7.12)$$

where v^2 - the spectral density of white noise;

$$X_r = \frac{2\mu}{T} \int_0^T x(t) z_r(t) dt \qquad (2.7.13)$$

- short-term crosscorrelation function between received signal x (t) and the expected signal

Page 91.

Then in accordance with rule (2.7.9) recording condition under

the receptor of the r version of signal takes the following form:

$$X_2 > X_1, \text{ is accepted } z_2 \ (t) ;$$
$$X_1 > X_2, \text{ is accepted } z_1 \ (t).$$
$$(2.7.14)$$

The obtained rule of solution (2.7.14) indicates that the optimum according to Kotelnikov receiver must be mutually correlated type receiver.

The decisive schematic of this receiver can be constructed as follows (Fig. 2.7.2). It contains two multipliers, to which are supplied received signal x (t) and signals μz_1 (t), μz_2 (t) from reference oscillators. From output/yields multiplier of the voltage/stresses are supplied to integrators, then the results of integration X_1 X_2 at the torque/moment of reading t = T are compared between themselves. Equipment/device of comparison issues the number of that symbol (Y_1 or Y_2), for which voltage/stress X_r, r = 1; 2, is more. After this is realized the jettisoning of the voltages in integrators and schematic is ready for the reception of the following cell/element of signal.

The main advantage of systems with resistive load is comprised
in the fact that the rule of solution (2.7.14) does not depend on the
transmission factor of the channel of communication/connection μ.
Actually, by taking into account relationship/ratio (2.7.13) and by
carrying out obvious reductions, the rule of recording the r version
of signal can be recorded in the form

$$\frac{1}{T}\int_0^T x(t)\,z_r(t)\,dt > \frac{1}{T}\int_0^T x(t)\,z_l(t)\,dt, \qquad (2.7.15)$$

with $r \neq l$.   Hence it follows that in systems with active pause
for the construction of the optimum according to Kotelnikov receiver
does not need a priori knowledge of the "scale" of the expected
signals, and is necessary only the knowledge of their form.

Of course, the created by the reference oscillators of sensing
transducer must strictly coincide in form with the expected signals
$z_r(t)$
and must be accurately synchronized with them. Howerver, the "scale"

of the signals of reference oscillators can be arbitrary, convenient
for the practical realization of circuit and, by certainly identical
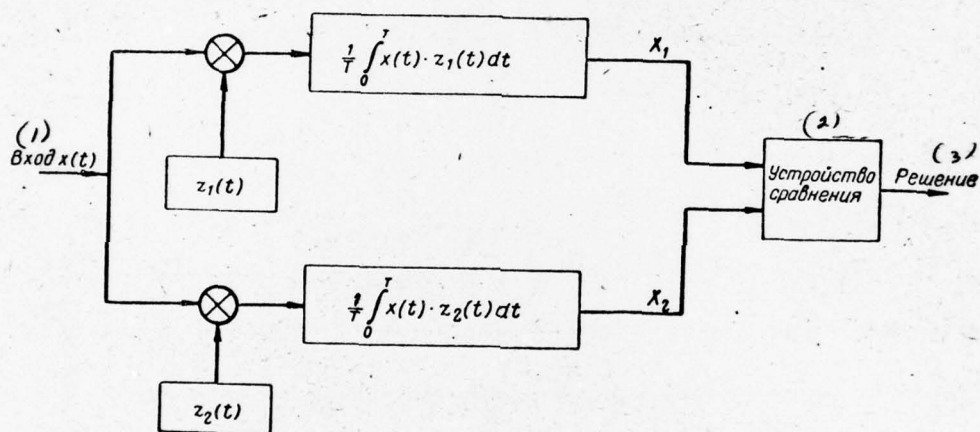for all reference oscillators. Page 92.

Fig. 2.7.2

Key: (1). / input x (t). (2). Equipment/device of comparison. (3).
Solution. Page 93.

This very important property of systems with active pause is retained
with signal fading [36]. Subsequently, if this is not specified
especially, we will examine the only systems with active pause. Let
us note that the realization of the decisive diagram of the optimum
receiver on multipliers is not the only possible. In the following
paragraph will be given examples of the decisive diagrams,
constructed on the so-called matched filters.

Further, the composite probability of the error with
piece-by-piece reception in the diagram in question is equal to [20]

$$p = \frac{1}{2}(p_1 + p_2) = \frac{1}{2}\left[1 - \Phi\left(h\sqrt{1 - \rho_{12}}\right)\right], \quad (2.7.16)$$

where $\rho_{12} = \frac{1}{P_0}\left[\frac{\mu^2}{T}\int_0^T z_1(t)\, z_2(t)\, dt\right]$ is a coefficient of the cross

correlation between signals z_1 (t) and z_2 (t);

$h^2 = \dfrac{P_c T}{v^2}$ - the ratio of the energy of signal to the spectral density of fluctuating interference;

$\Phi(u) = \sqrt{\dfrac{2}{\pi} \int_0^u e^{-\frac{x^2}{2}} dx}$ is an integral of probability or the function of Crump; this integral in elementary functions is not expressed; however for it are comprised detailed tables [3, 35].

As can be seen from relationship/ratio (2.7.16), the complete probability of error in the case in question it depends, in the first place, on the ratio of the energy of signal to the spectral density of fluctuating interference and, in the second place, on the coefficient of the cross correlation between signals. Therefore arises the question concerning how one should select the form of the utilized signals in order at the assigned value h^2 to ensure the greatest freedom from interference (minimum probability of error p).

Since $\Phi(u)$ is the monotonically increasing function of its argument, the greatest freedom from interference is reached when using such signals, with which ρ_{12} is minimal.

The greatest value of the coefficient of the correlation of ρ_{12} equal to +1 is reached at $z_1(t) = z_1(t)$. For such signals we have

$$\Phi(h\sqrt{1-1}) = 0$$

and, therefore, the probability of error is always equal to 1/2. In this finds its reflection that obvious fact, that identical signals cannot be distinguished. The minimum value of the correlation coefficient equal to -1 is reached at $z_1) t) = z_2(t)$, i.e., when signals they differ only in terms of sign. In other words, in this case the harmonic components of signal $z_2(t)$ have the same amplitudes as and components of signal $z_1(t)$, but their phases are shifted by angle π. The signals, for which $\rho_{12} = 1$, were called the name opposite. For such signals the probability of error is equal to

$$p = \frac{1}{2}\left[1 - \Phi\left(\sqrt{2}\,h\right)\right]. \qquad (2.7.17)$$

graph/diagram of the dependence of the probability of error p on value h^2 is represented in Fig. 2.7.3 (curve 1). Formula (2/7/17) determines (at the assigned value h^2) a maximally possible freedom from interference in the binary transmission system of discrete information.

Thus, the greatest freedom from interference during the transmission of discrete information possess the systems of coherent reception with opposite signals, which obtained the name of systems with phase modulation (FM). By the simplest example of this system is the system, in which are utilized the signals of the form

$$z_1(t) = A \cos(\omega t + \varphi) \; \text{и}$$
$$z_2(t) = -A \cos(\omega t + \varphi) = A \cos(\omega t + \varphi + \pi).$$

The advantages of phase modulation for the transmission of discrete information were known even in the beginning of thirtieth years. However, practical use FM long time blocked the phenomenon of the spontaneous jump/migration of the phase of the voltage in the reference oscillator of receptor, which brought to the so-called "reverse/inverse" work, by which the symbols "0" are accepted as "1" and vice versa.

Page 95.

The qualitative change began in the beginning of the fiftieth years, when prof. Petrovich n. T. proposed relative phase, or as it still call, phase-difference modulation (FRM), free from the indicated deficiency/lack and which possesses virtually such high freedom from interference, as FM.

With FRM the transmitted information is embedded not in the very value of the phase of the transmitted cell/element of signal, but in a phase difference of datum and preceding/previous cell/elements. For a binary system FRM this phase difference can take values 0 and π. Then for the transmission, for example, of symbol Y_1 is emitted the cell/element of the signal whose phase coincides with the phase of the preceding/previous cell/element, and for transmission Y_2 - cell/element with the phase, to the opposite phase of the preceding/previous cell/element and, etc.

At present the systems from FRM are very promising for the transmission of discrete information. The detailed information about the potential possibilities of such systems, about the realization principles of the construction of their equipment, about the newest developments of Soviet and foreign communicating systems with FRM the reader will find in work [14].
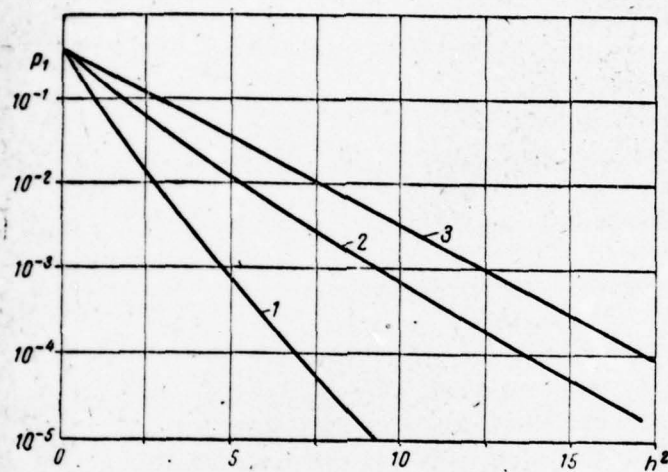
Fig. 2.7.3.

Page 96.

Besides systems with opposite signals practical interest are of the systems of coherent reception with the so-called orthogonal signals. In such systems signals z_1 (t) and z_2 (t) satisfy the condition of orthogonality under range from 0 to T:

$$\int_0^T z_1(t) z_2(t) \, dt = 0. \qquad (2.7.18)$$

Condition (2.7.18) is satisfied, for example, for a system with the frequency shift keying, which uses the signals

$$z_1 \ (t) = A \ \cos \ (k_1 \omega_0 t + \phi) \ \text{and}$$

$$z_2 \ (t) = A \ \cos \ (k_2 \omega_0 t + \phi),$$

where $\omega_0 = 2\pi/T$.

From relationship/ratio (2.7.18) it follows that when using
orthogonal signals the coefficient of their cross correlation of ρ_{12}
is equal to 0. Then expression for for probability error assumes the
form

$$p = \frac{1}{2}[1 - \Phi(h)]. \qquad (2.7.19)$$

Dependence of probability of error p as a function of h^2 is
represented in Fig. 2.7.3 (curve 2). From the comparison of
relationship/ratios (2.7.17) and (2.7.19) it is evident that for
opposite signals is provided the higher freedom from interference,
than for orthogonal. Moreover for achievement to one and the same
probability of error (reliability of reception) in system with

opposite signals is necessary two times the smaller power of
transmitter, other conditions being equal, than under system with
orthogonal signals.

    Optimum decisive diagram for signals with random initial phase
(incoherent reception). In practice frequently are encountered the
cases, when to define the values of the initial phase is impossible,
since its indeterminancy/uncertainty can be caused both conditions of
signal conditioning under transmitter and by the sufficiently rapid
changes in the state of the channels of propagation. On the other
hand, the determination of the values of the initial phase not is
always expedient, since expenditures on the system of a precise
tuning of phase can not justify the obtained increase reliability of
reception.

Page 97.

In a number of cases economically it is more profitable to obtain the
same increase in reliability by means of an increase in the power of
transmitted signal [36]. Communicating system, consider the a priori
information about the initial phase during processing received

signal, were called the name the systems of incoherent reception. Their advantage in the fact that with sufficiently high requirements for the authenticity of reception they provide, as will be shown below, relatively small energy loss in comparison with the systems of ccherent reception. At the same time such systems can be simpler realized.

For signals with active pause and the indefinite initial phase recording condition under the receptor of signals z_1 (t) and z_2 (t) assumes the form

$$V_2 > V_1, \text{ is accepted } z_2 \text{ (t)};$$

$$V_1 > V_2, \text{ is accepted } z_1 \text{ (t)}.$$

$(2.7.20)$

Here

$$V_r = V \overline{X_r^2 + X_r'^2} =$$

$$= \frac{2\mu}{T} V \overline{\left[\int_0^T x(t) z_r(t) \, dt \right]^2 + \left[\int_0^T x(t) z_r^*(t) \, dt \right]^2} \quad (2.7.21)$$

is a value envelope short-term crosscorrelation function between
received signal x (t) and the expected signal $\mu \overset{\cdot}{z}_r(t)$, at the moment
of time T; $z_r^{\cdot}$ (t) is the signal, conjugate/combined with $z_r(t)$
according to gilbert (him it is possible to obtain from $z_r(t)$ by the
path the phase displacement of all harmonic components to angle $\pi/2$).

This rule of the solution again tells about in the fact that the
optimum receiver for signals with indefinite initial phase also must
be mutually correlated type receiver. Moreover for recording the
taken signals in it must be calculated the values envelope short-term
crosscorrelation function V_r by specific relationship (2.7.21).

The decisive schematic of the receptor, which works in accordance with rule (2.7.20), can be constructed in the manner that this is shown in Fig. 2.7.4. It contains two reference oscillators, that reproduce the form of the expected signals z_1 (t) and z_2 (t) with an accuracy to the phase of high-frequency filling.

Page 98.

Let us note that the "scale" of the signals of reference oscillators both; and in diagram in Fig. 2.7.2, can be arbitrary, but compulsorily identical for both generators. The output voltages of reference oscillators enter two pairs of multipliers either directly or through phase inverters (FV) angle of 90°. At the output/yield of each phase inverter by means of the rotation of the phases of all harmonic components of the initial signal $z_r(t)$ are formed the signals z_r^* (t), conjugate/combined with initial according to gilbert. Voltages from the output/yield of each multiplier are integrated, as a result of which are formed the voltages, proportional to values X_r and X_r'. These voltages enter the nonlinear equipment/devices with square-law characteristics (square law detectors), and then the schematics of addition. After the addition of values X_r^2 and $X_r'^2$ are formed the voltages, proportional to values V_r^2.
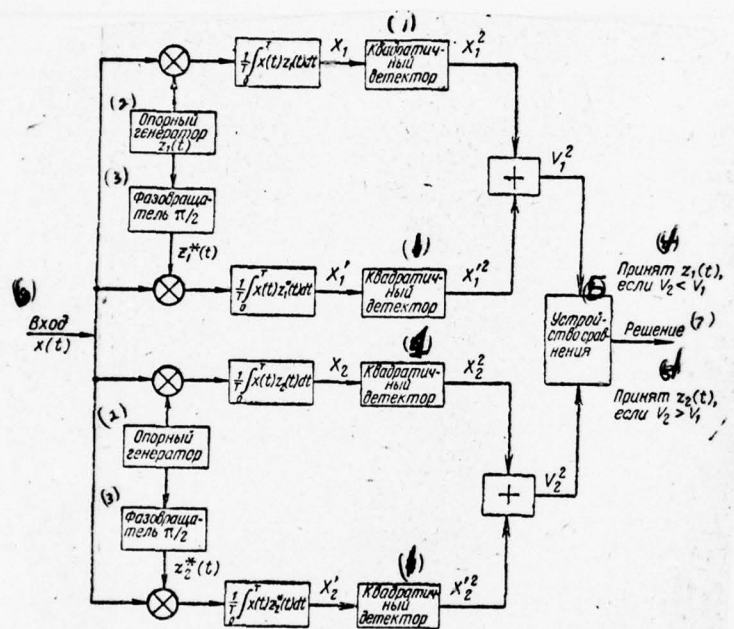
Fig. 2.7.4.

~~Fig. 2.7.4.~~

Key: (1). Square law detector. (2). Reference oscillator. (3). Phase
inverter $\pi/2$. (4). Is accepted z_1 (t) if $V_2 < V_1$. (5). Comparison
device. (6). Input / x (t). Page 99.

These voltages enter then into equipment/device of the comparison, in
which at the torque/moment of the termination of the element of
signal t = T conducts the reading and the comparison of values V_r^2. At
the output/yield of equipment/device of comparison is accepted the
solution to recording that signal (z_1 (t) or z_2 (t)), for which the
voltage V_r is more. Another version of the realization of rule
(2.7.20), based on the application/use of matched filters, is brought
in §2.8.

Let us examine briefly the question concerning the composite
probability of the error of piece-by-piece reception in diagram in
Fig. 2.7.4. As it follows from relationship/ratio (2.7.20), the
probability of error p_1 during the transmission of signal z_1 (t) is
equal to the probability of the fulfillment of inequality $V_2 > V_1$,
and the probability of error p_2 during the transmission of signal z_2
(t) - the probability of the fulfillment of inequality $V_2 < V_1$. The

composite probability of the error

$$p = \frac{1}{2}(p_1 + p_2)$$

was obtained and investigated in work [36]. Just as in the case of coherent reception, it finds to be dependent on value h^2 - ratio of the energy of the cell/element of signal to the spectral density of fluctuating interference and on a difference in the forms, utilized for the transmission of the information of signals. In this case for the assigned value h^2 by the greatest freedom from interference (inequality (2.7.20) are fulfilled most "strongly") they possess the systems, in which the utilized signals $z_r(t)$ are mutually orthogonal with any phase displacement of one of them, i.e.,

$$\left.\begin{aligned}\frac{1}{T}\int_0^T z_r(t)\,z_l(t)\,dt &= 0; \\ \frac{1}{T}\int_0^T z_r(t)z_l^*(t)\,dt &= 0\end{aligned}\right\} \qquad (2.7.22)$$

for all $r \neq l$.   Such signals were called name "orthogonal in the intensive sense" [36].

As examples of systems with orthogonal in the intensive sense signals can serve the systems of frequency telegraphy (ChT), of system, in which different symbols transmit by the mutually being noninterrupted in time momentum/impulse/pulses, etc.

end section.

Page 100.

For the optimum binary system of incoherent reception, which works in accordance with rule (2.7.20) and which uses orthogonal in the irtensive sense signals, the composite probability of error is equal to [36]

$$p = \frac{1}{2} e^{-\frac{h^2}{2}}. \qquad (2.7.23)$$

This expression determines the potential interference rejection of the binary systems of incoherent reception. The dependence of the probability of error p on value h^2 is represented in Fig. 2.7.3 (curve 3). From the figure one can see that the indeterminancy/uncertainty of phase comparatively little decreases the freedom from interference of systems with active pause.

Energy loss upon transition from coherent reception (curve 2) to incoherent for sufficiently high requirements for the authenticity of reception ($p = 10^{-4} - 10^{-5}$) does not exceed 15o/o (1 dB).

In conclusion let us note that in this paragraph the principles presented are base for the construction of the optimum broadband transmission systems of discrete information.

§2.8. Concept of matched filter. Optimum schematics of coherent and incoherent reception on matched filters.

   The examined in the preceding/previous paragraph optimum decisive schematics of coherent and incoherent reception can be realized not only on correlators with multipliers, but also on correlators in the form of the filters whose characteristics are agreed in the best way with the characteristics of the adopted against the background interferences of signals. In the literature such filters were called the name matched.

   Po.opredelenih, if $z_r(t)$ be the signal, taken against the background of white noise, then matched filter for this signal is called the linear filter, which has pulse response [5, 23]:

$$G(t) = az_r(t_0 - t), \qquad (2.8.1)$$

where a are certain constant, equal to the maximum amplification of filter;

t_0 - the fixed/recorded moment of time, with which is observed
the signal at the output/yield of filter.

Page 101.

From relationship/ratio (2.8.1) it is evident that the pulse
response of the filter, matched with signal $z_r(t)$, differs from the
function, which describes this signal, only in terms of constant
coefficient a, by displacement in time for value t_0 and by the sign
of the argument of time t. Emphasizing the latter fact, they say that
the pulse response of matched filter is mirror reflection relative to
axis t_0 the function $z_r(t)$, which determines the instantaneous values
of signal.

In order to visualize function G (t), let us turn to Fig. by
2.8.1, in which is shown the signal $z_r(t)$. It is obvious that the
function $z_r(t-t_0)$, shown by dotted line, delays with respect to
signal for a period t_0. Function $z_r(t_0-t)$ is mirror by it with
respect to axis t_0. After multiplying $z_r(t_0-t)$ to factor of
proportionality s (in figure a = 1), we will obtain pulse response
(2.8.1).

Very frequently in the literature is encountered the
determination of matched filter, datum with the aid of spectral
representations. Let $S_r(\omega)$ and $\varphi_r(\omega)$ be amplitude and phase spectra
of signal $z_r(t)$ respectively, and K (ω) and Φ (ω) - the
amplitude-frequency and phase-frequency response of linear filter.
Then the filter, matched with signal $z_r(t)$, is called the filter,
amplitude-frequency and phase-frequency responses of which are
determined by the relationship/ratios:

$$K(\omega) = aS_r(\omega); \qquad (2.8.2)$$
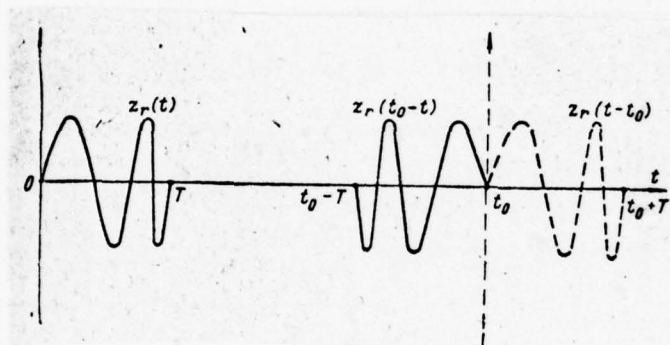$$\varphi(\omega) = -[\varphi_r(\omega) + \omega t_0]. \qquad (2.8.3)$$

Fig. 2.8.1

~~Fig. 2.8.1~~. Page 102.

In other words, the frequency characteristics of this filter with an accuracy to the constant amplitude factor a and the constant delay t_0 completely are defined by the spectral characteristics of signal, i.e., as "are agreed" with it. Of course, determinations (2.8.2), (2.8.3) and (2.8.1) are equivalent. The proof of this position can be

found in works [5, 6, 23].

Matched filters possess the following important properties.

1. Among all possible linear filters the matched filter provides at output/yield (at the moment of time $t = t_0$) the greatest ratio of the peak value of signal to the RMS value of white noise. This relation is proportional $h = \sqrt{\dfrac{P_c T}{v^2}}$. It is determined only by energy of signal $P_c T$, by spectral noise density v^2 and in no way depends on the form of the utilized signal (duration of signal, the width of the spectrum or from any other special feature/peculiarities of structure), which is very important for applying broadband noise-like signals.

Omitting the simple formal proof of this property, which can be found, for example, in [5, 23], let us examine its physical interpretation. In connection with the fact that the phase-frequency response of matched filter satisfies relationship/ratio (2.8.3), all harmonic components of signal $z_r(t)$ at the output/yield of this filter have at the moment of time $t = t_0$ one and the same phase, as a result of which occurs the arithmetical addition of their amplitudes.

Actually, the component of the signal of frequency ω at certain
moment of time t has the following resultant phase in the
output/yield of the filter:

$$\Theta(t) = \omega t + \varphi_r(\omega) + \varphi(\omega) = \omega t + \varphi_r(\omega) - $$
$$- \varphi_r(\omega) - \omega t_0 = \omega \ (t - t_0). \qquad (2.8.4)$$

With $t = t_0$, independent of frequency θ (t) = 0. Consequently, at
this moment of time the harmonic components of signal are accumulated
in phase and form peak overshoot (Fig. 2.8.2).

Page 103.

Harmonically the components of white noise, which have random phases
(see §2.6), crossing the matched filter, also will be obtained the
phase shifts $\phi(\omega) = - [\varphi_r(\omega) + \omega t_0]$, which, however, will not change
their random character. Because of this the result of the addition of
the components of noise at torque/moment t_0 at the output/yield of
filter will be random with by the negligible probability of the fact
that they will be formed in phase. The effect of matched filter on

the amplitudes of the components of signal and noise completely is
determined by the amplitude-frequency characteristic of filter
(2.8.2), repeating the amplitude spectrum of signal. Because of this
the filter provides the best isolation/liberation of the most intense
regions of the spectrum of signal, and weak regions of the spectrum
it attenuate/weakens, whereupon the weakening the greater, the lesser
the intensity of the corresponding component. Filter as "emphasizes"
those components, which most strongly affect the peak value output
the voltage of signal. The weakening of noise spectrum, uniform at
the input of filter, is observed at all frequencies, with the
exception only those, that correspond to the maximums of the spectrum
of signal. Thus, the frequency characteristic of matched filter
provides the maximum possible signal-to-noise ratio at output/yield.

2. As noted, the maximum signal level is reached on output of
matched filter at the moment of time $t = t_0$. Value t_0 characterizes
the delay time in the signal after the passage of matched filter.
From formula (2.8.2) it follows that t_0 must be not less than the
moment of time T of the termination of input signal, i.e.,

$$t_0 \geqslant T. \qquad (2.8.5)$$
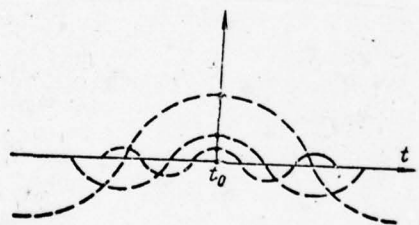
Fig. 2.8.2.

Page 104.

Otherwise the optimum filter would develop on its output/yield voltage even before to its input into torque/moment t = 0 will enter

signal [1].

FOOTNOTE [1]. Actually, since $z_r(t)=0$ with $t \geqslant T$, as it follows from (2.8.1), in the case $t_0 < T$ possible it would be $G(t)\neq 0$ with $t < 0$, i.e., even to the torque/moment of the effect of input signal. ENDFOOTNOTE.

It is obvious that this filter virtually could not be carried out. In the general case the delay time t_0 can be arbitrary, provided in this case was satisfied relationship/ratio (2.8.5). Usually to avoid an excessive signal delay at the output/yield of filter and for simplification in the structure of the matched filter they select

$$t_0 = T. \qquad (2.8.6)$$

3. Matched filter possesses the property of invariance relative to amplitude, temporary situation and the initial phase of signal [23]. This means that independent of a delay of the received signal in the transmission channel of a change in its amplitude and initial

phase of high-frequency filling of the characteristic of matched filter they do not change. Actually,, as this follows from relationship/ratios (2.8.2) and (2.8.3), to signal that which is differing from datum only by amplitude, by delay and the initial phase, will correspond the only another values constant a and t_0 without a change in the frequency characteristics of matched filter. The property of invariance has great practical value. In the information circuits amplitude, delay and the initial phase of the received signal, as a rule, are subjected to random changes. However, instead of constructing the enormous number of filters, each of which would be agreed with the signal, having the determined values of amplitude, delay and the initial phase, sufficient to construct only one filter, matched with all signals of this form.

4. Finally, the remarkable property of the matched with siganl $z_r(t)$ filter lies in the fact that this filter is the equipment/device, which measures the crosscorrelation function between received signal x (t) and the expected signal $z_r(t)$.

Page 105.

Actually, the output potential of any linear filter at certain moment of time t is determined according to superposition theorem by following expression [38]:

$$u_{\text{вых}}(t) = \int\limits_{-\infty}^{\infty} u_{\text{вх}}(t') \, G\,(t - t') \, dt', \qquad (2.8.7)$$

where $u_{\text{вх}}(t)$ is input voltage;

G (t) - the pulse response of filter.

Then, assuming that $u_{\text{вх}}(t)$ there is received signal x (t), and also by taking into account relationship/ratio (2.8.1) and (2.8.5), we will obtain that output potential of the matched filter

$$u_{\text{вых}}(t) = a \int\limits_{-\infty}^{\infty} x\,(t') \, z_r\,(t' - \tau) \, dt', \qquad (2.8.8)$$

where $\tau = t - T$.

Comparing **this expression** with formula (2.3.12), we comprise
that the voltage $u_{\text{ВЫХ}}(t)$ differs only in terms of constant factor a
that crosscorrelation function between received signal x (t) and the
expected signal $z_r(t)$. The difference within integration limits in
formulas (2.8.8) and (2.3.12) does not have a value, since the
duration of received signal x (t) is limited by range from 0 to T and
outside this interval becomes zero.

Thus, the matched with the expected signal $z_r(t)$ filter is
completely equivalent to the correlator of Fig. 2.3.4a, which
measures the crosscorrelation function of processes x (t) and $z_r(t)$.
This fact has great practical value, since it shows that the
schematics of the optimum receivers can be constructed also on
matched filters. Moreover in a number of cases (especially when using
broadband signals in channels with multi-beam characteristics) such
decisive schematics are realized considerably simpler than on
multipliers.

The expressed positions relative to the properties of matched

filter can be illustrated with the aid of following simple example [6].

Page 106.

Figure 2.8.3a depicts the schematic of the formation of broadband signal, which contains one delay line in removal/outlets, the dialing/set of narrow-band filters $\omega_1, \omega_2, \ldots, \omega_n$ and amplitude corrector $S_r(\omega)$. The exciting narrow pulse is delayed in the removal/outlets of line for a period $\tau_1, \tau_2, \ldots, \tau_n$. Then from its spectrum in each removal/outlet is isolated the corresponding spectral component of the form/shaped signal.
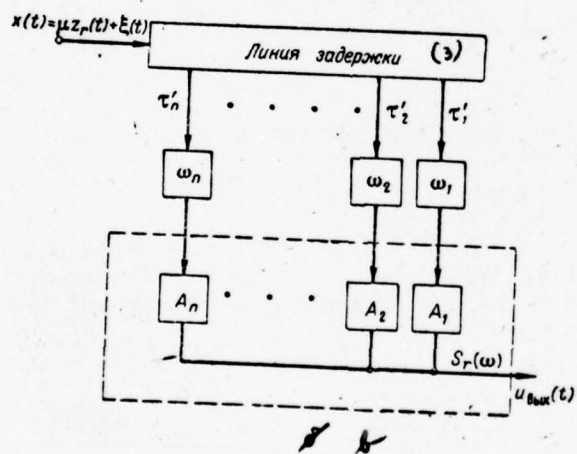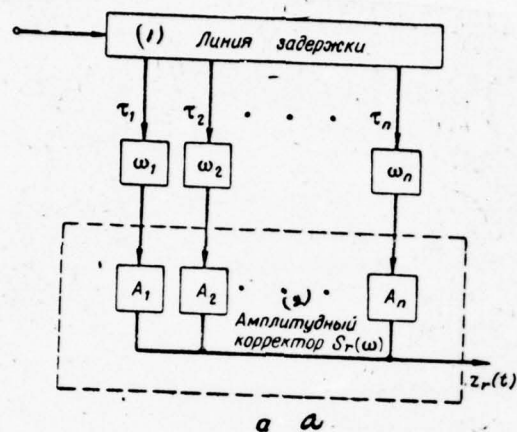
Fig 2.8.3.

Key: (1). Delay line. (2). Amplitude corrector. (3). Delay line. Page 107.

The amplitudes of the components of signal A_1, A_2, ..., A_n are form/shaped with the dialing/set of amplitude corrector's attenuators (attenuators) [1].

FOOTNOTE [1]. Amplitude corrector can be carried out also in the form of linear filter with characteristic $S_r(\omega)$. ENDFOOTNOTE.

The formed harmonic components of signal are summarized in common/general/total busbar at amplitude corrector's output/yield, forming signal $z_r(t)$. Its amplitude spectrum is determined by corrector's characteristic $S_r(\omega)$. The phase spectrum of signal $\varphi_r(\omega)$ depends on the selection of the initial phases, determined by delay factors τ_1, τ_2, ..., τ_n. Selecting in an appropriate manner of the value of amplitudes A_n and of delays τ_n, it is possible to form broadband signal with sufficiently noise-like structure.

It is not difficult to see that the diagram in Fig. 2.8.3a is
changed form version examined into §1.1 the diagrams of the formation
of broadband signal (see Fig. 1.1.3), if we in the latter interchange
the position filters and delay lines.

Let us examine, as must be constructed the matched with signal
$z_r(t)$ filter.

First, from condition (2.8.2), set/assuming for simplicity a =
1, we obtain, that the amplitude-frequency characteristic of filter
coincides with the amplitude spectrum of signal $S_r(\omega)$. Consequently,
in the construction of filter it is necessary to provide the
dialing/set of narrow-band filters for frequencies $\omega_1, \omega_2, ..., \omega_n$ and
amplitude corrector, analogous to the corrector of shaper.

In the second place, the phase-frequency response of filter
differs from phase spectrum only in terms of sign. Therefore for the
synthesis of matched filter it is necessary to change only the delay
time in each component so that the obtained by them phase shifts

would change sign. This can be reached with the aid of delay line in the removal/outlets, connected "mirror" with respect to the line of shaper. As a result the matched with signal filter assumes the form of Fig. 2.8.3b. The delay time in the removal/outlets $\tau'_1, \tau'_2, \ldots, \tau'_n$ is selected so that is satisfied the condition

$$\tau_1 + \tau'_1 = \tau_2 + \tau'_2 = \ldots = \tau_n + \tau'_n = t_0. \qquad (2.8.9)$$

Then the phases of all harmonic components of the adopted useful signal $\mu z_r(t)$ are formed in phase in the output/yield of filter at the moment of time t_0.

Page 108.

With satisfaction of condition (2.8.6) at the torque/moment of the termination of the cell/element of signal T is provided the ceiling voltage of useful signal and the maximum excess by it interference ε (t). The character of a change in the output voltage of filter according to (2.8.8) reflect/represents directly in time/temporary

scale the short-term mutually correlated function of the adopted and "reference" signals x (t) and $z_r(t)$.

Let the interference ξ (t), which enters the input of filter, be so small that it can be disregarded, i.e., $x(t) \approx \mu z_r(t)$.   In this case on the basis (2.8.8) the output voltage of the filter

$$u_{\text{вых}}(t) = a\mu \int\limits_{-\infty}^{\infty} z_r(t') z_r(t-t+T) dt'. \qquad (2.8.10)$$

Whereas it follows from this expression, voltage $u_{\text{вых}}(t)$ maps into time the short-term autocorrelation function of the useful signal. In other words, matched filter transformed signal into its short-term autocorrelation function. If the amplitudes of harmonic components $A_1, A_2, \ldots, A_n$

are approximately identical, and frequencies $\omega_1, \omega_2, \ldots, \omega_n$ are selected different and multiple $\omega_0 = 2\pi/T$, voltage at the output/yield of filter it coincides with correlation function (2.5.9). Its change in time is shown in Fig. 2.8.4. As can be seen from figure, output potential of filter has the oscillatory character with equal in medium frequency the to spectrum of signal $z_r(t)$, and as the slowly being changed envelope. The maximum ceiling voltage occurs

in range $\pm$ 1/F relative to the torque/moment of the termination of signal. At the moment of time the reception of the subsequent cell/element of signal at the output/yield of filter there is no voltage, caused by the preceding/previous cell/elements of signal.

Let us define as one should construct the diagrams of the examined in the preceding/previous paragraph optimum receivers, utilizing matched filters. Page 109.
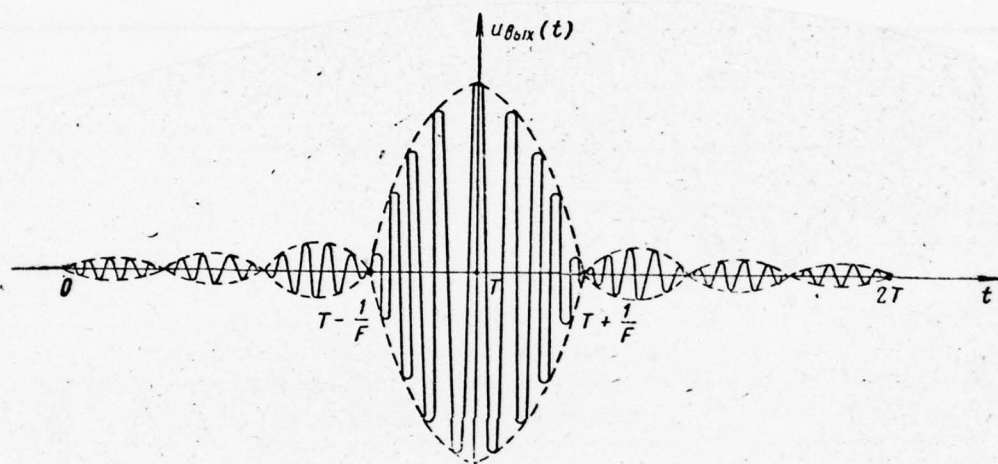
Fig. 2.8.4.

~~Нида С.Ми~~ Page 110.

Since $z_r(t)$ is identically equal to zero with $t < 0$ and $t \geqslant T$, of (2.8.8) it follows that the output potential of matched filter at the torque/moment of the termination of the cell/element of signal T

$$u_{\text{вых}}(T) = a \int_0^T x(t) z_r(t) \, dt = \frac{a}{\mu} X_r \qquad (2.8.11)$$

and with an accuracy to the constant factor a/μ coincides with the value of short-term crosscorrelation function X_r between received signal x (t) and the expected signal $\mu z_r(t)$. Therefore in the case of the coherent reception of signals with active pause the decisive diagram of the optimum receiver assumes the form, shown in Fig. 2.8.5 (compare with Fig. 2.7.2). Received signal x (t) enters the matched with signals z_1 (t) and z_2 (t) filters.

The voltages from the output/yield of matched filters at the torque/moment of reading t = T enter the equipment/device of the comparison, which selects that signal $z_r(t)$, for which obtained

greatest voltage. It is obvious that this diagram realizes the rule of solution (2.7.14) and the probability of the error in it is determined by relationship/ratios (2.7.17) and (2.7.19) for opposite and orthogonal signals respectively. In this case just as for a circuit on multipliers, in the case of using signals with active pause the "scale" of the output voltages of filters, determined by factor a/μ, can be arbitrary, convenient for a practical realization and identical for both matched filters.
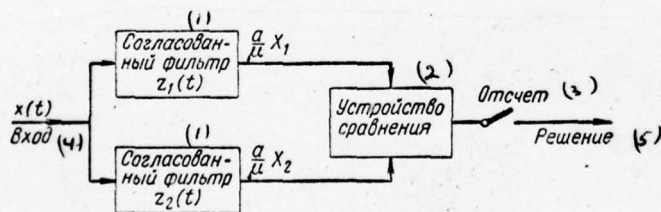

Fig. 2.8.5.

Key: (1). Matched filter z_1 (t). (2). Equipment/device of comparison. (3). Reading. (4). x (t) / input. (5). Solution. Page 111.

Let us note, that even so at the torque/moment of time T of the

output potential of matched filter and on the output/yield of
correlator with multiplier they coincide with an accuracy to constant
factor however for the moments of time t < T these diagrams give
substantially different voltages. Figure 2.8.6 shows the
exemplary/approximate form cf voltage from broadband signal on the
output/yield of correlator with multiplier. As can be seen from
figure, this voltage is the monotonically growing from zero to
function. Voltage on the output/yield of matched filter is the
oscillatory function (see Fig. 2.8.4) with the amplitude, which grows
from zero to $\frac{a}{\mu} X_r$.

As a result of the oscillating character of the output voltage
of the matched filter of requirement for accuracy of reading
(synchronization of reading) in diagram in Fig. 2.8.5 significantly
higher than in diagram in Fig. 2.7.2. In diagram with matched filters
the permissible displacement of fiducial mark must be much less than
the period of high-frequency oscillation, i.e., is much less than
value $2\pi/K\omega_0$ (in the general case much less than value $\frac{1}{f_{cp}}$, where f_{cp}
is the medium frequency of the spectrum of signal). Otherwise is
possible large error and even a sign change of output voltage. In
diagram on multipliers are permissible the considerably greater
displacement of fiducial mark, provided they remained much less than
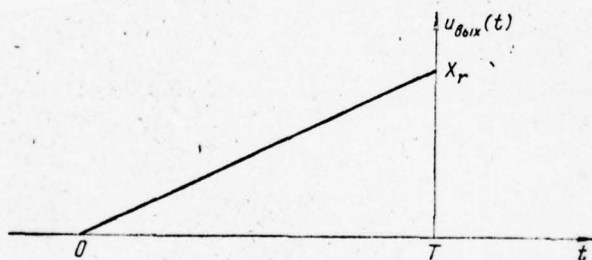the duration of the cell/element of signal T.

Fig. 2.8.6. Page 112.

However, when using a coherent reception this special
feature/peculiarity of diagrams on multipliers is unessential, since
in such diagrams are presented very stringent requirements for the
accuracy of the synchronization of received signal with reference
oscillators (also order $2\pi/K\omega_0$). Requirements for synchronization

(either reading or reference oscillators) can be considerably lowered
when using circuits of incoherent reception [36].

The use of matched filters in the diagrams of incoherent
reception is based on the isolation/liberation of the envelope of
output potential of filter. If $u_{вых}(t)$ is an instantaneous value of
the output voltage of filter at the moment of time $0 \ll t \ll T$, then
the envelope $\dot{E}_{вых}(t)$ of this voltage is equal to

$$E_{вых}(t) = \sqrt{u_{вых}^2(t) + u_{вых}^{*2}(t)}, \qquad (2.8.12)$$

where $\dot{u}_{вых}(t)$ is the function, conjugate/combined according to
gilbert with $u_{вых}(t)$. It is possible to sfow (for example, see [36]),
that the value of envelope at the moment of time t = T with an
accuracy to constant factor coincides with value V_r, by specific
relationship (2.7.21), i.e.,

$$E_{вых}(T) = k_E V_r, \qquad (2.8.13)$$

where k_E - certain constant coefficient.

Then the optimum decisive diagram of incoherent reception
assumes the form, presented in Fig. 2.8.7. Received signal x (t)
enters the matched with signals z_1 (t) and z_2 (t) filters, output
voltages of which pass through the squaring equipment/devices (square
law detectors). At the torque/moment of reading T of the voltage from
the output/yield of detectors, proportional to values V_1^2 and V_2^2, are
introduced into equipment/device of the comparison, which selects
that signal, for which obtained larger voltage. The probability of
the error of piece-by-piece reception in this diagram when using
orthogonal in the intensive sense signals is determined by
relationship/ratio (2.7.23). As concerns requirement for accuracy of
reading, preliminarily let us note that in diagram in Fig. 2.8.7
permissible displacement of fiducial mark must be much less than
value 1/F, where F is the conditional frequency band, occupied by
signal. For the utilized broadband signals always is fulfilled the
inequality

$$\frac{1}{F} \gg \frac{1}{f_{op}}. \qquad (2.8.14)$$

Therefore in the diagrams of the incoherent reception of requirement for the synchronization of reading it is considerably easier than with coherent reception. In more detail this question is examined in chapter 3.

(1) Согласованные фильтры   (2) Квадратичные детекторы

$z_1(t)$ $k_\xi V_1$ $k_\xi^2 V_1^2$

(3) Устройство сравнения   Отсчет (4)

Решение (5)

$z_2(t)$ $k_\xi V_2$ $k_\xi^2 V_2^2$

Fig. 2.8.7.

Key: (1). Matched filters. (2). Quadratic detectors. (3). Equipment/device of comparison. (4). Reading. (5). Solution.

end section.

Pages 113-140.

§2.9. Fundamental cell/elements of correlators in the systems of broadband communication/connection.

In the preceding/previous paragraphs it was established/installed that the correlators of the receivers of broadband communicating systems are the multiplying, retarding and integrating equipment/devices. Furthermore, the necessary elements of

the decisive networks can be the summing and subtractors, which form part, for example, comparison circuit.

In the present paragraph is given the concept of operating principle, the possible diagrams and the construction of such equipment/devices. The more detailed study of these problems can be found in works [21, 23 and, etc].

Multiplying equipment/devices of correlators. The problem of the multiplying equipment/device (multiplier) is the formation/education of the product of two input signals, i.e., obtaining the output voltage, instantaneous values of which are proportional to the product of the instantaneous values of two conducted/supplied to its input tensions. In mutually correlated systems as output voltages are adopted x (t) and supporting/reference $z_r(t)$ the signals. In autocorrelation systems (see Chapter 4) multiplier must form product undelayed x (t) and delayed by the determined time x (t-τ) of signals.

Is known at present the large number of versions of the schematics of the multiplying equipment/devices. However, the work of

all them is based on the multiplication of input voltage either with
the aid of the special cell/elements, which possess the physical
effect of multiplication or with the aid of electron-tube or
transistor diagrams. Accordingly, are distinguished the
direct/straight and indirect methods of multiplication. Direct
methods utilize the physical effects, proportional to the product of
two measured values.

Page 114.

They can be based, for example, on the application/use of the Hall
effect.

The essence of the Hall effect consists of the following (Fig.
2.9.1). Let current I (t), that takes place through the cross section
of fine/thin metallic plate, be proportional to input voltage x (t),
and magnetic field H (t), perpendicular to the conducting plate, over
which occur/flow/lasts the current, it is proportional to the voltage
of reference signal $z_r(t)$. Then in the direction, perpendicular to
direction of flow and magnetic field, appears the so-called Hall
voltage, proportional to product I (t) H (t) and, consequently, also

$x(t)z_r(t).$

As the material of the conducting plate can be used bismuth, silicon, tellurium or semiconductor. Magnetic field is created by coils air-ccred.

The advantage of the direct methods of multiplication is the fact that they make it possible to directly obtain the unknown product. The accuracy of the work of direct/straight multipliers is very high. The divergence of the resulting stress from the product input can not exceed 1-2o/o. A deficiency/lack in such multipliers is large volume of expenditure on accessory equipment (electric power supply, supplementary amplifiers). Therefore the direct methods of multiplication wide acceptance were not received.

An at present preferred use find the indirect methods of the multiplications, with which the product of input voltage is obtained because of the formation/education from them of the corresponding sums and differences. Such multipliers are constructed for circuits by the use of electron tubes or transistors and require smaller expenditures. Although the accuracy of the indirect methods somewhat lower than of straight lines ones however for practical target/purposes it turns out to be completely sufficient.
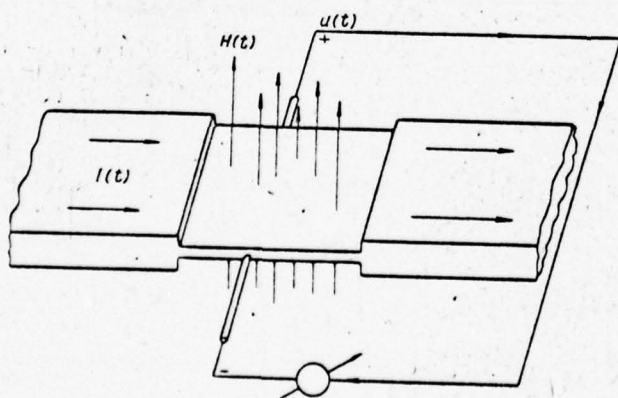
Fig. 2.9.1. Page 115.

From such multipliers it is possible to note the multipliers, which work on so-called quarter root-mean-square method [21], i.e., in accordance with the algebraic formula

$$xz = \frac{1}{4} [(x+z)^2 - (x-z)^2], \qquad (2.9.1)$$

where x and z - input voltage.

The functional diagram of the multiplier, which realizes this formula, is represented in Fig. 2.9.2. As can be seen from figure, the product of input voltage x (t) and $z_r(t)$ is obtained by the formation/education of sums and differences in them with the subsequent quadrature and the secondary subtraction. The special feature/peculiarities of the construction of the summing and subtractors are examined below. The squaring equipment/devices in diagram can be carried out on electron tubes or the semiconductor diodes, which have approximately square-law characteristics. A deficiency/lack in the diagram in Fig. 2.9.2 is the presence of two

squaring cascade/stages, which leads to certain of its complication.
Furthermore, the characteristic of any electron tube is not strictly
quadratic. Therefore for a decrease in the error in the work of
circuit it is necessary to complicate the construction of most
squaring equipment/devices. A very interesting example of simple
indirect multiplier is the diagram, published in work [48]. It is
represented in Fig. 2.9.3a. The operating principle of this
multiplier can be explained with the aid of the simplified equivalent
diagram in Fig. 2.9.3b. Under the action of the applied input voltage
x (t) through tubes $Л_1$ and $Л_2$ occur/flow/last currents $i_1 = i_0/2$ (1 +
ax) and $i'_2 = i_0/2$ (1 - ax), where i_0 is a cathode current $Л_1$ and $Л_2$:
a is the constant coefficient of proportionality, determined by the
mutual conductance of these tubes. Cathode currents $Л_3$ and $Л_5$ under
the action of input voltage $z_r(t)$ are equal to $i_3 = i_1/2$ (1 + bz)
and $i_5 = i_2/2$ (1 - bz), where b is the constant coefficient,
determined by slope/transconductance $Л_3$ and $Л_5$. Then taking into
account values i_1 and i'_2 we obtain that

$$i_3 = \frac{i_0}{4}(1+ax)(1+bz) = \frac{i_0}{4}(1+ax+bz+abxz);$$
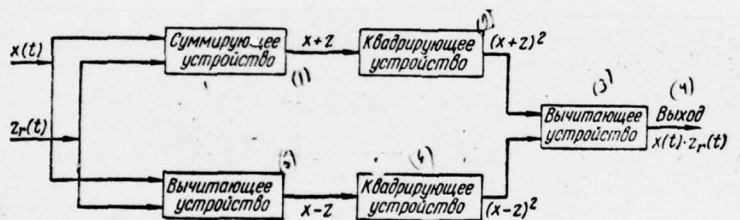$$i_5 = \frac{i_0}{4}(1-ax)(1-bz) = \frac{i_0}{4}(1-ax-bz+abxz).$$

Fig. 2.9.2.

Key: (1). Adder. (2). Squaring equipment/device. (3). Subtractor.
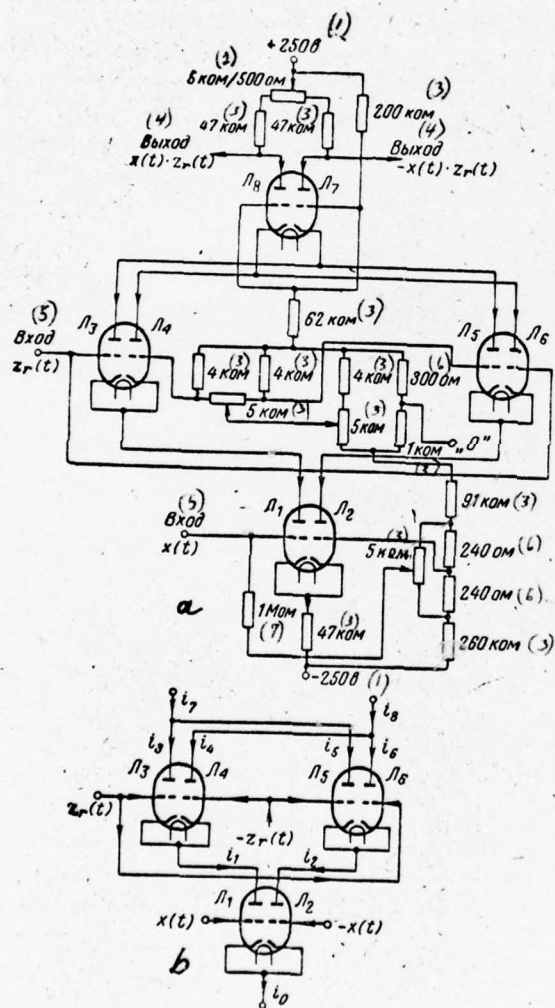(4). Output/yield. (5). Subtractor. (6). Squaring equipment/device.
Page 116.

Fig. 2.9.3

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Fig. 2.9.3.

Key: (1). V. (2). kΩ/Ω. (3). kΩ. (4). Output/yield. (5). Input. (6).
ohm. Page 117.

Cathode current Π_7 is equal to the sum of these currents:

$$i_7 = i_3 + i_5 = \frac{i_0}{2}(1 + abxz).$$

Analogously discussing, it is possible to show that

$$i_8 = i_2 + i_4 = \frac{i_0}{2}(1 - abxz).$$

A difference in currents i_7 and i_8 gives

$$i_7 - i_8 = i_0 abxz.$$

Thus, in common/general/total anode circuit Π_7 and Π_8 will be isolated the output voltage, proportional to product $x(t) z_r(t)$. According to the data [48] this multiplier works in the very wide frequency band of input voltage from 0 to 2 MHz with the sufficiently high accuracy of multiplication (order 1-2o/o).

Finally, as sufficiently simple multipliers is found a use of a diagram of circular balanced modulators (Fig. 2.9.4). Let us assume that all diodes of modulator are identical and have the exponential characteristics of form $i_A = A (e_A^{\gamma u} - 1)$, where A and γ are some constants; i_A - the current of diode; u_A - the voltage applied to diode. Then it is possible to show that the current of the load

$$i_{\text{R}} = \left[A\gamma^{2'} + \frac{A\gamma^2}{24}(x^2 + z^2) \right] xz = Bxz,$$

where B is certain constant value.

Consequently, the stress in the load of modulator is proportional to the product of input voltage x (t) and $z_{\pi}(t)$. Let us note that for the undistorted multiplication it is necessary to support constant value B, which depends on the sum of the squares $(x^2 + z^2)$ of the amplitudes of input voltage, i.e., it is necessary to support with constants the effective values of input voltage.
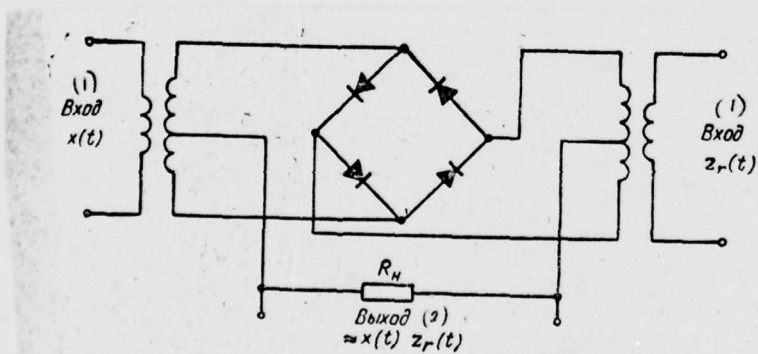
Fig. 2.9.4.

Key: (1). Input. (2). Output/yield. Page 118.

Circular balanced modulators as multipliers find a use at the
sufficiently high frequencies of input voltage - from several dozen
kilohertz of up to several dozen and even hundreds megahertz. In work
with broadband signals (noises) in them it is expedient to utilize
peak transformers on ferrite rings. The application/use of such
multipliers is very convenient in the correlators, to input of which
are supplied the voltages of different carrier frequencies, and
integration of product is conducted at the intermediate frequency,
equal to a difference in frequencies of input voltage. Therefore
double-balanced modulators are utilized as multipliers in the
so-called diagrams of synchronous heterodyning (see §3.2 and §3.5).

It must be noted that the modulator circuit as multiplier
differs somewhat from the mixer stage of receiver. This is explained
by the fact that its output voltage turns out to be proportional to
the product of input voltage, while the usual mixer stage of receiver
usually is controlled by the considerably larger voltage of
heterodyne.

Integrating equipment/devices of correlators. The integrating
cascade/stage of correlator serves for the formation/education of the
average value of product $x(t)z_\tau(t)$ (or x (t) x (t-τ)) for time of the

duration of signal T:

$$X_r = \frac{1}{T} \int_0^T x(t) z_r(t) dt. \qquad (2.9.2)$$

The simplest integrator is the integrating RC network (Fig. 2.9.5). Input $u_1(t) = x(t) z_r(t)$ and the output $u_2(t)$ voltages in it are connected by the relationship/ratio

$$RC \frac{du_2}{dt^2} + u_2 = u_1. \qquad (2.9.3)$$

By integrating the left and right side of this equation, it is possible to show that the output voltage into any point in time t is

equal

$$u_2(t) = \frac{1}{RC} e^{-\frac{t}{RC}} \int u_1(t) e^{-\frac{t}{RC}} dt. \qquad (2.9.4)$$

For the moment of time t = T << RC it is possible to count $e^{-\frac{T}{RC}} \approx 1$.   Then the solution is simplified and assumes the form

$$u_2(T) \approx \frac{1}{RC} \int_0^T u_1(t) \, dt. \qquad (2.9.5)$$

From (2.9.5) it follows that with sufficient slow response of the integrating circuit the output voltage approximately is proportional to integral of input voltage. This dependence the more

precisely, the more RC in comparison with T. However, increase RC for
an increase in the accuracy in the integration unavoidably leads to a
decrease in the value of input voltage. For example, let u_1 (t) is
constant during the duration of the cell/element of signal. Then
$u_2(T) \approx \frac{U_1 T}{RC}.$ With T = 10 ms and RC = 5 s we have u_2 (T) = U_1/500,
i.e., output voltage is attenuate/weakened as compared with input 500
times.

Page 119.

Consequently, during the application/use of the integrating RC
network it is impossible to obtain simultaneously the sufficiently
high value of output voltage and a small error of integration.

Sometimes in correlators as the integrating equipment/devices
are utilized the filters of low frequency. The examined integrating
RC network is the simplest example of such filters. To the filters of
low frequency as integrators are largely inherent the noted above
contradictions.

The better/best results, than the application/use of the integrating circuit, make it possible to obtain the widely utilized schematics of electronic integrators. The simplest electronic integrator is represented in Fig. 2.9.6. Let the cascade/stage work without the currents of the first grid. Then its grid circuit can be considered as the integrating circuit, which consists of resistor/resistance R_g and the input capacitance C_{BX}, equal to

$$C_{BX} = C_{gk} + (C_{ag} + C)(1 + K),$$

where C_{gk} and C_{ag} are the interelectrode capacitances of tube; K - the factor of amplification of cascade/stage.
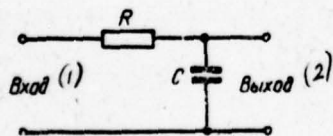
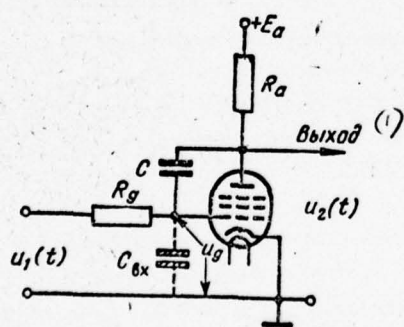

Fig. 2.9.5.

Key: (1). Input. (2). Output/yield.

Fig. 2.9.6.

Key: (1). Output/yield. Page 120.

If $C \gg C_{gk}, C_{ag}$ amplification factor is taken the order of several
dozens, then is input capacitance $C_{BX} \approx KC$. Then the voltage in
the circuit of control electrode with $T \ll R_g C_{BX}$ takes analogously
to (with 2.9.5) the form

$$u_g(T) \approx \frac{1}{KR_gC} \int_0^T u_1(t)\,dt. \qquad (2.9.6)$$

Consequently, the start of amplifier is equivalent to an increase in
the time constant of the integrating circuit in K once. Respectively
in K once will increase accuracy in the integration. The output
voltage, equal to

$$u_2(T) = -Ku_g(T) \approx -\frac{1}{R_gC} \int_0^T u_1(t)\,dt,$$

differs only in terms of sign from output potential of the
integrating circuit.

In the real schematics of integrators are utilized not one, but several step/stages of amplification, whereupon capacitance/capacity C is included from the output/yield of the last/latter cascade/stage to the input of the first. In such diagrams is provided the factor of amplification of the order of several hundreds or thousands. The diverse variants of the construction of the schematics of electronic integrators are given in [14]. Thus, electronic integrators make it possible by sufficiently simple means to obtain high accuracy in the integration and simultaneously high output voltage.

In contemporary correlators are applied also the integrators in the form of band-passs filter. Such integrators are utilized when to the input of the multiplier of correlator are supplied the voltages of different carrier frequencies. In this case the band-pass filter is tuned for different from zero difference frequency. As an example of this integrator serves tuned amplifier with the duct of high quality in anode circuit (Fig. 2.9.7). To the input of amplifier in the course of time from 0 to T enters voltage u_1 (t) certain difference frequency $k_1 \omega_0 = k_1 \, 2\pi/T$ proportional to the product of tensions x (t) and $z_r(t)$ Duct in anode circuit has quality Q, very narrow band (order 1/T) with resonance frequency $k_1 \omega_0$.
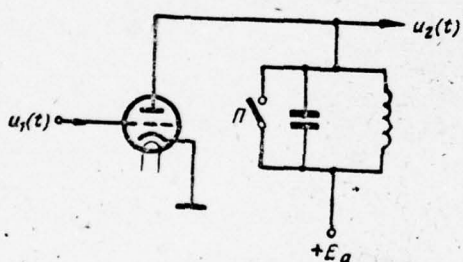
Fig. 2.9.7. Page 121.

The output voltage of amplifier is determined by input voltage u_1 (t)
and by the pulse reaction of plate circuit, which in this case takes
the form

$$G(t) = ae^{-\alpha t}\cos k_1\omega_0 t, \qquad (2.9.7)$$

where a - certain constant, which depends on the factor of amplification of cascade/stage;

$$a = \frac{k_1 \omega_0}{2Q}.$$

If the quality of duct is sufficiently high, so that for the moments of time $t \leqslant T$ correctly the inequality

$$\alpha t \ll 1,$$

that pulse reaction (2.9.7) assumes the form

$$G(t) \approx a \cos k_1 \omega_0 t. \qquad (2.9.8)$$

Let the input voltage be harmonic oscillation of the type

$$u_1(t) = A \cos k_1 \omega_0 t. \qquad (2.9.9)$$

Then output voltage for the moments of time $0 < t \leqslant T$ is equal

$$u_2(t) = \int_0^t u_1(x)\, G(t-x)\, dx. \qquad (2.9.10)$$

By substituting in this expression of value u_1 (t) and G (t) from (2.9.9) and (2.9.8), we will obtain

$$u_2(t) = \frac{aAt}{2} \cos k_1 \omega_0 t. \qquad (2.9.11)$$

Consequently, voltage u_2 (t) is the oscillatory function with the linearly growing in time amplitude (Fig. 2.9.8). At the torque/moment of the termination of the cell/element of signal T the output voltage is maximal.
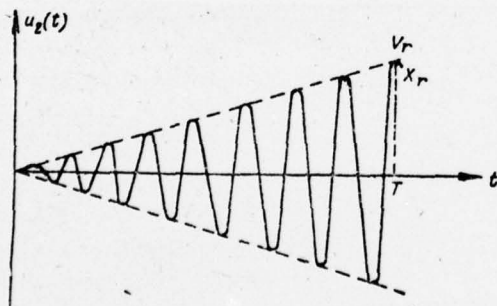
Fig. 2.9.8. Page 122.

Its instantaneous value at this moment is proportional to the
short-term mutually correlated function X of signals x (t) and $z_r(t)$

but amplitude (envelope) is proportional V_r - envelope short-term
crosscorrelation function of these signals. For broadband signals
input voltage u_1 (t) is the harmonic oscillation of frequency $k_1\omega_0$
with being slowly changed in time with an amplitude of A (t) and
phase ϕ (t). It is possible to show that in this case the output
voltage also is the oscillatory function with the monotonically
growing amplitude (similar to Fig. 2.9.8). After completion of input
signal at the moment of time T of oscillation in plate circuit
rapidly they are extinguished by the key/wrench Π, which shunts duct.
Oscillator circuit is ready for the reception of the following
cell/element of signal.

The duct of the high quality, equipped with equipment/device for
the instantaneous extinction of natural oscillations in it, was
called the name kinematic filter. Such filters find a use in the
technology of broadband radio communication. Specifically, they are
utilized in a vzimno- correlation system of the type "Rake" (see
Chapter 3).

Instead of the usual oscillatory circuits in kinematic filters
frequently are utilized electromechanical resonators (piezoelectric
or magnetostrictive). As equipment/devices of extinction in them are

applied electronic keying circuits. Let us note that such keying circuits are necessary also in all considered above diagrams of integrators.

Retarding equipment/devices. The retarding equipment/devices are intended for a formation from the output signals x (t) either $z_\gamma(t)$ of the delayed for a period τ signals x (t-τ) or $z_\gamma(t-\tau)$. The need for this delay appears during processing the multiple-pronged received signals x (t) in mutually correlated or autocorrelation broadband systems, during the construction of the matched with broadband signals $z_\gamma(t)$ filters and in a series of other cases.

As the retarding equipment/devices can in principle be used magnetic tapes, cathode-ray tubes with the accumulation of charges (charge-storage tubes), of delay line.

The preferred propagation as the retarding equipment/devices of broadband systems received delay lines. This is explained together of their design and operating advantages over other types of the retarding equipment/devices: comparative simplicity, relatively small overall sizes, the high accuracy of the installation of the required

delay time and the possibility of its arbitrary adjustment, high electrical characteristics (for example, a small distortion of output voltage), etc.

Distinguish electromagnetic (type of cut cable), ultrasonic, magnetostrictive and electrical (artificial lines with lumped parameters L and C) delay lines.

Electromagnetic and electrical delay lines make it possible to obtain with the permissible overall sizes a comparatively small delay time of the order of ones or at best of several dozen microseconds. Therefore they can be utilized, for example, in the autocorrelation broadband systems, for which are necessary short delay times, or for the construction of the broadband signals of a comparatively simple form in mutually correlated communicating systems.

Page 123.

Magnetostrictive delay lines make it possible to obtain in practice the delay time of the order of several dozen microseconds

and in this sense do not have special advantages over electrical
delay lines.

The ultrasonic lines provide a delay in the order of ones of
milliseconds. Their fundamental advantage consists of long delay time
of the unit of length, which is explained by comparatively small
velocity of propagation of ultrasound in conducting medium (1500-6000
m/s) [23]. As conducting medium in these lines are utilized either
liquid substances (mercury, mixture of water with ethyl alcohol), or
solids (magnesium alloys, vitreosil). Ultrasonic lines with liquid
conducting medium are comparatively bulky. Furthermore, in them it is
structurally difficult to obtain the intermediate step/stages of
delay time by means of the setting of supplementary removal/outlets.
The installation of the latter is necessary during processing
multiple-pronged signal in mutually correlated systems or with the
formation of the complex matched filters.

From the indicated deficiency/lacks are largely released the
lines with solid conducting medium. Most frequently are applied delay
lines in magnesium alloy. They are comparatively cheap, simple in
production, are miniature/small and do not require special drift. The
operating principle of this line consists of the following (Fig.

2.9.9). Conducting solid medium can be, for example, bars with gash/propyls. On its end/leads (and also in a series of intermediate points) are establish/installed the quartz crystal transducers of the electrical oscillations in ultrasonic (input) and ultrasonic in electrical (output/yield), which adhere on the surface of bar. The frequency characteristics of such lines are determined in essence by the characteristics of the quartz converters, which possess pronounced resonance properties at frequencies, determined by the thickness of quartz plate and which are of the order from several hundreds kilohertz to dozens megahertz [23]. Therefore ultrasonic lines realize usually a signal delay of high or intermediate frequency. Such lines have good qualitative indices. They found use, for example, in the broadband communicating system of the type "Rake" (see §3.5).
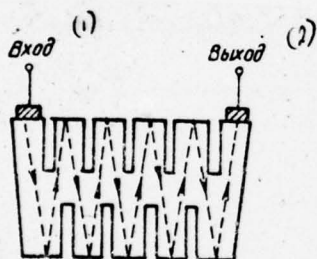
Fig. 2.9.9.

Key: (1). Input. (2). Output/yield. Page 124.

Summing and subtractors. The problem of the summing and subtractors is formation/education of sum or difference in two or several input voltage. In work with broadband signals such

equipment/devices are fulfilled with electronic or transistor
diagrams. In this case is realized simultaneously the supplementary
amplification of input signals according to voltage or power. Since
the input voltage are summarized with their signs, forming algebraic
sum, there is no fundamental difference between that which total and
subtractors. Is known at present the large number of schematics of
such equipment/devices. Let us give some examples.

As adder it is possible to utilize, for example, a directly
electron tube (transistor). In this case the summing voltages are
supplied to the different electrodes of tube, for example to control
electrode and cathode (Fig. 2.9.10a). Then output voltage is equal to
$$u_{вых} = K(u_k - u_g),$$
where K is a factor of amplification of cascade/stage, u_k and u_g
are the voltages, supplied on cathode and control electrode
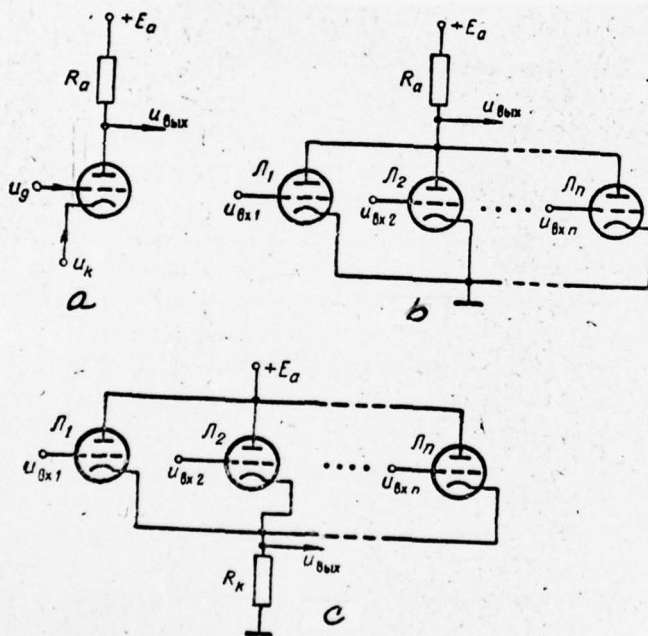respectively.

Fig. 2.9.10.

~~Page number~~ Page 125.

A deficiency/lack in this diagram in the fact that when using triodes
the entry impedance of cathode circuit is small (order 1/S, where S -
the slope/transconductance of tube) and therefore it must be supplied
from low-resistance source.

An addition of input voltage can be realized also with the aid
of the circuits of the multiple operation of a series of amplifiers
for common/general/total anode or cathode load (Fig. 2.9.10b, c).
Advantage of such diagrams in the high value of entry impedances and
the weak coupling between input circuits.

As summator it is possible to utilize an amplifier with deep
negative feedback because of resistor/resistance R_{oc} (Fig. 2.9.11).
The number of cascade/stages of amplification in diagram must be odd,
and their common/general/total amplification factor by sufficiently
large [23]. Then when selecting $R_i = R_{oc}$, **where i = 1, 2, ..., n,**
the output voltage of diagram is equal to $u_{\text{вых}} = -\sum_{i=1}^{n} u_{\text{вх}i}$,   i.e., $u_{\text{вых}}$
it differs only in terms of sign from the sum of input voltage. The
advantage of the diagrams of addition with negative feedback is

comprised in insensitivity to load changes.

In conclusion it should be noted that a selection of the type of the schematic of adder is dictated in each concrete/specific/actual case by the character of the solved problem.
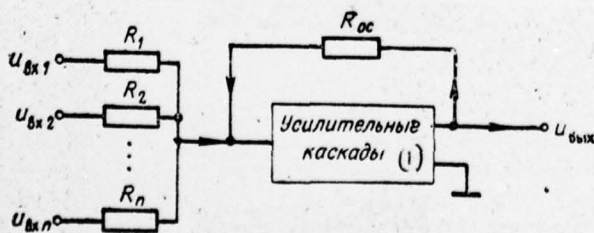

Fig. 2.9.11.

Key: (1). Amplifier stages.

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