STAT

TRANSLATION

ELECTRIC COMMUNICATIONS
(ELEKTROSVYAZ')

NO. 1, JANUARY 1957

SVYAZ'IZDAT - MOSCOW
Pages 3 - 80

PREPARED BY
TECHNICAL DOCUMENTS LIAISON OFFICE
MCLTD
WRIGHT-PATTERSON AIR FORCE BASE, OHIO

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

100

TABLE OF CONTENTS

STAT

Page

STAT

Page

[redacted]

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

10C

PARAMETERS OF BINARY CODING SYSTEMS

by

V.I.Siforov

This article introduces the concepts of the mean probability of decoding error and the optimal mean probability of decoding error at transmission of information by binary digits. Adduced are proofs that the mean probability of decoding error increases in approximately direct proportion to duration of transmission. Also introduced are the concepts of the probability of symbol decoding error, relative probability of decoding error, authenticity of coding system, and effective channel capacity. The physical sense of these parameters is described as applied to binary coding systems.

Introduction

As of the present there exists a great number of various known binary coding systems. To these pertain, especially, the error-correcting codes based in the utilization of but a part of the combinations of elementary binary-digit signals out of the total of possible combinations, and investigated by Hamming (Bibl.1, Laemmel (Bibl.2), Reed (Bibl.3), Silverman and Balser (Bibl.4), Siforov (Bibl.5), and others.

It is necessary to establish single parameters reflecting the characteristics of these coding systems in order to carry out their quantitative comparison. The present article is intended to promote the establishment of a single system of such parameters by tentatively introducing novel parameters, revealing their physical meanings and furnishing the necessary quantitative relationships between the novel and the known parameters.

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

Mean Probability of Decoding Error

Let us investigate a code combination of n elements each of which may have only two values. The sequence of positive and negative current sendings depicted in Fig.1 may serve as an example of such a code combination.

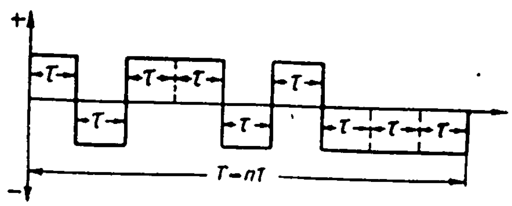


Fig.1

On designating the two possible values of each element by the symbols 0 and 1, the said code combination may be imagined as representing a sequence of zeros and unities. Thus for instance, the combination of positive and negative current sendings depicted in Fig.1 may be represented in the form of 101101000.

The total sum of all possible sequences, each containing n symbols, obviously equals $M = 2^n$. Out of this total let us select only a part, equal to $N = 2^{Rn}$, where $R < 1$, for use in a communication channel. The cumulation of the selected N sequences represents in itself a form of an alphabet of the coding system, while the individual sequences represent the letters of that alphabet.

If noise is present in the communication channel, distortions may affect any element in the code combination, i.e., as per the designations we have adopted, a zero may be transmuted into a unity and vice versa. Let us designate by p the probability of such a transformation.

The noise-induced transformation of the elements of code combinations may lead to the transformation of alphabet letters at decoding. Let us designate by y_1, y_2, $y_3, \ldots y_N$ the corresponding probabilities of such transformation or, which means the same thing, the probabilities of the error in decoding at transmission of letters numbered 1, 2, 3, ..., N.

The mean probability of decoding error will be

$$y_c = \frac{\sum_{i=1}^{N} y_i}{N}. \qquad (1)$$

If each N letter transmits an identical number A times, the mathematical expectation of the number of letters liable to transformation at decoding will equal the sum of

$$\sum_{i=1}^{N} Ay_i.$$

On dividing that sum by the total number of AN transmitted letters we obtain the value of y_c according to Bibl.1. Hence it follows that the value of y_c is approximately equal to the ratio of the number of incorrectly decoded letters to the total number of all transmitted letters provided that the said total number be sufficiently high and that all letters of the alphabet be transmitted at the same rate.

In this way, if a sufficiently high number of B letters is transmitted in a communication system, and if all these letters occur with the same frequency, the number of incorrectly decoded letters will equal $y_c B$.

The above-cited example of transmission of all letters of the alphabet the same number of times is not an optimally suitable instance. To reduce the total number of incorrectly decoded letters it is necessary to transmit less often the letters with a higher probability of incorrect decoding and more often the letters with a lower probability of incorrect decoding.

Let $p_1, p_2, p_3, \ldots p_N$ correspond to the probability of appearance of transmitted letters numbered 1, 2, 3, ..., N. Thereupon the amount of information H (entropy) transmitted through the system will be, as related to a single letter, and as is known from Bibl.6:

$$H = -\sum_{i=1}^{N} p_i \log_2 p_i. \qquad (2)$$

It can be demonstrated that at a given cumulation of the values of the magnitudes $y_1, y_2, y_3, \ldots y_N$, i.e., probabilities of transformation of alphabet letters

STAT

numbered 1, 2, 3, ... N, and at a selected constant value of entropy H there exists the optimum suitable cumulation of the values of probabilities $p_1, p_2, \dots p_N$ at which the mathematically expected of the number of incorrectly decoded letters will be at its minimum. The probability values characterizing the optimum suitable statistics of transmitted letters for a given coding system and a given entropy may be obtained from the following equations:

$$\left.\begin{array}{l} p_1 = \Phi_1(y_1, y_2, \dots y_N H) \\ p_2 = \Phi_2(y_1, y_2, \dots y_N H) \\ \dots\dots\dots\dots\dots \\ p_N = \Phi_N(y_1, y_2, \dots y_N, H) \end{array}\right\}, \qquad (3)$$

where $\Phi_1, \Phi_2, \dots \Phi_N$ stand for certain functions whose uncovery and exploration are the object of this article.

Equations (3) represent the conditions of the optimal utilization of a coding system.

Let us call as "optimal mean probability of decoding error", y_{co}, the ratio of the mathematically expected number of incorrectly decoded errors to the total number of transmitted letters at the optimal utilization of a coding system. The magnitude of y_{co} is, generally speaking, a function of probabilities $y_1, y_2, \dots y_N$ and entropy H.

The optimal mean probability of decoding error is approximately equal to the ratio of the number of incorrectly decoded letters to the total number of transmitted letters provided that that total number be sufficiently high and the requirements [eq.(3)] for an optimum utilization of the coding system be satisfied.

In this way, if a sufficiently high number of B letters is transmitted and the requirements [eq.(3)] for an optimum utilization of the coding system be satisfied, the number of the incorrectly decoded letters will equal $y_{co}B$, in a given communication system.

Relationship between Mean Probability of Decoding Error and Duration of Transmission

Let us first review the simplest example: transmission of k letters of alphabet, with the probability of decoding error being the same for any letter and denoted by y. Thereupon the probability of the correct decoding of any arbitrarily selected letter would equal 1 - y, and the probability of the correct decoding of all k transmitted letters would equal $(1 - y)^k$ in accordance with the theorem of the multiplication of probabilities.

The probability of error in the decoding of the cumulation of transmitted k letters, i.e., the probability that at least one of the transmitted k letters would be incorrect, is:

$$y_{general} = 1 - (1 - y^k), \qquad (4)$$

or

$$y_{general} = 1 - \left[1 - ky + \frac{k(k-1)}{1 \cdot 2} y^2 - \ldots\right].$$

Assuming here that the probability y is sufficiently low and thus assuming the existence of inequality $ky \ll 1$, we will have

$$y_{general} \approx ky. \qquad (5)$$

Equation (5) indicates that at sufficiently low probabilities of decoding error for each letter of the alphabet the probability of decoding error for the cumulation of these letters will be in direct proportion to the number of letters in that cumulation. In other words, it may be assumed that - all other conditions being equal - the probability of error at the transmission of any telegram is directly proportional to the duration of transmission of that telegram.

Let us now investigate the more general instance of inequalities among the probabilities of decoding errors for individual letters despite the transmission of all letters of the alphabet at the same rate.

STAT

100

Consider that a cumulation of two letters is transmitted through a communication channel, and that the related alphabet consists of N letters. Let us designate the possible first letters of the two-letter combination by $a_1, a_2, a_3, \ldots a_N$ and the possible second letters of that combination by $b_1, b_2, b_3, \ldots b_N$. Let us further assume that $y_1, y_2, y_3, \ldots y_N$ correspond to the probability of decoding error for all letters of the alphabet at the transmission of the first possible letter of the two-letter combination, while $y'_1, y'_2, y'_3, \ldots y'_N$ correspond to the probability of decoding error for all letters of the alphabet at the transmission of the second possible letter of that two-letter combination.

The probability of correct decoding at the transmission of the two-letter combination $a_i b_k$ is, in accordance with the theorem of the multiplication of probabilities, equal to the product of

$$(1-y_i)(1-y'_k),$$

where $1 - y_i$ and $1 - y'_k$ correspond to the probability of the correct decoding of the letters a_i and b_k.

The mathematical expectation of the number of correctly decoded two-letter combinations at single transmission of all the possible two-letter combinations will be expressed by the double sum

$$\sum_{k=1}^{N} \sum_{i=1}^{N} (1-y_i)(1-y'_k).$$

On dividing this expression by the total number of all the possible two-letter combinations, which obviously equals N^2, we obtain the mean probability of the correct decoding of two-letter combinations, in the form of

$$\frac{1}{N^2} \sum_{k=1}^{N} \sum_{i=1}^{N} (1-y_i)(1-y'_k).$$

The mean probability of the decoding error for two-letter combinations will be

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

$$y_c = 1 - \frac{1}{N^2}\sum_{k=1}^{N}\sum_{l=1}^{N}(1-y_l)(1-y'_k).$$

Assuming here that

$$\left.\begin{array}{l} y_l \ll 1 \\ y'_k \ll 1 \end{array}\right\},$$

we obtain

$$(1-y_l)(1-y'_k) \approx 1-(y_l+y'_k).$$

Substituting this relationship into the equation for y_c we will have

$$y_c \approx \frac{\sum_{l=1}^{N} y_l}{N} + \frac{\sum_{k=1}^{N} y'_k}{N}, \qquad (6)$$

i.e., at sufficiently low probabilities of decoding error for each letter the mean probability of decoding error for the two-letter combination equals the sum of the mean probabilities of decoding error included in this combination of letters.

The above postulate can be easily applied to multiletter combinations and also to cases of optimum coding.

It can be assumed from the above statements that the mean probability of decoding error y_c and the optimum mean probability of decoding error y_{co} for a sequence of letters increase, at sufficiently small values of these probabilities, in direct proportion to the duration of transmission.

Special Parameters of Binary Coding Systems

The assumption of the direct relationship between the mean probability of decoding error for a sequence of letters and the duration of transmission of that sequence, as postulated in the preceding Section, makes it possible to characterize a coding system by the probability of decoding error as related to a time unit, a single letter or a single symbol.

[redacted]ymbol decoding error" (β) is how we will term the ratio of the

STAT

mean probability of decoding error for a sequence of letters (y_c) to the total number of the symbols (m) comprised in that sequence.

In accordance with this definition, we may write

$$\beta = \frac{y_c}{m} \frac{1}{\text{symbol}}. \tag{7}$$

In this particular instance when the probabilities of decoding errors for divers letters are identical and equal to y, then, in accordance with eq.(5):

$$y_c = ky,$$

where k is the number of letters in the transmitted sequence, and

$$m = kn,$$

where n is the number of symbols in each letter; wherefore

$$\beta = \frac{y}{n} \frac{1}{\text{symbol}} \tag{8}$$

On knowing the probability of symbol decoding error (β) and the total number (m) of the symbols contained in the message transmitted, it is easy to determine the mean probability of decoding error for that message, as according to eq.(7)

$$y_c = m\beta. \tag{9}$$

The formula for the above-reviewed particular instance of identical probabilities of coding error for all letters will be, in accordance with eq.(8):

$$y = n\beta. \tag{10}$$

Before we pass on to determining the other specific parameters of binary coding systems let us cite the definitions of mean rate of transmission and system capacity, as furnished by the information theory:

"Mean rate of transmission of information R" is the term applied to the ratio

100

of the amount of information transmitted during a given time interval to the duration of that interval. The magnitude of R is usually expressed in binary digits per second. If the duration of elementary sending (symbol) be adopted as the time unit, the magnitude of R will be expressed in binary digits per symbol.

As introduced in the second chapter of the work cited previously, the magnitude R represents the mean rate of transmission of information expressed in binary digits per symbol. Thus, if a letter consists of n symbols, and the total number of all letters of the alphabet equals

$$N = 2^{Rn},$$

then the amount of information related to a single letter will be, in accordance with the above definition

$$\log_2 N = Rn,$$

and the amount of information relating to a single symbol will be

$$\frac{Rn}{n} = R.$$

"Channel capacity at binary transmission C" is the term applied to the largest amount of information that can be transmitted during a given time unit through a communication channel without incurring distortions, at the presence of noise, and of the possibility of selecting any coding system.

The binary coding theory includes the postulate (Bibl.7) that channel capacity at binary transmission C can be expressed by the equation

$$C = 1 + p \log_2 p + (1 - p) \log_2 (1 - p), \qquad (11)$$

where p is the probability of transformation of a single symbol.

In order to characterize the extent of utilization of a communication channel for transmission of information, it is relevant to introduce the concept of effective channel capacity η, which we will construe as referring to the ratio of the mean

STAT

rate of transmission of information R to channel capacity C, i.e.,

$$\eta = \frac{R}{C}. \tag{12}$$

Obviously, the magnitude of η ranges from 0 to 1.

The amount of information transmitted by a system H as related to a single letter, is expressed by the following equation in accordance with Bibl.6

$$H = -\sum_{i=1}^{N} p_i \log_2 p_i, \tag{13}$$

where p_i is the probability of appearance of the i^{th} letter;

N is the total number of letters of the alphabet.

The amount of information related to a single letter, i.e., the mean rate of transmission of information R, will therefore be

$$R = \frac{H}{n} = -\frac{1}{n}\sum_{i=1}^{N} p_i \log_2 p_i, \tag{14}$$

where n is the number of symbols in a single letter.

In the specific instance when the probabilities of decoding error for divers letters are identical and equal to y, we will have

$$p_i = \frac{1}{N}. \tag{15}$$

Substituting this equation into the general equation (14) we obtain

$$R = \frac{\log_2 N}{n}, \tag{16}$$

which corresponds with the equations previously cited for this specific instance.

We have already described the concept of the probability of symbol decoding error (β). It seems to us that this concept should be complemented by also characterizing a coding system by the relative probability of decoding error (α) which we will construe as referring to the ratio of the probability of symbol decoding error (β) to the mean rate of transmission of information (R), i.e.,

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

100

$$\alpha = \frac{\beta}{R} \frac{1}{\text{bit (binary digit)}} \tag{17}$$

The relative probability of decoding error α thus represents the probability of decoding error in relation to a single symbol and to a unit of the rate of transmission of information. Inasmuch as the value β has the magnitude of $\frac{1}{\text{symbol}}$ and the rate R has the magnitude of $\frac{\text{bit}}{\text{symbol}}$, the value α has the magnitude of $\frac{1}{\text{bit}}$, i.e., is expressed in reverse binary digits.

On knowing the relative probability of decoding error (α), the total number of symbols contained in the transmitted message (m), and the mean rate of transmission of information (R), it is easy to determine the mean probability of decoding error for that message (y_c) in accordance with eqs.(7) and (17), pursuant to the following equation:

$$y_c = mR\alpha. \tag{18}$$

Thus for instance, if $\alpha = 10^{-5}$ reverse binary digits, R = 0.1 binary digits per symbol, and $m = 10^4$ symbols, then $y_c = 0.01$.

As contained in eq.(18), the product of mR represents in itself the total number of binary digits contained in a transmitted telegram. Therefore, the relative probability of decoding error α represents in itself the probability of occurrence of error in the transmitted telegram as related to the number of binary digits in the message contained in that telegram.

Assuming that

$$mR = D, \tag{19}$$

where D is the total number of binary digits in the message contained in the telegram, we will obtain in accordance with eq.(18)

$$y_c = D\alpha. \tag{20}$$

STAT

This equation makes it possible to determine – on knowing the parameter α and on setting the allowable value of mean probability y_c of the occurrence of error in the telegram – the largest allowable amount of information in any telegram. Thus, for instance, if $\alpha = 10^{-6}$ reverse binary digits, and if the allowable value of mean probability of the occurrence of error in a telegram is $y_c = 10^{-3}$, i.e., if on the average the distortion of one telegram out of a thousand is deemed allowable, then the largest amount of information in the one telegram will be:

$$D = \frac{y_c}{\alpha} = \frac{10^{-3}}{10^{-6}} = 10^3 \text{ binary digit}$$

At transmission of telegrams with the amount of information exceeding 10^3 binary digits per telegram, the degree of distortion will exceed the allowable limits.

The smaller is the relative probability of decoding error (α) the greater will be the allowable amount of information in each telegram at a given rate of errors in the process of telegram reception. Therefore, the value

$$Q = \frac{1}{\alpha} \tag{21}$$

may be termed the authenticity factor of the coding system. The higher is the magnitude of Q the better is the coding system, in the sense of ensuring the transmission of the largest possible amount of information in each telegram at a given degree of reliability of reproduction of telegrams.

BIBLIOGRAPHY

1. Hamming, R.W. – Error Detecting and Error Correcting Codes, Bell System Technical Journal, Vol.29, No.2 (1950)
2. Laemmel, A.E. – Efficiency of Noise-Reducing Codes. Communication Theory Papers Read at a Symposium on "Applications of Communication Theory" held at Institite of Electrical Engineers, London, September 22nd – 26th, 1952, Butterworth Scientific Publications, 1953

3. Reed, A. - Class of Multiple-Error-Correcting Codes and the Detecting Scheme. Tech. Report No.44, Lincoln Laboratory, MIT, 9 October 1953

4. Silverman, R.A. and Balser, B. - Coding for Constant-Data-Rate Systems". Proc.,IRE, V.42, No.9 (1954); Vol.43, No.6 (1955)

5. Siforov, V.I. _ Noiseproof Feature of Systems with Error-Correcting Codes, Radiotekhnika i Elektronika, Vol.1, No.2 (1956)

6. Dolukhanov, M.P. - Introduction to the Theory of Transmission of Information along Electrical-Communication Channels, Svyazizdat Publishing House for Communications Literature, Moscow, 1955

7. Barnard, G.A. - Simple Proofs of Simple Cases of the Coding Theorem. Imperial College, London. Paper Read at 3rd Symposium on Theory of Information, 12 - 16 September 1955

Article Received by the Editors, 2 August 1956

10

A METHOD FOR APPROXIMATE COMPUTING OF THE RELATIONSHIP BETWEEN THE FREQUENCY AND TRANSFER CHARACTERISTICS OF RADIO-ENGINEERING CIRCUITS

by

S.N.Krize

This article reviews a simple method of computing transient processes according to frequency responses, and vice versa, as based on approximate calculations of the Fourier integral. The conclusion of this article is complemented with experimental data satisfactorily coinciding with the computations made by this method.

Notwithstanding the very considerable number of papers published anent the relationship between the frequency and transient responses of radio engineering circuits, there does not yet exist at present any sufficiently simple method for computing the transient response of a system as according to a given frequency response, or for plotting the frequency response as according to a known transient one. And yet, in practice, the labors of the engineers engaged in the field of radio engineering and communications very often involve the necessity of rapid and uncomplicated plotting of the transient process at the output of a system as according to its known frequency response. There also arises often a necessity for the reverse task: the determining of the frequency response of a system as according to a given transient process at its output. The technological development of electronic calculating machines makes it essentially possible to carry out such computations rapidly. But nevertheless, at present it is not yet possible to state that the numerical methods of analysis are no longer important, because electronic machines are not yet widespread enough to be used for solving all the computing tasks arising in engineering work. Therefore, the devolopment of some simple methods of numerical analysis is as

STAT

pressingly needed as ever.

The calculation of a transient process for a given characteristic response can be, as is known, effected by either of these two methods: accurately, by means of the Fourier integral; or roughly, by methods of approximation based on replacing the continuous frequency spectrum by a discrete spectrum, i.e., by replacing the Fourier integral by a series with a finite number of members. Among these approximate methods of computing frequency response one of the most fruitful is the method developed by Professor P.K.Akul'shin and based on calculating the reaction of a system to a periodic sequence of square pulses. The calculation of transient processes by the Fourier-integral method requires prior knowledge of the analytic formula for the frequency response of the system. In practice, this response is not infrequently expressed in the form of a graph and its equation is unknown. Moreover, the integration of extremely complex functions from the complex variable represented by the equations for frequency responses in uncommonly cumbersome computations which hamper the utilization of this method for solving the computing aspects of engineering work.

The utilization of the methods of approximation based on employment of the Fourier series entails difficulties connected with selecting the number of members in the dissociation of that series and summing up a great quantity of such members, which is usually necessary for obtaining reliably accurate calculations.

As described below, the method of approximate transfer-characteristic computing according to a given frequency response, and vice versa, in the analytic or graphic form, is based on replacing the determined integral by a finite sum according to the known Simpson formula. This method has the advantage of being quite satisfactorily accurate even at a small number of members in a sum ($n = 4$) as indicated by experimental data. The ultimate computing formulas do not involve the loss of much time or the employment of complex mathematical apparatus. The increased accuracy of computing stems obviously from the circumstance that at the output of the system we calculate the continuous spectrum instead of the discrete one, using for this purpose

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

an approximate expression of the Fourier integral.

Another advantage of this method is the possibility of computing the transient process at the output of a system – not only in cases of the action of a square pulse at the input of the system but also in cases of action of pulses of a more complicated form approximating the pulses encountered with in practice.

The starting formula is constituted by the Fourier integral establishing a definite relationship between the transient process at the output of a system $h(t)$, the equation for its frequency response and the spectrum of the acting pulse

$$h(t)=\frac{2}{\pi}\int\limits_0^\infty f_1(\omega)K(\omega)\cos\varphi\sin\omega t\,d\omega, \qquad (1)$$

where $K(\omega)\cos\varphi = \mathrm{Re}[K(\omega)]$ is the real part of the equation for frequency response. For brevity's sake let us introduce the designation $K(\omega)\cos\varphi = F(\omega)$; thereupon

$$h(t)=\frac{2}{\pi}\int\limits_0^\infty f_1(\omega)F(\omega)\sin\omega t\,d\omega, \qquad (2)$$

where $F(\omega)$ is the real part of the spectral function of the output signal.

For simplicity let us assume that we are dealing with a video pulse whose corresponding representative part of the function $F(\omega)$ is depicted in Fig.1. The spectrum of an r-f pulse is symmetrical in relation to the carrier frequency which is to be regarded as the beginning of the coordinates – thereupon all the below-cited results will prove correct for the pulse envelope.

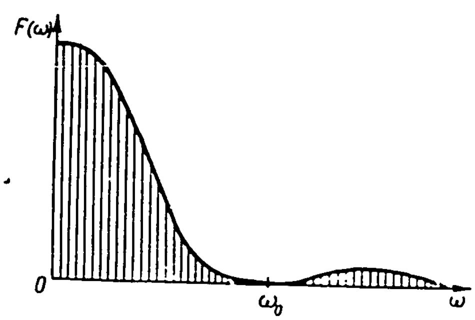


Fig.1

The approximate equations for computing definite integrals are suitable for subintegral functions with finite bounds, and therefore in eq.(2) it is necessary to change over to a frequency-bounded signal spectrum. As is known, the principal part of the energy of a signal is concentrated in the low-frequency part of its spectrum within the frequency range of 0 to $f_o = \frac{\omega_o}{2\pi}$ (Fig.1). The subsequent maximums have a low amplitude and exert a nugatory influence on the character of the determining curve, and that only in the initial periods of time. This is the foundation for assuming in actual conditions that the passband of the system is comprised within the range of 0 to f_o, where f_o is the frequency of the first zero of the spectral function $F(\omega)$. This frequency is obviously determined by the requirement:

$$\cos\varphi(\omega_0)=0. \qquad (3)$$

On changing over in eq.(2) to a finite bounded integral equal to the system passband f_o we obtain

$$h(t)=\frac{2}{\pi}\int_0^{\omega_0} f_1(\omega)F(\omega)\sin\omega t\,d\omega. \qquad (4)$$

Let us replace the integral (4) by a finite sum upon employing the Simpson formula which, notwithstanding its simplicity, yields a satisfactory accuracy at a low number of members of the sum. Let us recall that in that equation the segments of the curve are interpolated at intervals by fragments of parabola. Generally speaking, the Simpson formula has the following aspect:

$$\int_a^b f(x)\,dx=\frac{b-a}{3n}(y_0+4y_1+2y_2+\ldots+4y_{n-1}+y_n), \qquad (5)$$

where n is the number of intervals into which the frequency response is graded. We assume that $n = 4$, $a = 0$, $b = \omega_o$, $y_k = f_1(\omega_k)\,F(\omega_k)\,\sin\omega_k t$, in which connection, in this particular instance $y_4 = 0$; therefore

$$h(t)=\frac{\omega_0}{6\pi}(y_0+4y_1+2y_2+4y_3). \qquad (6)$$

When a square-fronted pulse acts upon the input of a system, its spectrum is determined by the formula

$$f(\omega)=\frac{1}{\omega};$$

STAT

accordingly

$$\frac{\omega_0}{6\pi} y_0 = \frac{f_0}{3} F_0(\omega) \lim_{\omega \to 0} \left[\frac{\sin \omega t}{\omega} \right] = \frac{f_0 t}{3};$$

$$\frac{\omega_0}{6\pi} 4y_1 = 0{,}85 F_1(\omega) \sin\left(\frac{\pi}{2} f_0 t\right),$$

$$\frac{\omega_0}{6\pi} 2y_2 = 0{,}212 F_2(\omega) \sin(\pi f_0 t),$$

$$\frac{\omega_0}{6\pi} 4y_3 = 0{,}283 F_3(\omega) \sin\left(\frac{3\pi}{2} f_0 t\right).$$

On substituting these results into eq.(6) we ultimately obtain

$$h(t) = \frac{f_0 t}{3} + 0{,}85 F_1(\omega) \sin\left(\frac{\pi}{2} f_0 t\right) + 0{,}212 F_2(\omega) \sin(\pi f_0 t) + \\ + 0{,}283 F_3(\omega) \sin\left(\frac{3\pi}{2} f_0 t\right), \qquad (7)$$

where $F_1(\omega)$, $F_2(\omega)$ and $F_3(\omega)$ are ordinates of the real part of the system's response, and f_0 is the frequency of its first zero (Fig.2).

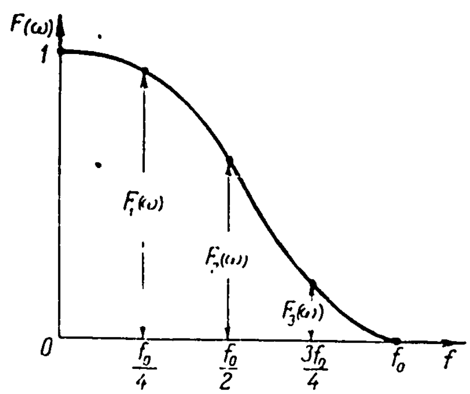


Fig.2

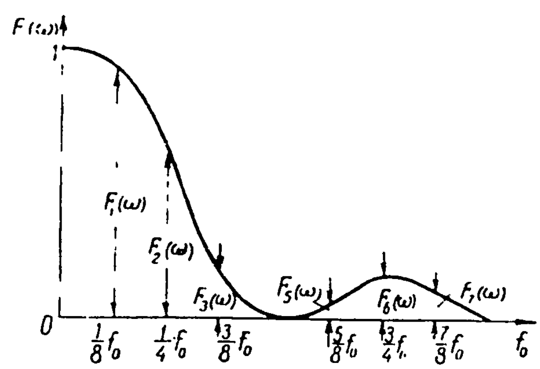


Fig.3

An evaluation of the error involved in replacing the integral by a finite sum may be effected by means of the ordinary equations employed in such cases (see, for example, Bibl.1).

The accuracy of determining the points on curve h(t) may be considerably im-

10

proved if necessary, especially at the beginning of the transient process, if the frequency f_0 be assumed as corresponding to the second zero of the frequency spectrum, and if the integration interval be broken down into eight stages (Fig.3). In this connection $y_4 = 0$ and $y_8 = 0$ and six ordinates of the frequency spectrum are subject to being determined

$$h(t) = \frac{f_0 t}{6} + 0{,}85\, F_1(\omega)\sin\left(\frac{\pi}{4} f_0 t\right) + 0{,}212\, F_2(\omega)\sin\left(\frac{\pi}{2} f_0 t\right) + 0{,}283\, F_3(\omega)\sin\left(\frac{3\pi}{4} f_0 t\right) + 0{,}17\, F_5(\omega)\sin\left(\frac{5\pi}{4} f_0 t\right) + 0{,}07\, F_6(\omega)\sin\left(\frac{3\pi}{2} f_0 t\right) + 0{,}12\, F_7(\omega)\sin\left(\frac{7\pi}{4} f_0 t\right). \quad (8)$$

The obtained results may be employed for computing the transient process also in cases when a pulse of any form is acting on the input of a system. In this connection, it is necessary to know the spectrum of that pulse in addition to the frequency response, because it can be used to plot correspondingly the pulse spectrum at the output. Thereupon, the transient process can be computed by substituting the ordinates of frequency response in eq.(7) with the corresponding ordinates of the output spectrum curve.

There are no problems to computing analogously the spectrum at the system output according to a given transient response. For this purpose, we use the reversed Fourier transform

$$U(\omega) = \int_0^\infty h(t)\cos \omega t\, dt \quad (9)$$

We will carry out the computations for the time interval from $t = 0$ to $t = t_y$, where t_y is the time of determination of the process, (Fig.4). On replacing the integral (9) by its approximate equation we will have according to eq.(5):

$$U(\omega) = \frac{t_y}{3n}[4y_1 + 2y_2 + \ldots + 4y_{n-1} + 1], \quad (10)$$

in which connection

100

$$y_0 = 0, \quad y_n = 1, \quad y_k = h_k(t)\cos\omega t_k .$$

Assuming that n = 4, which is usually wholly sufficient, we obtain

$$U(\omega) = \frac{t_y}{12}\left[4h_1(t)\cos\left(\frac{\omega t_y}{4}\right) + 2h_2(t)\cos\left(\frac{\omega t_y}{2}\right) + 4h_3(t)\cos\left(\frac{3\omega t_y}{4}\right) + 1\cos\omega t_y\right]. \qquad (11)$$

The real part of the equation for the frequency response of the system may be found from the equation

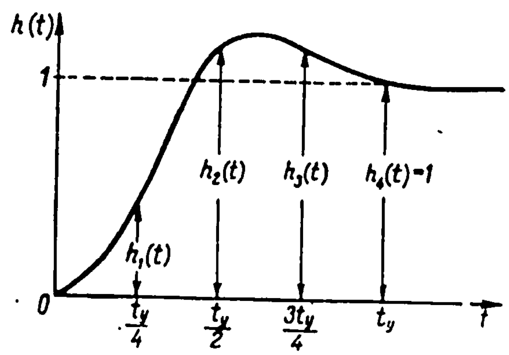


Fig.4

$$\mathrm{Re}\,[K(\omega)] = \frac{U(\omega)}{f_1(\omega)} = \omega U(\omega), \qquad (12)$$

where $f_1(\omega) = \frac{1}{\omega}$ is the spectrum of the connecting pulse.

In all methods of approximate computation special importance is attached to the verifying of the accuracy of the obtained results. Such verifying has been effected by these two means: by means of comparing the results of computations made with approximate and with exact equations (in the event that computing with exact equations entails no great difficulties), and by means of conducting an appropiate experiment. In both cases the comparison was applied to the least exact approximate equation (7) which appears to be the simplest and whose computing requires the determination of only three ordinates of the frequency response.

A comparison of the results of the computations by exact and by approximate equations has been conducted for a two-stage amplifier for which the real part of the response equation is expressed by the following equation:

$$\mathrm{Re}\,[K(\omega)] = \frac{1-\left(\frac{f}{f_0}\right)^2}{\left[1+\left(\frac{f}{f_0}\right)^2\right]^2}. \qquad (13)$$

The results of the computation according to eq.(13) are illustrated in Fig.5. On determining three ordinates of the frequency response and substituting them into eq.(7) we obtain

$$h(t)=\frac{f_0 t}{3}+0{,}71\sin\left(\frac{\pi}{2}f_0 t\right)+0{,}11\sin(\pi f_0 t)+0{,}05\sin\left(\frac{3\pi}{2}f_0 t\right). \qquad (14)$$

The points computed by means of the approximate equation (14) and also the

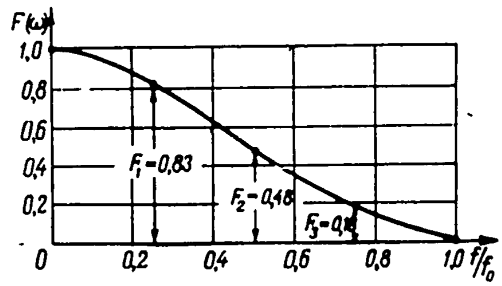


Fig.5

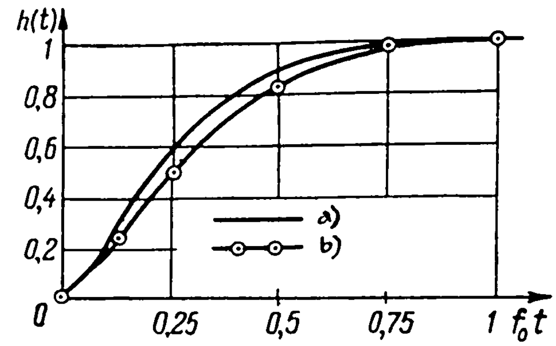


Fig.6

a) Exact formula; b) Approximate formula

transient response computed by means of the exact formula are depicted in Fig.6.

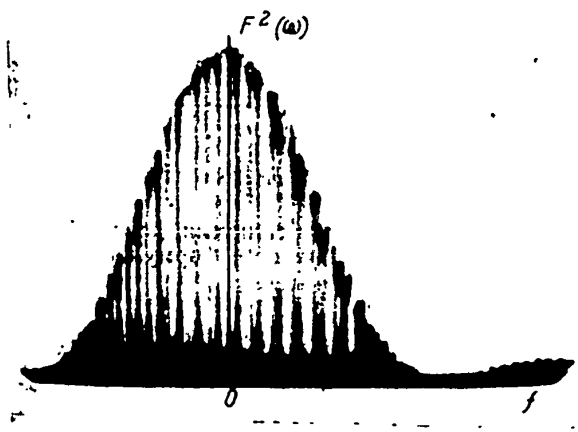


Fig.7

Fig.8

The comparison of the obtained curves attests to the satisfactory consonance

10

of results.

Figures 7 and 8 depict oscillograms of the spectrum of a high-frequency pulse and of its envelope. The computation of the envelope according to the approximate equation (7), made on the basis of a photograph of the spectrum, is indicated by dots in Fig.8. The satisfactory consonance of computed and experimental data attests to the feasibility of using the above-obtained formulas in practical work.

BIBLIOGRAPHY

1. Krylov, A.N. - Lessons in Approximate Computations, GTTI, 1954
2. Sneddon, I. - Fourier Transforms, IL, 1955
3. Zemanian, A. - Approximate Method of Evaluating Integral Transforms, Journal of Applied Physics, No.2 (1954)

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

100

PROTECTIVE ACTION AND DECOUPLING IN PERISCOPIC ANTENNA SYSTEMS

by

V.D.Kuznetsov and A.V.Sokolov

This article describes the protective action (with regard to reverse-direction reception) and decoupling in periscopic antenna systems used on radio relay lines.

The important parameters of the antenna systems used on radio lines are the magnitude of the protective action of the antenna with regard to reverse-direction reception and the degree of decoupling between antennas.

The magnitude of the protective action of an antenna with regard to rear reception determines the possibility of application of a two-frequency system for the allotment of working frequencies on a radio relay line, i.e., a system in which reception from both directions occurs at a single frequency and transmission to both directions occurs at another single frequency. In this case, only two working frequencies are necessary for organizing communication along a single r-f line. The magnitude of decoupling between the receiving and transmitting parts of the antenna system determines the extent of possible similarity between the receiving and transmitting frequencies, the requirements for filtering system, and the extent of the "echo" that can be created by overradiation of neighboring antennas.

For a line with a large number (240 or more) of channels and a considerable length (1000 and more km) it is assumed that the protective action of the antennas should be no less than 60 db at the use of a two-frequency allotment system. At a smaller number of channels or a shorter line length the requirements may be correspondingly lowered.

The decoupling between the transmitting and receiving parts of an antenna system is a good indicator of the satisfaction of these requirements at a line fre-

STAT

100

quency allotment system at which the receiving and transmitting frequencies alternate. As for the frequency allotment system at which all receiving frequencies are concentrated on one side of the working frequency range and all transmitting frequencies on another side, there the decoupling becomes less essential from the viewpoint of constructing a filter system, and the requirements as to the magnitude of that decoupling may be formulated from the viewpoint of the possible "echo" alone. As a rule it is assumed that in systems with alternate transmitting and receiving frequencies the necessary decoupling should range from 60 to 80 db.

In systems with unidirectional distribution of transmitting and receiving lines, reception and transmission are handled by single antennas in separate types of equipment. Evidently, in such systems a decoupling of the order of 40 db between the receiving and transmitting parts in one direction may be considered sufficient.

In a periscopic antenna system the magnitude of protective action with regard to rear reception and the magnitude of decoupling are fundamentally determined by the positions of reflecting mirrors relative to each other and by the design execution of the system as a whole, and in practice cannot be computed with sufficient accuracy. Therefore, in order to obtain actual data concerning this problem, measurements have been conducted on active radio relay lines with periscopic antenna system. No special measures or adjustments for increasing the decoupling were undertaken prior to the measurements.

The periscopic antenna system on which the measurements were conducted is characterized by the following design and electrical data:

1. Upper reflecting mirror - compact, flat, inclined at a 45° angle to the horizontal, representing a circle with a diameter of 3.2 m when projected on the horizontal plane.

2. Lower reflecting mirror - compact, concave surface of the ellipsoid of rotation, with foci located in the middle of the upper mirror and in the site of the antenna-feed aperture, representing a circle with a diameter of 3.2 m when projected

STAT

on the horizontal plane.

100

3. Antenna feed - a horn measuring 1 m in length and having a square aporture measuring 45 × 45 cm^2 in area, energized from a square waveguide with two symmetrical and mutually perpendicular dipoles. The one dipole is used for transmission and the other for reception.

4. The distance from the middle of the lower mirror to the feed aperture is 5 m. Minimum distance from mirrors to supporting structure is 1.5 m.

Fig.1

a) Periscopic antenna system; b) Flat mirror; c) Horn feed; d) Ellipsoid mirror

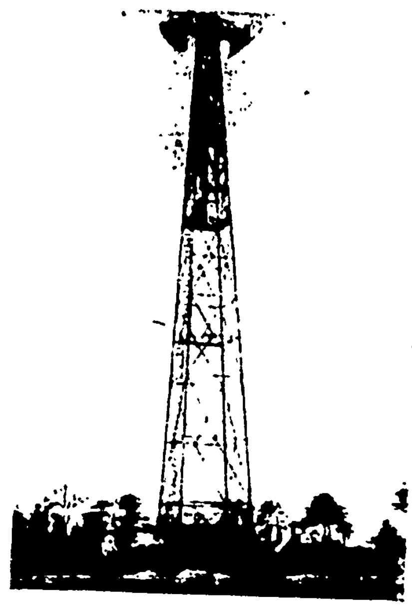


Fig.2

The upper mirror is suspended in a special attachment fastened to the top of the supporting structure. The lower mirror is mounted on a special foundation near the instrument shed. Both mirrors are fitted with attachments for their rotation in vertical and horizontal directions. The horn feed is installed inside the instrument shed and the radiation of energy occurs through a **phenoplast-covered square** window in the shed's wall. The horn also is fitted with an attachment for its rotation. Photographs of parts of the antenna are shown in Fig.1.

The gain of such an antenna system amounts roughly to 30 db, and the energy losses in the reflecting-mirror arrangement amount to about 3 db at operation in the 2000-megacycle range.

The related measurements have been carried out in a periscopic antenna system whose reflecting mirrors were located at the top and bottom of a freely standing metallic tower.

The overall view of an intermediate station with antennas of this type is depicted in Fig.2. The schematic positions of reflecting mirrors, horn feeds and instruments sheds are shown together with the principal distances in Fig.3.

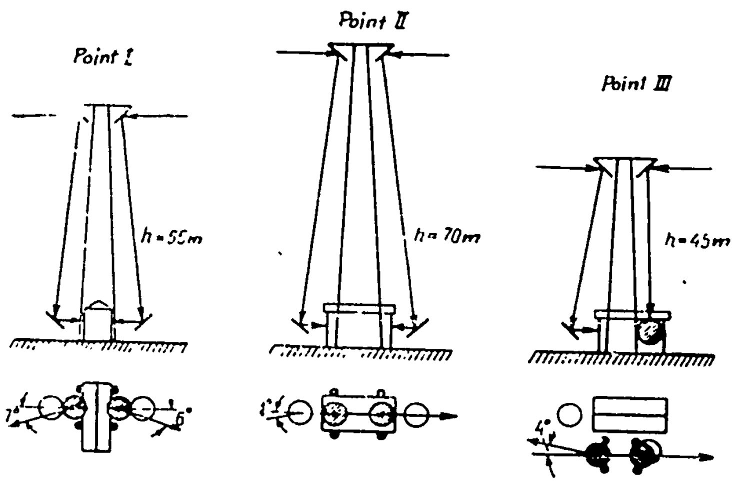


Fig.3

a) Point I; b) Point II; c) Point III

The measurements were carried out in three points with variously positioned reflecting mirrors on towers of varying height.

In point I the instrument shed is located between the tower legs right across

STAT

the line route. The spacing between the lower reflecting mirrors is 14 m. The spacing between the upper reflecting mirrors is 9.6 m. Tower height: 55 m.

10

In point II the instrument shed is likewise located between the tower's legs, but is directed along, and not across, the line route. Spacing between lower reflecting mirrors: 18.6 m. Tower height: 70 m.

In point III the instrument shed is located on a side of the tower, along the direction of the line route. The lower reflecting mirror is so positioned as to cause a polarization reversal in one direction in the reflecting-mirror system. Tower height: 45 m.

The following measurements have been carried out in the antenna systems of these points:

1. Measurement of decoupling or measurement of protective attenuations between the divers combinations of horn-feed dipoles.

2. Measurement of the protective action of the antenna at reverse-direction reception for different polarization combinations.

The measurements were conducted by means of the method of comparing the magnitudes of the received signals by means of a signal generator with a gradated output level and a measuring receiver.

These measurements led to the obtainment of the following data:

The attenuation between the different dipoles of a single feed horn range at 50 to 52 db is are not affected by the height of suspension of the upper reflecting mirrors.

The attenuation between the analogous dipoles of feed horns in different directions is in the following relationship to the height of suspension of the upper reflecting mirror:

m	db
55	68 to 70
70	52 to 55
45	78 to 80

This attenuation decreases sharply at an increase in height of suspension.

The attenuation between the opposite dipoles of feed horns in different directions is in the following relationship to the height of suspension of the upper reflecting mirror:

m	db
55	82 to 84
70	57 to 60
45	58 to 60

This attenuation is greater than the corresponding attenuations between analogous vibrators owing to the use of different polarization, but it also decreases at an increase in the height of suspension. In point III this law is reversed owing to the reversal of polarization in one of the directions.

The above cited data indicate that in such an antenna system the decoupling between the receiving and transmitting parts may be made sufficiently high.

The strength of the reverse-direction reception also depends on the height of suspension of the upper reflecting mirror, and increases with that height.

The protective action of the antenna at reverse-direction reception is in the following relationship to the height of suspension of the upper reflecting mirror – at the same polarization as that of the incoming signal:

m	db
55	50 to 52
70	40 to 42
45	53 to 55

At a differing polarization this relationship changes thus:

m	db
55	60 to 62
70	52 to 53
45	60 to 63

The difference in polarization results in the additional weakening of reception by 8 to 12 db.

The results of the measurements of reverse-direction reception indicate that it is feasible to apply the two-frequency system of working frequency allotment on radio relay lines of limited length and having a limited number of channels with similar antenna systems.

It is interesting to note that the reception of a signal from a straight direction by a dipole whose polarization does not correspond with that of the incoming signal results in a weakening by 26 to 32 db.

The obtained measurement results make it possible to formulate the following conclusions:

In periscopic antenna systems of the above-described design the employment of differing polarization for performance in opposite directions improves the noiseproof protective feature to 53 - 60 db for reverse-direction reception. This makes it possible to use the two-frequency system of frequency allotment on lines of limited length and on lines with a number of channels of the order of 60.

The high noiseproof feature obtained at low antenna supports indicates that the principal factor determining rear reception is constituted by sideward radiation of the lower reflecting mirror, striking the antenna support structure and the reflecting mirror positioned in an opposite direction. Accordingly, in the event of using a lower reflecting mirror of considerable size or in the event of using shorter waves, the noiseproof feature of the antenna system may be considerably greater.

Received by Editors on 26 August, 1956

INFLUENCE OF ANTENNA DIRECTIVITY AT LONG-DISTANCE TROPOSPHERIC PROPAGATION OF ULTRASHORT WAVES

by

V.N.Troitskiy

The relationship between the mean level of field intensity of tropospheric waves and antenna directivity is obtained here.

In long-distance tropospheric propagation of radio waves the antennas used are sometimes of a type having an extremely high gain and great directivity. In certain cases this leads to specific manifestations of a relationship between signal level and antenna directivity.

In a previous study (Bibl.1) we dealt with the problem of the magnitude of that level in cases of not very strongly directional antennas. It was demonstrated that in such cases the field intensity formed by irregularities of the l dimension will be determined by the following equation:

$$E_l^2 = 64\pi^2 \frac{p^4 M^2}{\alpha l} e^{-4\pi p \sin\varphi_0} E_0^2 \int e^{-\frac{8\pi p}{R} x} dx, \tag{1}$$

where α and M are the parameters characterizing the intensity of nonuniformities;

$p = \frac{1}{\lambda}$

λ is the wavelength;

φ_0 is one half of the angle between the tangents and to earth's surface, plotted through the points of location of antennas;

E_0 is the field intensity in free space;

R is the distance between receiving and transmitting points;

x is the height of reflecting layer, as computed from the point of intersection of tangents toward the earth's surface.

In the case of directional antennas the amplitudes of all reflected waves will

10(

vary in direct proportion to the product of the patterns of the receiving and transmitting antennas. Therefore, eq.(1) may be represented in the following form:

$$E_l^2 = 64\pi^2 \frac{p^4 M^2}{al} e^{-4\pi p \sin\varphi_0} E_0^2 \int_0^\infty f_1(\alpha) f_2(\alpha) e^{-\frac{8\pi p}{R}x} dx, \qquad (2)$$

where $f_1(\alpha)$ and $f_2(\alpha)$ are the patterns of the receiving and transmitting antennas in the vertical plane;

α is the angle of slide toward the layer located on altitude x.

Considering the smallness of the angle α, we have

$$\alpha = \frac{2x}{R}. \qquad (3)$$

The influence of antenna directivity on the magnitude of field intensity may be evaluated by employing eq.(2) upon first substituting into it a concrete form of the functions $f_1(\alpha)$ and $f_2(\alpha)$. For an approximate appraisal of this influence we will use the simplest approximation of the functions $f_1(\alpha)$ and $f_2(\alpha)$ by means of rectangles having a width of α_0, where α_0 is the angle of antenna directivity in the vertical plane.

On considering the antennas as identical, we obtain

$$E_l^2 = 64\pi^2 \frac{p^4 M^2}{al} e^{-4p \sin\varphi_0} E_0^2 \int_0^{\frac{\alpha_0 R}{4}} e^{-\frac{8\pi p}{R}x} dx. \qquad (4)$$

In this way,

$$E_l^2 = 64\pi^2 \frac{p^4 M^2}{al} e^{-4\pi p \sin\varphi_0} \left(1 - e^{-2\pi p \alpha_0}\right) E_0^2.$$

Consequently,

$$E_l^2 = 64\pi^2 \frac{p^4 M^2}{al} E_0^2 \left[e^{-4\pi p \sin\varphi_0} - e^{-4\pi p\left(\sin\varphi_0 + \frac{\alpha_0}{2}\right)}\right]. \qquad (5)$$

The first item corresponds to the case of nondirectional antennas, while the second takes into account the directivity of antennas. Both items are analogous in form, and therefore calculations of the total field may be carried out similarly as in the [redacted]tional antennas. The sole difference consists in that the

STAT

10 ultimate result will contain two items corresponding to the two items of eq.(5).

$$\bar{E}^2 = 0{,}036\frac{RF_0\lambda^{1/3}E_0^2}{\sin^{11/3}\varphi_0} - 0{,}036\frac{RF_0\lambda^{1/3}E_0^2}{\left(\sin\varphi_0+\frac{\alpha_0}{2}\right)^{11/3}}.$$

Therefore,

$$\bar{E} = 0{,}19\frac{\sqrt{RF_0}\,\lambda^{1/6}}{\sin^{11/6}\varphi_0}\left[1-\frac{1}{\left(1+\frac{\alpha_0}{2\sin\varphi_0}\right)^{11/3}}\right]^{1/2}E_0, \qquad (6)$$

where F_o is the magnitude characterizing the intensity of irregularities (the determination of that magnitude is described in Bibl.1).

Accordingly, the multiplier

$$\left[1-\frac{1}{\left(1+\frac{\alpha_0}{2\varphi_0}\right)^{11/3}}\right]^{1/2},$$

(inasmuch as φ_o is small, $\sin\varphi_o \approx \varphi_o$) determines the relationship between field intensity and antenna directivity. In this way, antenna directivity will manifest itself only when the ratio $\frac{\alpha_o}{\varphi_o}$ is sufficiently low. In practice, the influence of this directivity will be noticeable only when $\frac{o}{\varphi_o} < 2$. At a distance of R = 300 km $\varphi_o \approx 2 \times 10^{-2}$, and therefore in order to effect the above-cited inequality it is necessary that $\alpha_o < 2^o$.

At a pronounced increase of directivity in the vertical plane there will be observed a considerable decrease in the weakening coefficient $\frac{E}{E_o}$. Therefore, the contraction of the antenna pattern in a vertical plane of more definite limits is undesirable. In communication by long-distance tropospheric propagation it is advisable to employ antennas with higher directivity in the horizontal plane and lower in the vertical. It is though to be considered that an increase in the vertical-plane directivity of antennas involves some increase in the undistorted transmission band. Thus, the selection of the appropriate antenna type should be made upon considering both these factors.

STAT

BIBLIOGRAPHY

1. Troitskiy, V.N. - Propagation of Ultrashort Waves Far Beyond the Limits of the Horizon, Radiotekhnika, No.5 (1956)

Article Received by Editors on May 19, 1956

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

10

THE USSR STATE TELEVISION STANDARD 7845-55

by

S.V.Novakovskiy and D.I.Yermakov

This article reviews the principal parameters of the black-and-white television broadcasting system established by the USSR State Standard 7845 in 1955.

General Characteristics of the Standard

The Committee for Standards, Measurements and Measuring Devices confirmed on 31 December 1955 the All-Union State Standard No.7845-55 for the Principal Parameters of Television Broadcasting Systems. This Standard was drafted by the Scientific Res.Inst.of the Ministry of Communications, with the participation of a number of other interested organizations, on the basis of a 625-line television-standard proposal adopted in 1944. The cumulative experience of the years of the operation and production of transmitting and receiving television equipment has made it possible to evaluate critically the basic postulates of the 1944 proposal and to draft a revised standard which takes into account the current state of television technology in the USSR and other countries.

In connection with the plan for a further widespread development of television broadcasting during the Sixth Five-Year Period, the regulation of this field of communications by State Standards has become absolutely necessary both for the construction and operation of the network of television stations and for the conduct of intercity and international exchange of programs and for the manufacture of television equipment.

The Standard determines only the principal parameter of (black-and-white) television broadcasting systems. The parameters which do not affect such systems per se are not comprised within the Standard and should be provided for in other standards and engineering requirements in accordance with the current state of technology.

The selection of such parameters as the number of scanning lines, number of frames per second, width of radio channel, method of scanning, polarity of transmission, methods of modulation, polarization of radiation, level of black in the radiated signal, and the separation of the carrier frequencies of the image and sound transmitters is made on the basic premise of the necessity of ensuring the normal performance of the television sets already in use, without having to modify these sets in any manner.

Number of Image Scanning Lines and Clarity of Image

The State Standard regularizes the following parameters introduced in the Soviet Union as early as in 1944: number of lines - 625; nominal number of frames per second - 25, or 50 at interlaced scanning; image aspect ratio - 4:3; and maximum video-signal frequency band - 6 mc.

The fixing of the video frequency band at 6 mc, as compared with the European standard adopted by the Comite Consultatif International Radiophonique, which fixes that band at 5 mc for a line number of 625, has the advantage of making possible an increased clearness of image and a more efficient materialization of an adaptable color television system.

The discrimination of a television system determines the geometric clearness of image, i.e., the number of the little details (elements) of the image that are discernible to the naked eye and depend on the number of scanning lines Z, number of frames per second n', and frequency band Δf passed by the electrical channel. The interrelationship of these values at interlaced scanning of 2:1 is determined by the following equations:

$$n = MN, \qquad (1)$$

$$M = k_1 Z(1 - \beta), \qquad (2)$$

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

$$\Delta f = \frac{Nn'Z}{4(1-\alpha)} = \frac{nn'}{4k_1(1-\alpha)(1-\beta)}. \quad (3)$$

where M and N is the number of the discernible horizontal and vertical black and white lines, respectively;

α and β are the relative durations of the horizontal and vertical sweep fly-backs, respectively;

k_1 is a coefficient taking into account the loss of clearness along the vertical owing to the horizontal structure of the scanning pattern.

Usually $k_1(1 - \beta) \approx 0.7$ for interlaced scanning. According to the Standard, $\alpha = 0.16$ to 0.18; $\beta = 0.074$ to 0.08; and $n' = 50$.

In the specific case when vertical clearness differs from horizontal,

$$N = kM = kk_1Z(1-\beta), \quad (4)$$

$$\Delta f = kk_1Z^2n'\frac{(1-\beta)}{4(1-\alpha)}, \quad (5)$$

where k is the image aspect ratio.

Equations (1) and (2) remain applicable for this case.

Figure 1 contains a graph plotted for $N = V(Z)$ according to eq.(3) for $n' = 50$ cycles and $\alpha = 0.18$. This also contains a line plotted for equal vertical and horizontal clearness at $k = \frac{4}{3}$ in consonance with eqs.(4) and (5).

At an image brightness of 100 to 1000 apostilbs* the least angular distance φ' between divers details of a drawing consisting of black and white stripes constitutes 1.1 to 1.0 min. In the case of television image the φ' for an average eye may be assumed at 1.5 min, because the video signal at the transmission of a drawing consisting of thick black and white stripes has a form that is more sinusoidal than rectangular; therefore, in a television image this signal causes a gradual, and not instantaneous, change in brightness. The discriminating ability of the human eye

---

* 1 Apostilb = 1 lux at white light.

STAT

with respect to such lines is smaller than with respect to lines with abrupt changes in brightness. Moreover, the eye becomes fatigued quicker when observing a television image than when observing the original image, owing to pecularities of the scanning pattern (line structure, interferences, etc.), and therefore the discriminating ability of the eye is reduced.

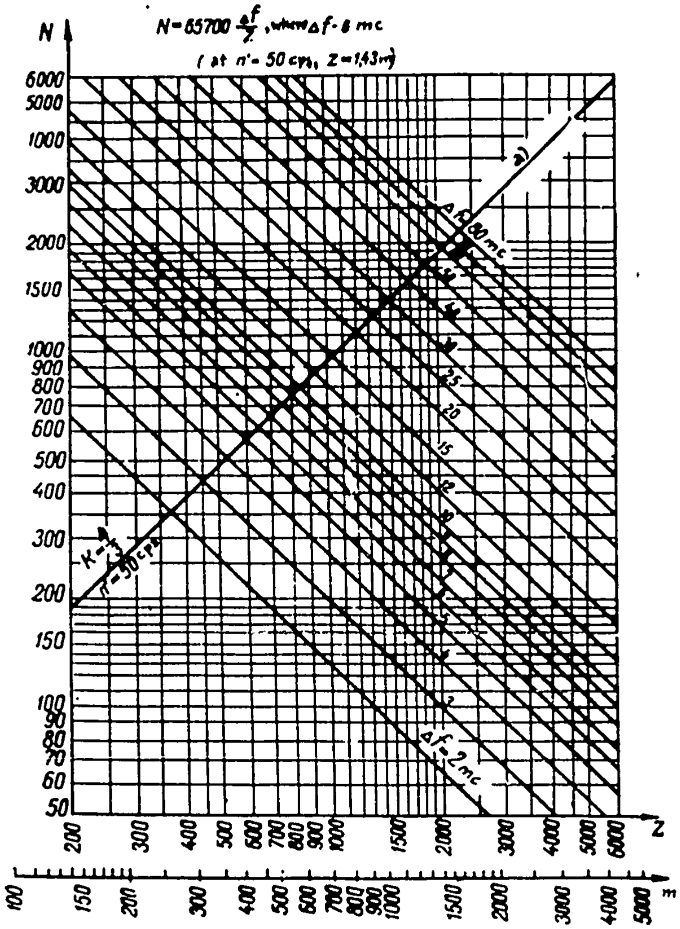


Fig.1

a) Line of equal clarity

The angular dimension of the line φ' is connected to the number of black and white horizontal lines p, the relative distance of observation ρ by the following relationship:

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

$$\varphi' = \frac{3450}{p\rho}\ min, \tag{6}$$

in which connection $\rho = \frac{d}{h}$, where d is the distance from the eye to the image and h stands for height of image.

The relationship $\varphi' = V(\rho)$ is plotted in Fig.2.

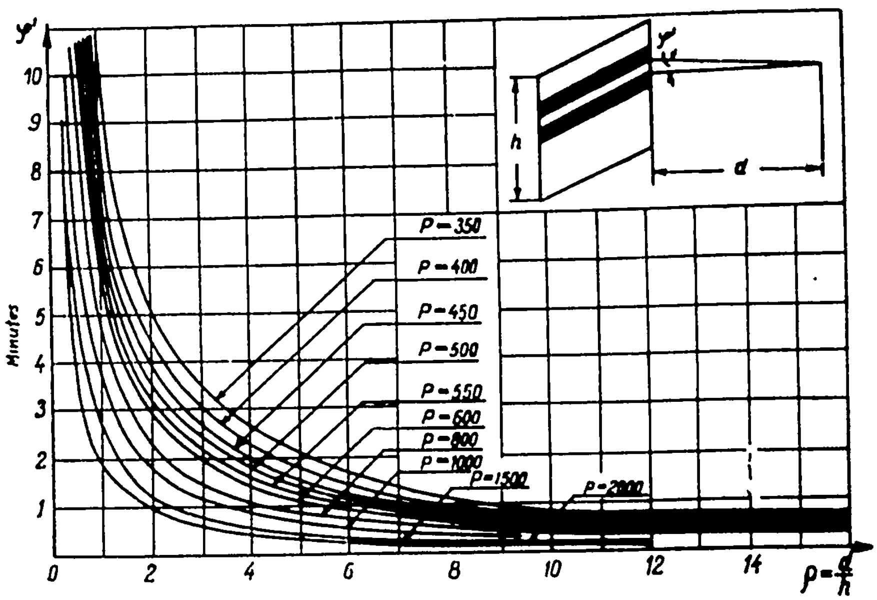


Fig.2

For comfortable viewing, $\rho = 4$ (in general).

From Fig.2 we find that at $\varphi' = 1.5$ min and $\rho = 4$, p = 575. Accordingly, M = = 575. In this instance the corresponding ratios in Fig.1 should be: Z = 825, N = = 800 (at $\Delta f = 10$ mc and n' = 50), and n = 460,000. A television system with such parameters ensures a good utilization of the discriminating ability of the human eye at $\rho \geq 4$.

For the parameters of the Soviet Standard, $\Delta f = 6$ mc, Z = 625 cycles, and n' = = 50, we find from Fig.1 that the corresponding ratios are N = 630, M = 436, whence n = 275,000.

Let us now review the question of how this Soviet television Standard ensures

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

the reproduction of geometric clearness in 35-mm movies. It is known that positive films printed on the contratype have a clearness of 25 to 30 black strokes per 1 mm of tape, i.e., 500 to 600 black lines (N') throughout the entire width of a tape frame. The connection between the number of lines N and the N' is established by this equation:

$$N = 2N'. \tag{7}$$

Accordingly, at N' = 500 it is necessary to ensure that N = 1000. The Soviet Standard provides for N = 630 (N' = 315).

However, the problem of selecting the number of image scanning lines and the width of the video signal spectrum cannot be resolved solely on the basis of data concerning the discriminating ability of human sight or the geometric clearness of a movie tape. It is necessary also to consider the economic indices of a television system and to make a compromise between the desired geometric clearness of image and the costs of the transmission system and of the television set. From this viewpoint considered, the Soviet Television Standard seems a rational one.

In effect, first of all, the Soviet Standard makes it possible to obtain an image of high quality. At transmission of program objects and not of testing tables there is but a slight improvement at an increase in the number of lines from 625 to 825 and in the frequency band from 6 to 10 mc. Secondly, the experience we have gained indicates that the costs of a television set built according to our Standard are not too high. Thirdly, at this Standard, the width of the radio channel does not exceed 8 mc, which makes it possible to use for broadcasting not only the frequencies upward of 100 mc but also the well-mastered 48 - 100 mc frequency range. Fourthly, the Soviet Standard makes it possible to utilize both radio and cable lines for interurban transmissions of television programs at reasonable economic losses.

Let us note that the following formula is a requirement for the inconspicuSTAT ness of the line structure of the scanning pattern

$$2Z_\alpha \geq \rho,$$

where Z_α is the number of active scanning pattern lines. At Z_α - 575 we have ρ = = 1150. From Fig.2 we can compute that at φ' = 1.5 min this is corresponded by $\rho \geq 2$.

Nominal Frequencies of Frames, Lines and Fields

The Standard establishes formally the nominal frame frequency of 25 cycles and field frequency of 50 cycles, which were introduced in 1944 and are applied in the existing television equipment.

The selection of precisely these frequencies in 1944 was based on the circumstance that image scanning at the time was adapted to synchronize with the urban electric networks, whose frequency in the USSR equals 50 cycles.

From the viewpoint of increasing image brightness, a field frequency of 50 cycles does not appear satisfactory, because then an increase in brightness will entail the flicker effect. However, the 50-cycle field frequency has been adopted in the existing television equipment and therefore it was not possible to alter this parameter of television systems. Special measures have to be applied in receivers in order to eliminate flicker at increased brightness.

As demonstrated by the experience gained in television broadcasting in the last years, the synchronization method at which the field frequency is synchronous with the power supply, feeding the transmitting station, has the disadvantages: a) interurban exchange of television programs is more difficult; and b) the system of synchronization in television sets does not have a high interference-killing feature.

In interurban exchange of television programs the pulse frequency should be kept as accurately constant as possible in order to preserve synchronization in television sets at rapid switching from a program from one city to a program from another city generated from a television station. Therefore, the Standard stipulates

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

100

a stability of line and field frequency amounting to ± 0.05%, which suffices to satisfy the above requirement*. At synchronization from a power-supply network such an accuracy is not ensured. To obtain a stability of ± 0.05% the pulses should evolve independently of the frequency of the network.

At such a system of synchronization the transmitting equipment and the television sets should ensure a satisfactory quality of image regardless of the frequency of the feeding network, which is another requirement set by the new Standard (assuring the reception of transmissions not synchronous with the frequencies of the feeding power networks).

At reception in the conditions of low signal-to-noise ratio it is necessary to employ synchronizing circuits with a high interference-killing feature; such circuits are those with a narrow passband and rapband. The use of such circuits is feasible only in case of a high stability of the pulse frequency, which is not achievable if the latter is synchronous with the frequency of the power-supply network.

<u>Image Aspect Ratio</u>

The Standard establishes formally the image aspect ratio of 4:3 which was introduced in the Soviet Union in 1944 and is applied in the existing equipment. Originally, the idea of transition to an aspect ratio of 11:8, such as is adopted in cimematography, was contemplated.

However, for the purpose of approximating the Soviet Standard to the television standards of other countries - which is important for facilitating international exchange of television programs, the image aspect ratio of 4:3 has been retained, as it is also observed in other countries. The resulting loss of part of image at transmission of motion pictures is small (about 3%).

* To be introduced as of 1 July 1957.

STAT

10

Form, Levels and Nominal Parameters of the Full Television Signal in the Envelope of Modulated Waves Radiated by Antenna of the Radio Transmitter

A full television signal radiated by the antenna of a radio transmitter and depicted in Fig.3 consists of image signals, quenching pulses and synchronizing pulses.

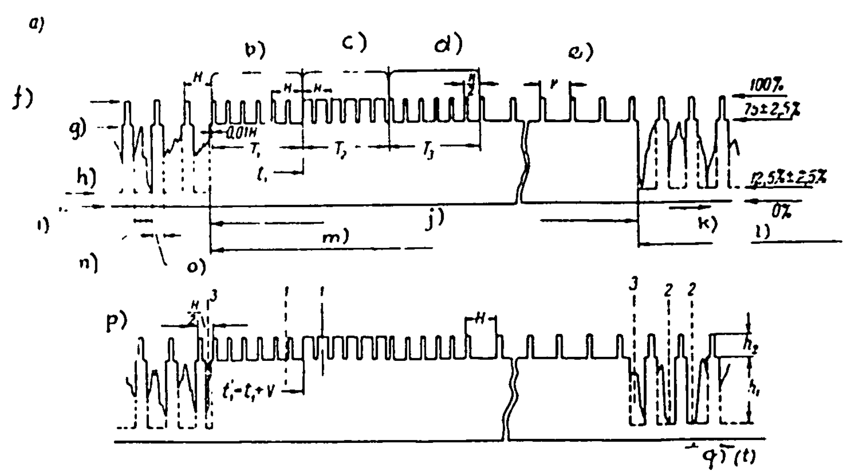


Fig.3

a) Black image scanning lines; b) Front equalizing pulses; c) Frame synchronizing pulses; d) Rear equalizing pulses; e) Line synchronizing pulses; f) Maximum level of carrier frequency voltage; g) Level of black; h) Level of white; i) Zero carrier frequency voltage; j) Frame quenching pulse; k) Time (t); l) Upper edge of image; m) Lower edge of image; n) Image signal; o) Line quenching pulse; p) Unclear image scanning lines; q) Time (t)

The parameters of the synchronizing and quenching pulses and their corresponding tolerances are cited in Figs.3, 4, 5, and 6, and in Tables 1, 2, and 3.

A study of the problem of the structure of the frame synchronization signal, conducted in connection with the drafting of the new Standard, has indicated that the change in the number of front equalizing pulses from 5 to 6 does not affect the accuracy of synchronization when integrating circuits are used. If, however, pulse

STAT

separation in television sets is effected by means of time delay or differentiation methods, there is less necessity for equalizing pulses.

The number of wide frame synchronizing pulses also may be reduced to 5, because, when integrating circuits are employed, the voltage at the output of the integrating circuits causes, when a synchro signal is fed in at the input of that circuit, the synchronization of the scanning generator at the moment of the termination of the second wide frame pulse. On this sector, the front of the integrated signal is still sufficiently steep to ensure correct interlaced scanning. Any further growth in voltage is very negligible.

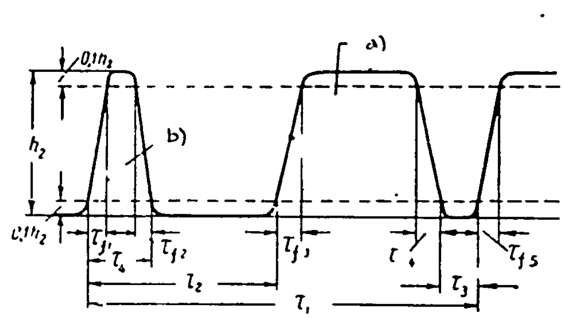


Fig. 4

a) Frame synchronizing signal;

b) Equalizing pulse

In the event of supplanting integrating circuits by time-delay methods or by differentiation, the number of wide frame pulses will be still less critical.

The introduction of rear equalizing pulses into the composition of the frame synchronization signal is, generally speaking, not obligatory, because the synchronization of the frame scanning generator within the television set occurs during a time interval preceding the appearance of these pulses. The time interval between two consequent moments of synchronization is sufficiently long (312.5 lines). Therefore, the difference in voltages at the output of the integrating circuits after an even or uneven field becomes practically nil and exerts no influence on the accuracy of synchronization during a subsequent scanning period.

According to the Standard, the number of wide pulses of the frame synchronizing signal, and also of the front and rear equalizing pulses of that signal may range at 5 to 6 in each group of the abovesaid pulses, independently.

This has absolutely no effect on receiver performance and it facilitates somewhat the employment of divers versions of synchro generator circuits.

The frame synchronizing signal adopted by the USSR Standard (Fig.3) and also by most European countries and by the USA appears to be the most complete one. This signal makes it possible to obtain an improved quality of interlaced scanning upon applying an integrating circuit, and it is distinguished by its high interference-killing feature.

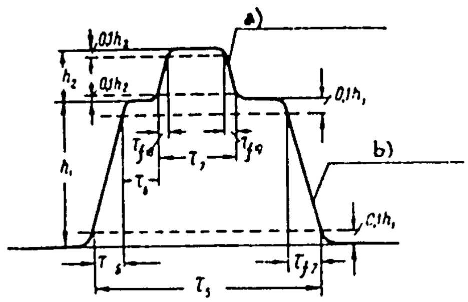


Fig.5

a) Line synchronizing pulse; b) Line quenching pulse

The recess (or front square) in front of the line synchronizing pulse and the first equalizing pulse (see Figs.3 and 5) is necessary for counteracting the influence of the contents of the transmitted image on the form of the synchronizing pulse.

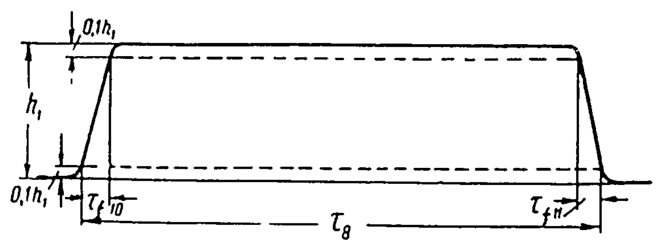


Fig.6

The moment of the appearance of the synchronizing pulse should be accompanied by the ending of all the transient processes connected with the presence of this or that signal level preceding the quenching pulse. In this way, the duration of the recess (front square) should knowingly exceed the duration of transient processes in the transmitter and receiver.

At a channel synchronization passband of 2 - 3 mc in the receiver the duration

STAT

10C

of nonstationary processes is 0.5 μ sec. To create some reserve, the width of the "recess" is set at (0.01 - 0.02)H, i.e., 0.64 - 1.28 μ sec. An excessively long duration of the "recess" might prove deleterius, because this would entail the necessity of reducing the duration of the return path of line scanning in the receiver. Moreover, this would result in a reduced interference-killing feature of line-scanning synchronization, because of the increased probability of the application of interference pulses to the recess.

As cited in Tables 1, 2 and 3, the values of tolerances for the duration of the fronts of synchronizing and quenching pulses, set by the State Standard 7845-55, are fully commensurate with the technical possibilities of the existing normal television equipment and, as demonstrated by practice, safeguard the necessary accuracy of synchronization and quality of scanning pattern.

The steeper fronts of synchronizing pulses at the transmitter's output are not expedient in view of the comparatively narrow band of the receiver's synchronization channel (2 - 3 mc). A too long duration of the front of synchronizing pulses would lead to smaller accuracy of synchronization and reduced interference-killing feature of synchronization in the receiver. A too long duration of the front and rear shape of line quenching pulses might lead to reduced fineness and straightness of the vertical sides of the scanning pattern. With respect to the frame quenching pulses, the same circumstance would lead to a noticeable instability of the moment of quenching of the upper and lower lines of the scanning pattern.

The least allowable duration (width) of synchronizing pulses is determined by the conditions of their effect on the synchronized circuit and by the properties of that circuit, and also it depends to a known degree on the amplitude of these pulses. As stipulated by the Standard (Fig.5 and Table 2), the duration of line synchronizing pulses exceeds several times the minimum necessary duration. This facilitates the conditions of the formation of pulses in synchro generator and increases the interference-killing feature of the synchronization circuits in the television set.

STAT

The frame synchronizing pulses (Fig.4 and Table 1) are designated wider than the line synchronizing ones, solely for the purpose of distinguishing them from the latter.

Table 1

a)	b)	
	in 1/10 H	in millisec
τ_1	1,0	64,0
τ_2	0,50	32,0
τ_3	0,06 – 0,08	3,84 ÷ 5,12
τ_4	0,035 0,045	2,24 – 2,88
τ_{f1}, τ_{f2}, τ_{f3}, τ_{f4}, τ_{f5}	not more than 0,004	not more than 0,26

a) Pulse; b) Duration of pulses or intervals between two pulses (τ) and duration of front and rear of pulses (τ_f)

Table 2

a)	b)	
	in 1/10 H	in millisec
τ_5	0,16 0,18	10,24 – 11,52
τ_6	0,01 – 0,02	0,64 – 1,28
τ_7	0,07 – 0,08	4,48 ÷ 5,12
τ_{f6}, τ_{f7}	not more than 0,007	not more than 0,45
τ_{f8}, τ_{f9}	" " 0,004	" " 0,26

a) Pulse; b) Duration of pulses or intervals between two pulses (τ) and duration of front and rear of pulses (τ_f)

Table 3

a)	b)		
	in 1/10 H	in 1/10 V	in millisec
τ_8	23 ÷ 25	0,074 – 0,08	1470 1600
τ_{f10}, τ_{f11}	not more than 0,1	not more than 0,00032	not more than 6,4

a) Pulse; b) Duration of pulse or interval between two pulses (τ) and duration of front and rear of pulses (τ_f)

The duration of quenching pulses is fundamentally determined by the return-
10 scanning time in the receiver and also by nonstationary processes at the end of return scanning; these processes should end within the time interval of quenching.

In a rationally designed and constructed receiver the transmitter power will be optimally used in the case when the shifting of the point of reception farther away from the transmitter is accompanied by simultaneous worsening of both synchronization and quality of image.

As indicated by practice, this requirement is met by the Standard-stipulated nominal level of black, 75 ± 2.5%, in the modulated-voltage envelope at the output of the transmitter.

Stability of the Level of Black

The Standard provides for a tolerance of ± 2.5% for the stability of the level of black in a radiated signal (deviations from the mean value of that level), regardless of the contents of image. Practice corroborates that this tolerance satisfies the requirements for a stable synchronization of the receiver.

Minimum Value of HF Voltage in the Image Signal (Level of White)

The Standard-stipulated value of the minimum voltage of the image carrier (12.5 ± 2.5%) at modulation is necessary for ensuring a satisfactory performance of the receivers using the frequency of beats between the sound and image carriers as the second intermediate sound signal frequency.

The above method of reception has found widespread application in the circuits of modern television receivers, especially in the frequency range of over 100 mc (this method makes it possible to eliminate the influence of the instability of the receiver's heterodyne frequency on the audio channel of the receiver).

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

10

Separation of Carrier Frequencies of Audio and Video Radio Transmitters

The Standard establishes this separation formally at 6.5 mc, and this rate had been introduced informally in 1944 and used since then in existing equipment.

Width of Radio Channel for Transmission of Television Programs

The width of radio channel for transmission of television programs has been set at 8 mc in 1944 and since then has always been applied in existing equipment and has been legalized by the Standard.

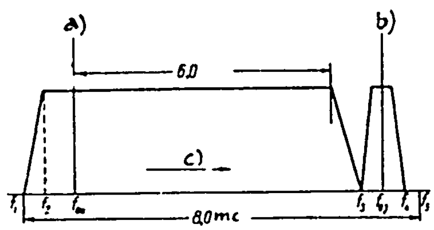


Fig.7

a) Image carrier frequency; b) Sound carrier frequency; c) Frequency

This value is a consequence of the Standard-stipulated 6.5 mc frequency separation, and of the need for reserving a 125 kc frequency band for the upper side oscillations of the audio signal transmitter, and another such frequency band for the protective zone at the upper boundary of the channel. A 1.25 mc frequency band is set aside for the part of the lower side oscillations of the video transmitter that are transmitted to the receiver, as stipulated by the Standard.

Figure 7 illustrates the nominal characteristics of the audio and video transmitters, as specified by the Standard.

The power levels for divers frequencies as measured in relation to the power level corresponding to the tops of synchro pulses are cited in Table 4.

The Power Relationship between the Image and Sound Transmitters

According to the Standard, the power of a sound signal transmitter should constitute 25 - 50% of the power of the image signal transmitter as corresponding to tops of synchro pulses.

10

Such a power ratio has been selected in view of the necessity for an identical zone of action of both transmitters at reception in normal television sets.

Table 4

a)	b)		f)
	c)	e)	
	d)	d)	
f_1 and less	20	—	$f_1 = f_{0u} - 1{,}250$ mc and less
f_2		—	$f_2 = f_{0u} - 0{,}750$ mc
f_3 and less	-	20	$f_3 = f_{0u} + 6{,}375$ mc and less
f_4	—	—	$f_4 = f_{0u} + 6{,}625$ mc
f_5 and less	20	20	$f_5 = f_{0u} + 6{,}750$ mc and less

a) Frequency; b) Attenuation of power radiated by the radio transmitters of: c) Image; d) Minimum db; e) Sound; f) Remarks

Polarization of the Electric Field of Radiated Waves

The Standard maintains the horizontal polarization of the electric field of radiated waves as introduced in 1944. At such a polarization the reflected signals and interferences from engine sparking systems are weaker than at vertical polarization, and the horizontal-plane directivity of the horizontal receiving antenna makes it possible to weaken the interferences. In this connection, the transmitting-antenna design becomes simplified.

Concurrent operation of many television stations involves the likelihood of mutual interferences among them. To weaken such interferences, the Standard permits the application of vertical polarization in specific cases, because a vertical receiving dipole weakens an interference having a horizontal polarization of the electric field of radiated waves.

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

10

Method and Polarity of Modulation of the Image Signal Transmitter

The Standard provides for the amplitude modulation with negative polarity and suppression of part of lower frequency sideband (Fig.7) that has been introduced in the USSR in 1944 for transmission of image signals.

At such a polarity (as compared with positive) there is an improved synchronization in receivers and easier designing of their automatic gain control circuits. This occurs because the synchro pulses correspond to the maximum and constant value of the radiated power. Interferences in the image appear in the form of black spots instead of the more noticeable white spots. This also facilitates the designing and construction of the transmitters, because the mean radiated power is considerably below its maximum considering that white is usually predominant in images.

Transmission of Audio Signals

A 0.25 mc channel is reserved for the transmission of audio signals, and the method used is that of frequency modulation with the maximum frequency deviations amounting to ± 50 kc. The nominal modulating frequency band for an audio transmitter should range at 30 - 15,000 cycles, and provision should be made for a rise in the modulating voltage gain at audio frequencies by means of an electric circuit with a time constant of 50 μ sec.

The FM method is retained in the Standard in view of its well-known advantages. The 30 - 15,000-cycle FM band is selected in order to obtain an opportunity for creating high-quality receivers reproducing such a band.

The maximum frequency deviation Δf_{max} has been selected so that: a) the instability of the receiver's heterodyne frequency will not lead to amplitude and phase distortions in view of the spread of the received-signal spectrum beyond the limit of the frequency band passed without distortions by the receiver along the section from input to the audio detector (for this purpose the magnitude of f_{max} should be not great); the signal-to-noise ratio at the output of the sound detector will

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

be high (for this purpose the magnitude of Δf_{max} should be high).

During the drafting of the USSR State Standard 7845-55 it was observed that ultrashort-wave FM systems tend to have their Δf_{max} at ±50 kc. In the interest of unification, the Television Standard has also adopted this value.

The rise in modulating voltage gain at high frequencies (pre-distortions), as provided for in the Standard, makes it possible to improve the signal-to-noise ratio in the receiver, because the energy in the audio signal spectrum is principally concentrated in the 150 - 2000-cycle frequency band. On the 10-kc frequency the attenuation of energy averages 10 db. Therefore, on higher frequencies the signal-to-noise ratio deteriorates.

The introduction of pre-distortions in a transmitter should be preferably accompanied by creating a corresponding drop in gain on higher frequencies in the receiver by means of a special RC circuit or a timbre regulator.

Article received by Editors on 5 June 1956

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

THE IDEAL POWER CONVERTER - A NEW ELEMENT OF ELECTRIC CIRCUITS

by

E.V.Zelyakh

There exists a newly introduced concept of a new electric circuit element represented by a four-terminal network in which the voltage ratio equals the current ratio at any load. This element, termed ideal power converter, is utilized for creating new converter circuits of nonreciprocal four-terminal networks, and for analyzing transmission in circuits with nonreciprocal four-terminal networks. The here-expounded theory is illustrated by instances of its application in, especially, transistor amplifiers.

Introduction

The theory of linear electric circuits is based chiefly on the following four elements: resistance, inductance, capacitance, and a voltage (or current) source. These two-terminal elements are utilized in designing converter circuits for practically any linear electric device.

Nonetheless, for purposes of analysis and computation, it has been found expedient to introduce into practice also the four-terminal element. This pertains to:

1) The ideal transformer (Bibl.1) (Bibl.2, p.249), described by these equations:

$$U_1 = \frac{1}{n} U_2, \quad I_1 = -nI_2, \tag{1}$$

or in matrix form

$$\begin{vmatrix} \dot{U}_1 \\ I_1 \end{vmatrix} = \begin{vmatrix} \frac{1}{n} & 0 \\ 0 & n \end{vmatrix} \times \begin{vmatrix} \dot{U}_2 \\ I_2 \end{vmatrix}. \tag{2}$$

where n is the real number. Here and later on U_1 and I_1 designate input voltage and input current, respectively, and U_2 and I_2 designate output voltage and output current, respectively, on the output of the four-terminal network (Fig.1)

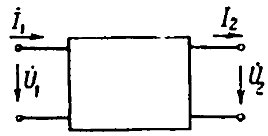


Fig.1

A generalization of this ideal transformer could be constituted by the ideal converter (Bibl.3, 4), which is described by the same eqs.(1) or (2) but whose n has a complex value.

2) The ideal gyrator (Bibl.5), described by these equations:

$$\dot{U}_1 = mI_2, \quad I_1 = \frac{1}{m}\dot{U}_2, \tag{3}$$

or in matrix form

$$\begin{vmatrix} U_1 \\ I_1 \end{vmatrix} = \begin{vmatrix} 0 & m \\ \frac{1}{m} & 0 \end{vmatrix} \times \begin{vmatrix} U_2 \\ I_2 \end{vmatrix}, \tag{4}$$

where m is the real number. Assuming that the number m is complex we will obtain yet another element which could be termed generalized ideal gyrator.

A common trait of the above-enumerated four-terminal elements is the circumstance that at any load resistance they convert voltage into current, subject to the conditions

$$\dot{U}_1 I_1 = \dot{U}_2 I_2, \quad \frac{\dot{U}_1}{I_1} \neq \frac{\dot{U}_2}{I_2}. \tag{5}$$

The first of these conditions signifies that apparent power at the input of the element remains equal to apparent power at the output of the element, at least. The second condition signifies that the input resistance of the element is not equal to its load resistance. Thus, the ideal transformer converts the load resistance into a resistance $\left(\frac{1}{n}\right)^2$ times greater by the modulus. The ideal gyrator con-

verts the load resistance into a resistance equal to the value of m^2 divided by load
10 resistance, i.e., it transforms the two terminal load network into a reciprocal two terminal network.

As early as in 1947 a paper by Kh.F.Fazylov (Bibl.6) and the subsequent discussions (Bibl.7) inspired this writer with the idea of introducing into the theory of electric circuits a new ideal four-terminal element characterized by the equations:

$$\dot{U}_1 = \frac{1}{K}\dot{U}_2, \quad I_1 = \frac{1}{K} I_2. \qquad (6)$$

or in matrix form

$$\begin{bmatrix} \dot{U}_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} \frac{1}{K} & 0 \\ 0 & \frac{1}{K} \end{bmatrix} \times \begin{bmatrix} \dot{U}_2 \\ I_2 \end{bmatrix}, \qquad (7)$$

where K is a complex number, generally speaking:

$$K = ke^{i\theta}.$$

As will be shown below, the new element converts power without converting resistance. Proceeding from this, the writer has termed that element "ideal power converter" (IPC).

At that time already, the writer applied the IPC when creating new circuits for converting the nonreciprocal four-terminal network and for analyzing the cascade connections of such networks. The obtained results were announced by the writer on 24 March 1953 at a Scientific-Technical Conference of the Leningrad Electrical Engineering Institute for Communications (LEEIC) named after M.A.Bonch-Bruyevich.

In December of 1954 there appeared an article (Bibl.8) in which an electric-circuit element coinciding with the IPC was discussed under the appellation of an "ideal amplifier-phase rotator".

In recent times this writer applied the IPC for analyzing transistor ampliSTAT, and reported the results on 6December 1955 at a seminar of the Chair of the Theory of Electro-Communications, LEEIC, and on 26 March 1956 at a Scientific-Technical Con-

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

100

ference in the LEEIC.

It was found that the IPC makes it possible to simplify the analysis of many circuits and make it more penetrating.

Therefore, the present article concerning the IPC and its possible uses is herewith recommended to the reader's attention*.

Properties of the Ideal Power Converter

The following properties of the ideal power converter can be inferred from eqs.(6) or (7):

1. The voltage at the output of the IPC equals the voltage at its input, multiplied by K. Analogously, the current at the output equals the current at the input multiplied by K. In this way, the IPC converts voltage and current identically. Let us conditionally term the number K "conversion factor".

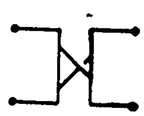


Fig.2

2. The ratio of the modulus of voltage to the modulus of current at the input of the IPC, and also the phase shift between that voltage and current, are equal to the corresponding ratio and phase shift at the output. Consequently, the input resistance of the IPC equals the load resistance of the IPC, i.e., the IPC does not convert the load resistance.

3. The apparent power at the output of the IPC equals the apparent power at its input multiplied by k^2.

4. As can be concluded from the property of invariancy of the phase shift, formulated in point 2, the active (and also the reactive) power at the output of the IPC is likewise equal to the corresponding power at the input of the IPC multiplied

* During the writing of this paper the writer learned about an article (Bibl.9) published in 1955, where results coinciding with some of the results stated here have been arrived at in a different way.

STAT

10

by k^2. In this way, the IPC appears to be an attenuator if $k < 1$, and an amplifier if $k > 1$. Considering this, we will introduce a graphical symbol (Fig.2) resembling the symbol for amplifier, to represent the IPC.

5. In a circuit constructed by cascade connection of an arbitrary number of four-terminal networks and an ideal power converter, the latter may be moved along the circuit (especially to the beginning or end of the circuit) without changing the matrix of the circuit as a whole. This may be easily demonstrated by computing the cascade connection matrices in the known manner (Bibl.2, p.213) and multiplying the matrices of the indicated four-terminal networks and the matrices of the ideal power converter

$$[a_k] = \begin{bmatrix} \frac{1}{K} & 0 \\ 0 & \frac{1}{K} \end{bmatrix}. \qquad (8)$$

Let us note that at such a multiplication the IPC's matrix may be regarded as simply the number $\frac{1}{K}$.

6. Cascade connection, in any order of sequence, of several IPC's with conversion factors $K_1, K_2, \ldots, K_n$ equivalent to one IPC with a conversion factor equal to the product of the above coefficients

$$K = \prod_{i=1}^{i=n} K_i \qquad (9)$$

As in the preceding case, the correctness of the above can be verified by multiplying the matrices of the IPC.

New Circuits for Converting the Nonreciprocal Four-Terminal Network

Let us recall that a four-terminal network is termed reciprocal if it is governed by the known principle of reversibility or reciprocality; in a case to the contrary, the network is termed nonreciprocal.

A nonreciprocal four-terminal network is, for example, an amplifier with vacuum-tube or crystalline triodes; it can be also an electromechanical device containing

STAT

a definite combination of electromagnetic and electrostatic converters (Bibl.10), the computed equivalent of certain symmetrical multiphase systems (Bibl.6, 11), and so forth.

As is known (Bibl.2, p.209), the following equations are applicable for a reciprocal four-terminal network but are not applicable for a nonreciprocal one:

$$a = \begin{vmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{vmatrix} = a_{11}a_{22} - a_{12}a_{21} = 1, \tag{10}$$

$$z_{12} = z_{21}, \quad y_{12} = y_{21}, \tag{11}$$

where the values a, z and y together with the indices stand for parameters - coefficients in the known forms of four-terminal network equations:

$$\left.\begin{aligned} \dot{U}_1 &= a_{11}\dot{U}_2 + a_{12}\dot{I}_2 \\ \dot{I}_1 &= a_{21}\dot{U}_2 + a_{22}\dot{I}_2 \end{aligned}\right\}, \tag{12}$$

$$\left.\begin{aligned} \dot{U}_1 &= z_{11}\dot{I}_1 + z_{12}\dot{I}_2 \\ \dot{U}_2 &= z_{21}\dot{I}_1 + z_{22}\dot{I}_2 \end{aligned}\right\}, \tag{13}$$

$$\left.\begin{aligned} \dot{I}_1 &= y_{11}\dot{U}_1 + y_{12}\dot{U}_2 \\ \dot{I}_2 &= y_{21}\dot{U}_1 + y_{22}\dot{U}_2 \end{aligned}\right\} \tag{14}$$

Further, it is known that a converter circuit composed of three resistances or conductances linked together in the form of a T- or P-circuit (Bibl.2, p.267) may be constructed for a reciprocal four-terminal network. As for the nonreciprocal four-terminal network, its converter circuit cannot be constructed out of either resistances or conductances alone; it is usually constructed by adding a complementary circuit element consisting of one or two subordinate voltage or current sources. The converter circuits of a nonreciprocal four-terminal network containing a single such source are depicted in Fig.3. Such circuits are widely applied in computing vacuum-tube or transistor circuits. However, the presence of current or voltage sources in these converter networks complicates greatly the computing of currents and voltages in the vacuum-tube or transistor circuits of which these networks are a part.

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

Let us demonstrate that the IPC can be used to construct converter circuits of

100

nonreciprocal four-terminal networks functioning without any current or voltage source.

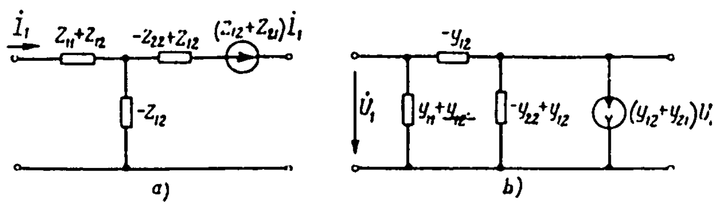


Fig.3

For this purpose, let us present the matrix [a] of a nonreciprocal four-terminal network in the form of the product of two matrices, in this manner:

$$\begin{vmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{vmatrix} = \begin{vmatrix} a_{110} & a_{120} \\ a_{210} & a_{220} \end{vmatrix} \times \begin{vmatrix} \frac{1}{K} & 0 \\ 0 & \frac{1}{K} \end{vmatrix}, \qquad (15)$$

where

$$\begin{vmatrix} a_{110} & a_{120} \\ a_{210} & a_{220} \end{vmatrix} = \begin{vmatrix} Ka_{11} & Ka_{12} \\ Ka_{21} & Ka_{22} \end{vmatrix}, \quad K = \frac{1}{\sqrt{|a|}}. \qquad (16)$$

The matrix standing in the left part of eq.(15) reflects the nonreciprocal four-terminal network. In the right part of that equation, the first matrix, which has a determinant of one, reflects some reciprocal four-terminal network, and the second matrix, which coincides with matrix (9), reflects obviously some ideal power converter.

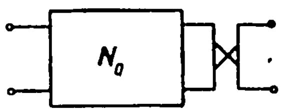
As can be concluded from the above, the linear nonreciprocal four-terminal network N is equivalent to the cascade connection (in any order of sequence) of the reciprocal four-terminal network N_o and the IPC (Fig.4) - with parameters determined according to eqs.(16)*.

* Let us note that the given equivalency loses its sense when the original nonreciprocal four-terminal network has a determinant of [a] equal to zero or infinity, i.e., when the network admits energy in one direction only.

STAT

As for the reciprocal four-terminal network N_0, let us call it derivative as distinguished from the original nonreciprocal four-terminal network N.

Upon utilizing the known rules of parameter computing it is possible to use eq.(16) to find the parameters of the other systems for the derivative four-terminal network and for the IPC:



$$\begin{vmatrix} z_{110} & a_{120} \\ z_{210} & z_{220} \end{vmatrix} = \begin{vmatrix} z_{11} & -\sqrt{-z_{12} z_{21}} \\ \sqrt{-z_{12} z_{21}} & z_{22} \end{vmatrix}$$

$$K = \sqrt{-\frac{z_{21}}{z_{12}}}, \qquad (17)$$

Fig.4

$$\begin{bmatrix} y_{110} & y_{120} \\ y_{210} & y_{220} \end{bmatrix} = \begin{bmatrix} y_{11} & \sqrt{-y_{12} y_{21}} \\ \sqrt{-y_{12} y_{21}} & y_{22} \end{bmatrix}, \quad K = \sqrt{-\frac{y_{21}}{y_{12}}}, \qquad (18)$$

where the parameters with triple indices pertain to the derivative four-terminal network and those with double indices - to the original four-terminal network.

Upon replacing the reciprocal four-terminal network in the circuit in Fig.4 by its T- or P-equivalent and considering eqs.(17) and (18) we will obtain two new converter circuits for the nonreciprocal four-terminal network (Fig.5), containing no current or voltage sources as distinguished from the circuits shown in Fig.3.

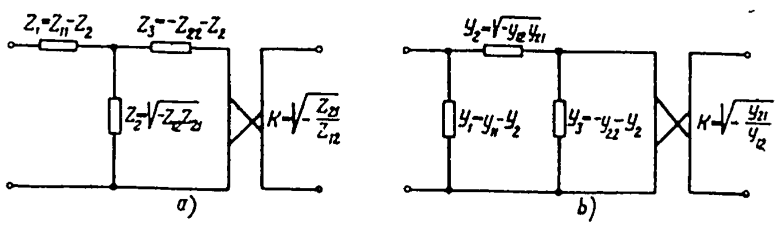


Fig.5

To illustrate the above let us cite instances of designing converter circuits for some specific devices.

Instance 1. This concerns a vacuum-tube stage with conductance between anode and grid (Fig.6a). The known circuit for converting such a stage is depicted in

Fig.6b. Now let us find a novel circuit for converting this device.

For this purpose, let us copy the matrix of the corresponding four-terminal network from the Tables cited in Bibl.2, p.255:

$$[y]=\begin{vmatrix} Y & -Y \\ Y-S & -(Y+G_i) \end{vmatrix}. \qquad (19)$$

Substituting the elements of that matrix into the formulas cited in Fig.5b, we obtain the sought-for converter circuit (Fig.6c).

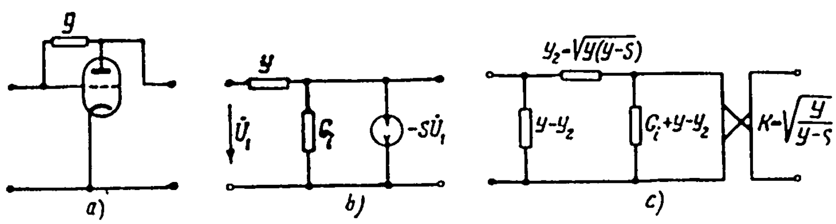


Fig.6

Instance 2. Given: a transmitter operating linearly and in one case connected to a common base (Fig.7a), in another case connected to a common emitter (Fig.8a), and in the third case, connected to a common collector (Fig.9a). Find: a new transistor converting circuit for each of these three cases, and express its values in the terms of the values of the elements of the known converter circuit for the common-base transistor depicted in Fig.7b.

This last circuit is corresponded by the matrix:

$$[z_b]=\begin{bmatrix} Z_e+Z_b & -Z_b \\ Z_b+Z_g & -(Z_b+Z_k) \end{bmatrix}. \qquad (20)$$

Substituting the elements of the above matrix into the formulas in Fig.5a, we will find the following formulas for a new converter circuit for the common-base transistor (Fig.7c)

$$\left.\begin{aligned} Z_{1б} &= Z_э + Z_б - \sqrt{Z_б(Z_б+Z_г)} \\ Z_{2б} &= \sqrt{Z_б(Z_б+Z_г)} \\ Z_{3б} &= Z_б + Z_к - \sqrt{Z_б(Z_б+Z_г)} \\ K_б &= \sqrt{\frac{Z_б+Z_г}{Z_б}} \end{aligned}\right\} \qquad (21)$$

The known converter circuit for the common-emitter transistor is depicted in Fig.8b.

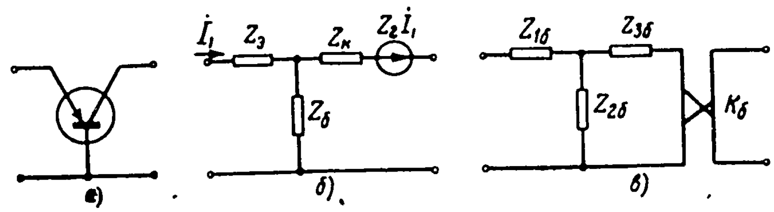


Fig.7

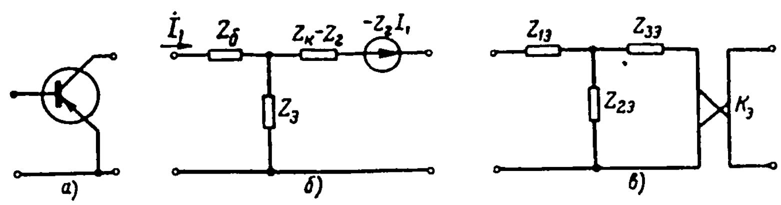


Fig.8

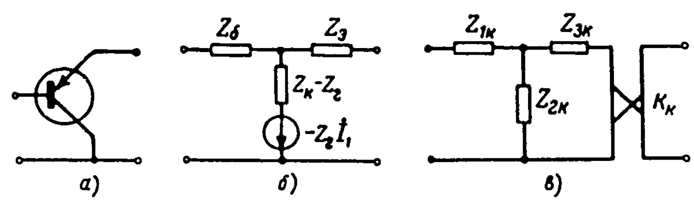


Fig.9

This circuit is corresponded by the matrix:

$$[z_e]=\begin{bmatrix} Z_e+Z_b & -Z_e \\ Z_e-Z_q & -(Z_e+Z_\kappa-Z_q) \end{bmatrix}. \tag{22}$$

Substituting the elements of the above matrix into the formulas in Fig.5a we will find the following formulas for a new converter circuit for the common-emitter transistor (Fig.8c)

$$\left.\begin{aligned} Z_{1e}&=Z_e+Z_b-\sqrt{Z_e(Z_e-Z_q)} \\ Z_{2e}&=\sqrt{Z_e(Z_e-Z_q)} \\ Z_{3e}&=Z_e+Z_\kappa-Z_q-\sqrt{Z_e(Z_e-Z_q)} \\ K_e&=\sqrt{\frac{Z_e-Z_q}{Z_e}} \end{aligned}\right\} \tag{23}$$

Finally, the known converter circuit for the common-collector transistor has the form shown in Fig.9b, and its matrix correspondingly equals

$$[z_\kappa]=\begin{bmatrix} Z_b+Z_\kappa & -(Z_\kappa-Z_q) \\ Z_\kappa & -(Z_e+Z_\kappa-Z_q) \end{bmatrix}. \tag{24}$$

Substituting the elements of the above matrix into the formulas in Fig.5a we will find the following formulas for the sought-for new converter circuit for a common-collector transistor (Fig.9c)

$$\left.\begin{aligned} Z_{1\kappa}&=Z_b+Z_\kappa-\sqrt{Z_\kappa(Z_\kappa-Z_q)}=Z_b+Z_\kappa(1-\sqrt{1-\alpha}) \\ Z_{2\kappa}&=\sqrt{Z_\kappa(Z_\kappa-Z_q)}=Z_\kappa\sqrt{1-\alpha} \\ Z_{3\kappa}&=Z_e+Z_\kappa-Z_q-\sqrt{Z_\kappa(Z_\kappa-Z_q)}=Z_e+Z_\kappa(1-\alpha-\sqrt{1-\alpha}) \\ K_\kappa&=\sqrt{\frac{Z_\kappa}{Z_\kappa-Z_q}}=\frac{1}{\sqrt{1-\alpha}} \end{aligned}\right\}, \tag{25}$$

where $\alpha = \dfrac{Z_3}{Z}$.

To indicate the order of values of the new transistor converter circuit, Table 1 cites the numerical values of that circuit for two types of transistor: a point-contact transistor with Z_e = 300 ohms, Z_b = 100 ohms, Z_k = 15 k-ohms,

Z_g = 40 k-ohms; and a junction-type transistor with Z_e = 30 ohms, Z_b = 200 ohms, Z_k = 10 m-ohms, and Z_g = 0.96 m-ohms. The data pertain to a range of sufficiently low frequencies.

Table 1

Values of the New Converter Circuit for Two Transistor Types

a)	b)	Z_1	Z_2	Z_3	K
c)	d)	—i 602	2 002	13 097	20
	e)	400—i 3 451	i 3 451	—24 700 — i 3 451	i 11,5
	f)	15 100 — i 19 365	i 19 365	—24 700 — i 19 365	— i 0,78
g)	d)	—13 628	13 858	386 340	69
	e)	230 — i 5 366	i 5 366	40 030 — i 5 366	i 179
	f)	800 000	200 000	-160 000	5

a) Type of transistor; b) Common electrode; c) Point-contact; d) Base; e) Emitter; f) Collector; g) Junction-type

Instance 3. Given: a balanced three-phase buster-transformer (Fig.10a). The resistances of the windings in each phase are equal to Z_A and Z_B respectively, and the mutual-inductance resistance equals Z_M. Find: a new circuit for converting the single-phase equivalent of the transformer.

For this purpose, we employ the previously determined (Bibl.11, p.64) matrix [y] of the given equivalent, upon copying it in the following manner:

$$[y] = \frac{1}{Z_A Z_B - Z_M^2} \begin{bmatrix} Z_A + 3Z_B & -(Z_A - i\, Z_M \sqrt{3}) \\ Z_A + i\, Z_M \sqrt{3} & -Z_A \end{bmatrix}. \qquad (26)$$

From matrix [y] we pass on to the counter matrix [z], because the latter matrix has a simpler form in the case given

STAT

100

$$[z] = \frac{1}{3}\begin{bmatrix} Z_A & -(Z_A - iZ_M\sqrt{3}) \\ Z_A + iZ_M\sqrt{3} & -(Z_A + 3Z_B) \end{bmatrix}. \qquad (27)$$

Substituting the elements of the above matrix into the formulas in Fig.5a, for the new converter circuit (Fig.10b) is valid

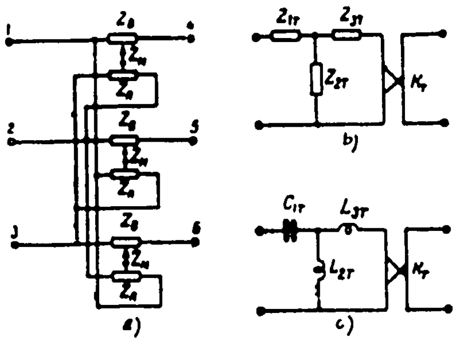


Fig.10

$$\left.\begin{aligned} Z_{1T} &= \frac{1}{3}\left(Z_A - \sqrt{Z_A^2 + 3Z_M^2}\right) \\ Z_{2T} &= \frac{1}{3}\sqrt{Z_A^2 + 3Z_M^2} \\ Z_{3T} &= \frac{1}{3}\left(Z_A + 3Z_B - \sqrt{Z_A^2 + 3Z_M^2}\right) \\ K_T &= \sqrt{\frac{Z_A + iZ_M\sqrt{3}}{Z_A - iZ_M\sqrt{3}}} \end{aligned}\right\}. \qquad (28)$$

In the particular case when the winding resistance may be ignored, the values of Z_A, Z_B, and Z_M will be imaginary:

$$Z_A = i\omega L_A, \quad Z_B = i\omega L_B, \quad Z_M = i\omega M.$$

In this connection, eqs.(29) will assume this aspect:

$$\left.\begin{aligned} Z_{1T} &= \frac{1}{3}\, i\omega\left(L_A - \sqrt{L_A^2 + 3M^2}\right) \\ Z_{2T} &= \frac{1}{3}\, i\omega\sqrt{L_A^2 + 3M^2} \\ Z_{3T} &= \frac{1}{3}\, i\omega\left(L_A + 3L_B - \sqrt{L_A^2 + 3M^2}\right) \\ K_T &= 1\, e^{i\vartheta}, \; \vartheta = \operatorname{arctg}\frac{M\sqrt{3}}{L} \end{aligned}\right\}. \qquad (29)$$

In the case given, the IPC is reduced to an ideal "element that regulates the phase alone" (Bibl.6, p.58).

For a single definite frequency ω_0 the two-terminal networks Z_{1T}, Z_{2T} and Z_{3T} may be realized in the form of capacitance C_{1T} and two inductances L_{2T} and L_{3T}, respectively (Fig.10c), which are determined by these equations:

STAT

$$\left.\begin{aligned} C_{1T} &= \frac{3}{\omega_0^2\left(\sqrt{L_A^2+3M^2}-L_A\right)} \\ L_{2T} &= \frac{1}{3}\sqrt{L_A^2+3M^2} \\ L_{3T} &= \frac{1}{3}\left(L_A+3L_B-\sqrt{L_A^2+3M^2}\right) \end{aligned}\right\}. \qquad (30)$$

Application of the IPC for Computing Transmission Along Circuits with Nonreciprocal Four-Terminal Networks

The above-obtained equivalent of the nonreciprocal four-terminal network makes it possible to reduce the computing of transmission along circuits with such networks to the computing of transmission along the more simple circuits with reciprocal four-terminal networks.

Let us first review a circuit with a single nonreciprocal four-terminal network (Fig.11a). In this circuit, let us replace the above network by its IPC-containing equivalent. Further, considering that the IPC does not convert resistances, we will conduct the computation according to the scheme in Fig.11b where the IPC is left out and a broken line indicates its position.

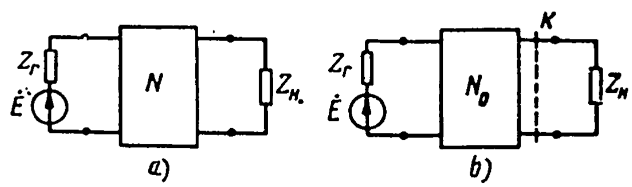


Fig.11

It is not difficult to see that the sought-for voltage and current in the circuit in Fig.11a will be the same at the input, but will be K times greater at the output than in the circuit in Fig.11b. Correspondingly, the input resistances of the four-terminal networks in the both circuits will coincide.

At a reversed direction of transmission, when the e.m.f. in the circuit in Fig.11a is transferred to the right-hand end, the computing may, as before, be conducted according to the circuit in Fig.11b, upon likewise transferring the e.m.f. there to the right-hand end. In this connection, the sought-for voltage and current

STAT

in right-hand end of the original circuit will be the same, but in the left-hand end will be K times smaller than in the circuit with the derivative four-terminal network. As for the output resistances of the four-terminal networks, they will obviously coincide in both circuits.

As can be further concluded from the above, the characteristic resistances of the said four-terminal networks will likewise coincide. With respect to the characteristic constant transmission (from the left to the right) of the nonreciprocal four-terminal network, which, as is known, is determined at a matched load by the following equation:

$$g_{c1} = \frac{1}{2} \ln \frac{\dot{U}_1 I_1}{\dot{U}_2 I_2},$$

for such a value, we will obtain the following equation:

$$g_{c1} = \frac{1}{2} \ln \frac{\dot{U}_{10} I_{10}}{\dot{U}_{20} I_{20}} \frac{1}{K^2} = g_{c0} - \ln K, \qquad (31)$$

where the values with the 0 index pertain to the derivative reciprocal four-terminal network.

In the same way we will find that the characteristic transmission constant for reverse direction of transmission will be determined by this formula:

$$g_{c2} = g_{c0} + \ln K. \qquad (32)$$

It is easy to see that the magnitude g_{c0} stands for the arithmetic mean of the values of g_{c1} and g_{c2}.

Analogously, it can be demonstrated that the iterative resistances of the nonreciprocal four-terminal network coincide with the iterative resistances of the derivative reciprocal four-terminal network, and the iterative transmission constants g_{k1} and g_{k2} of the nonreciprocal four-terminal network are determined by these formulas:

$$g_{k1} = g_{k0} - \ln K, \quad g_{k2} = g_{k0} + \ln K, \qquad (33)$$

100

where g_{k0} is the iterative transmission constant of the derivative reciprocal four-terminal network. The magnitude g_{ko} is thus the arithmetic mean of the values of g_{k1} and g_{k2}.

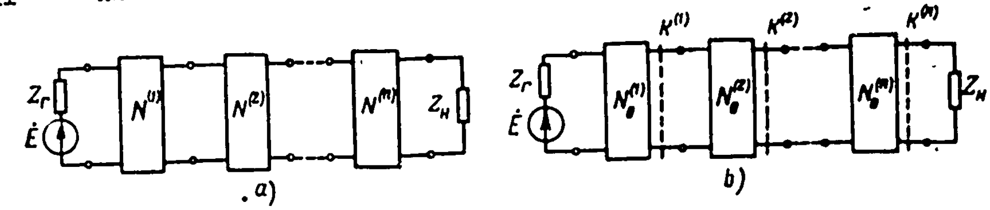


Fig.12

Let us now pass over to studying transmission along a circuit with several cascade-connected nonreciprocal four-terminal networks (Fig.12a).

As in the preceding example, in such a circuit we will replace each nonreciprocal four-terminal networks, $N^{(1)}$, $N^{(2)}$, ... , $N^{(n)}$ by their corresponding derivative reciprocal four-terminal networks $N_o^{(1)}$, $N_o^{(2)}$, ..., $N_o^{(n)}$ jointly with an IPC. Upon omitting the IPC we will obtain the circuit depicted in Fig.12b where the location of the IPC is indicated by a broken line.

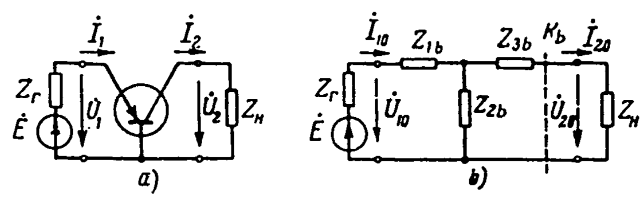


Fig.13

According to that circuit, the known methods of the circuit theory of passive circuits may be utilized for carrying out a computation of the currents and voltages at the input and output of any of the circuit's four-terminal networks. Thereupon it is possible to find the analogous values for the original circuit in Fig.12a, for which purpose the value of current or voltage at a given point in the circuit in Fig.12b has to be multiplied by the product of the conversion factors of all IPC's to

STAT

100

the left of that point.

The input resistances of the four-terminal networks are obviously identical in both circuits.

The same circuit in Fig.12b can also be used as the basis on which to compute voltages and currents in any point of the circuit in the event of transmission in return direction, and to transfer the e.m.f. to the right-hand end of the overall circuit. In this connection, in order to determine the magnitude of voltage or current at some given point in the original circuit, it is necessary to divide the corresponding magnitude of voltage or current in the derivative circuit by the product of the conversion factors of all IPC's lying to the right of that point.

As for the output resistances of the four-terminal networks, they are, like the input resistances, identical in both circuits.

Instance 4. Let us study the circuit of transmission through a common-base transistor as an example of a circuit with a single nonreciprocal four-terminal network (Fig.13a).

On replacing the transistor by its converter circuit as according to Fig.7c and omitting the IPC, we will obtain the circuit shown in Fig.13b, where the broken line designates the position of the IPC.

By looking at that circuit we will find directly the input resistance Z_{Vkh}, the current amplification factor K_i, and the voltage amplification factor K_u.

$$Z_{in} = \frac{\dot{U}_1}{I_1} = \frac{\dot{U}_{10}}{I_{10}} = Z_{1b} + \frac{Z_{2b}(Z_{3b} + Z_H)}{Z_{2b} + Z_{3b} + Z_H}, \tag{34}$$

$$K_i = \frac{I_2}{I_1} = \frac{K_b I_{20}}{I_{10}} = \frac{K_b Z_{2b}}{Z_{2b} + Z_{3b} + Z_H}, \tag{35}$$

$$K_u = \frac{\dot{U}_2}{\dot{U}_1} = \frac{K_b \dot{U}_{20}}{\dot{U}_{10}} = \frac{K_b Z_{2b}}{Z_{2b} + Z_{3b} + Z_H} \frac{Z_H}{Z_{in}}, \tag{36}$$

or, on taking eq.(21) into account

STAT

$$Z_{in} = Z_e + Z_b - \frac{Z_b(Z_b + Z_g)}{Z_b + Z_\kappa + Z_H}, \tag{37}$$

10C

$$K_i = \frac{Z_b + Z_q}{Z_b + Z_\kappa + Z_H}, \tag{38}$$

$$K_u = \frac{Z_H (Z_b + Z_q)}{(Z_e + Z_b)(Z_b + Z_\kappa + Z_H) - Z_b (Z_b + Z_q)}. \tag{39}$$

We can find just as simply the analogous values for transmission in return direction (from the right to the left) by transferring the e.m.f. to the right-hand end in the circuit in Fig.13b:

$$Z'_{in} = \frac{\dot{U}'_1}{I'_1} = \frac{\dot{U}'_{10}}{I'_{10}} = Z_{3b} + \frac{Z_{2b}(Z_{1b} + Z_G)}{Z_{1b} + Z_{2b} + Z_G}, \tag{40}$$

$$K'_i = \frac{I'_2}{I'_1} = \frac{1}{K_b} \frac{I'_{20}}{I'_{10}} = \frac{1}{K_b} \frac{Z_{2b}}{Z_{1b} + Z_{2b} + Z_G}, \tag{41}$$

$$K'_u = \frac{\dot{U}'_2}{\dot{U}'_1} = \frac{1}{K_b} \frac{\dot{U}'_{20}}{\dot{U}'_{10}} = \frac{1}{K_b} \frac{Z_{2b}}{Z_{1b} + Z_{2b} + Z_G} \frac{Z_G}{Z'_{in}}, \tag{42}$$

or on the basis of eq.(21)

$$Z'_{in} = Z_b + Z_\kappa - \frac{Z_b (Z_b + Z_q)}{Z_e + Z_b + Z_G}, \tag{43}$$

$$K'_i = \frac{Z_b}{Z_e + Z_b + Z_G}, \tag{44}$$

$$K'_u = \frac{Z_b Z_G}{(Z_b + Z_\kappa)(Z_e + Z_b + Z_G) - Z_b (Z_b + Z_q)}. \tag{45}$$

Equations (34) to (36) and (40) to (42) may also be used for computing transmission through a common-emitter or common-collector transistor; for this purpose it suffices to replace the index b in these equations by the corresponding index e or or c.

Instance 5. The circuit of transmission through a two-stage transistor amplifier (Fig.14a) may serve as an example of a circuit with several nonreciprocal four-terminal networks.

Let us replace each transistor in that circuit with a converter circuit as according to Fig.8c and let us omit the IPC, and we will thereupon obtain the circuit depicted in Fig.14b*. (For footnote, see next page) The subsequent compu-

STAT

tations of that circuit may be then conducted by means of the known methods for computing passive circuits.

Let us note that the usually applied methods of analyzing transistor amplifiers by means of the known conversion circuits containing sources of current or voltage

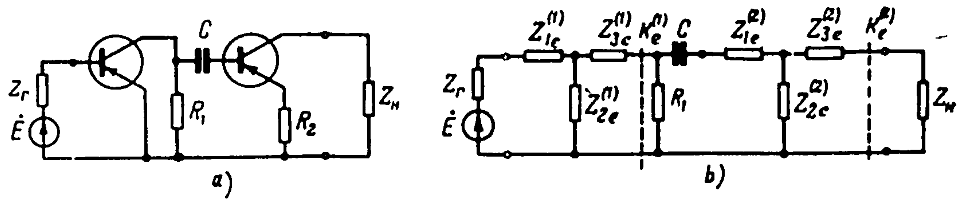


Fig.14

have to be based on Kirchhoff's law and on the derivation of sets of equations that will, moreover, differ depending on the right or left direction of transmission. As compared with the whole of such methods, the here-expounded procedure for analyzing transistor amplifiers appears to be, in our opinion, simpler and more graphic.

Conclusion

The theory expounded here may prove useful not only in the above-described simplifying of the analysis of circuits with nonreciprocal four-terminal networks but also in the modeling of circuits. Thus for instance, at the modeling of a balanced three-phase circuit consisting of cascade-connected multi-terminal networks containing also a booster-transformer, the said transformer can be modeled by means of the circuit depicted in Fig.10c; in this connection, in accordance with the foregoing explanations, the IPC may be omitted.

* Let us note that, in accordance with Fig.14a, $Z_e + R_2$ should be taken instead of Z_e when computing the values of $Z_{1e}^{(1)}$, $Z_{2e}^{(1)}$, $Z_{3e}^{(2)}$, and $K_e^{(2)}$ according to eqs.(23).

STAT

BIBLIOGRAPHY

1. Campbell, G.A. and Foster, R.M. - Maximum Output Networks for Telephone Sub-stations and Repeater Circuits, Trans of the AIEE, Vol.39, 1920

2. Zelyakh, E.V. - Foundations of the General Theory of Linear Electric Circuits, Publishing House of the USSR Academy of Scienees, 1951

3. Feige, A. - General Theory of Four-Terminal Networks, TFT, No.9, 1927

4. Yuryev, M.Yu. - Theory of Telephone Transmission and Its Practical Applications, Svyaz'tehkizdat Publishing House for Technical Literature on Communications, 1932, p.95

5. Tellegen, B.D.H. - The Gyrator, a New Electric Network Element, Philips Research Reports, Vol.3, 1948, pp.81-101; Vol.5, 1950, pp.81-86

6. Fazylov, Kh.F. - The General Method of Flow Distribution in Complex Circuits, Elektrichestvo, No.11, 1946

7. - Discussions of Kh.F.Fazylov's Article, Elektrichesvvo, No.6, 1947

8. Haus, H.A. - Equivalent Circuit for a Passive Nonreciprocal Network, J. of Appl. Phys., Vol.25, No.12, 1954

9. Aurell, C.G. - Representation of the General Linear Fourterminal Network and Some of Its Properties, Ericsson Technics, No.1, 1955

10. - Furduyev, V.V. - Reciprocality Theorems, Gostekhizdat State Technical Publishing House, 1948

11. Zelyakh, E.V. - Concerning the Theory of Balanced Multiphase System, Elektrichestvo, No.4, 1953

STAT

100

DISTORTIONS OF TELEGRAPH PULSES IN TONAL TELEGRAPHY CHANNELS AT SHARP FLUCTUATIONS OF THE SIGNAL LEVEL

by

V.N.Amarantov

This article compares the distortions of telegraph pulses owing to abrupt fluctuations in the signal level of AM and FM voice-frequency carrier telegraphy channels. It is demonstrated that the distortions in the FM channels are considerably smaller than those in the AM channels.

Introduction

Abrupt fluctuations in the signal level in long-distance wire communication channels stem from a variety of reasons. A basic reason for such fluctuations is momentary breakdowns of contact connections of conductors in equipment, on lines and in input devices. Considering that such breakdowns of connections are, as a rule, a consequence of concealed defects of parts and assembling, their detection entails major difficulties. The other reasons for the fluctuations in the signal level could be the accidental touching of wires, phenomena caused by atmospheric electricity (so far as overhead lines are concerned), commutation in repeater and terminal equipment circuits, momentary overloading of group repeaters by peak speaking currents, and the like.

The statistical data obtained from the observations of cable (coaxial and balanced-line) lines conducted in certain West European countries indicate that on the average in every 24-hour period there occur two or three fluctuations of over 0.7 - 1.0 nepers in the signal level per 100 km of line length (according to reports of the Comite Consultatif International Telephonique). In a number of cases the observed number of fluctuations was much higher than that.

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

These sudden fluctuations in the signal level are particularly harmful in tonal
100 telegraphy channels where they may lead to distortions in the transmitted texts. Let us now describe the effect of these fluctuations in the AM and FM tonal telegraphy channels.

AM Channels

To reduce the influence of fluctuations in signal level in the older AM tonal telegraphy channels, use is made of AGC devices whose performance is based on changes in vacuum-tube grid bias. Change in the bias is effected by modifying the grid current as depending on the signal level. In order to prevent the AGC circuit from reacting to signal-level fluctuations during the modulation, the action of that circuit is retarded by charging or discharging the capacitor. The duration of the retardation is timed to be much longer than the duration of the elementary pulse.

Such circuits reduce effectively the pulse distortions that occur at small fluctuations in the signal level, but they do not eliminate the distortions occurring at sharp fluctuations; the series of pulses occurring immediately after the moment of the sharp fluctuation becomes distorted. It can be demonstrated* that the maximum value of distortions will in this case be determined by the following approximate equation:

$$\delta \approx \left[\frac{1-e^{-\Delta p}}{2\Delta F}\left(1+\frac{1}{a}\right)+\frac{2}{3}\Delta p t_{dv}\right] v \cdot 100\%, \qquad (1)$$

where Δp is the fluctuation of signal level, in nepers;

v is the velocity of telegraphing, in bauds;

ΔF is the effective channel width;

a is the ratio of steadied value of current in receiving-relay winding to the value of the sensitivity current of that relay

t_{dv} is the time of motion of the relay's reed in the conditions of the given

---

* See Appendix 1

circuit, at a normal signal level.

The channels equipped with such AGC circuits are particularly sensitive to a decrease in the signal level, and the receiving relay will fail if $\Delta p > \ln \dfrac{2a}{1 + a}$ so that one or several pulses will get lost.

Figure 1 depicts the rating characteristic of maximum pulse distortions as depending on the extent of fluctuation in the signal level, at the following conditions: ΔF = 100 cycles; v = 50 bauds; a = 5; t_{dv} = 4.0 milliseconds. As can be seen from Fig.1, the distortions present at Δp = - 0.2 nepers will reach 10%, and at Δp = - 0.5 nepers the pulses will vanish.

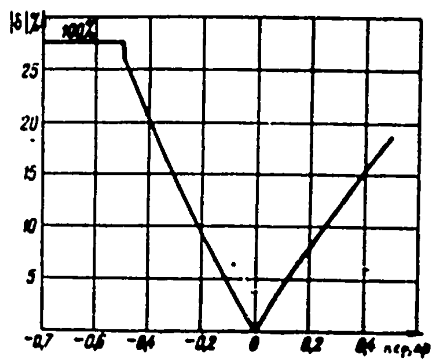


Fig.1

In the last few years there have appeared several new v - f telegraphy AM systems operating on a different method for reducing the effect of fluctuations in the signal level. Thus, Fig.2 depicts a simplified circuit of the receiving device

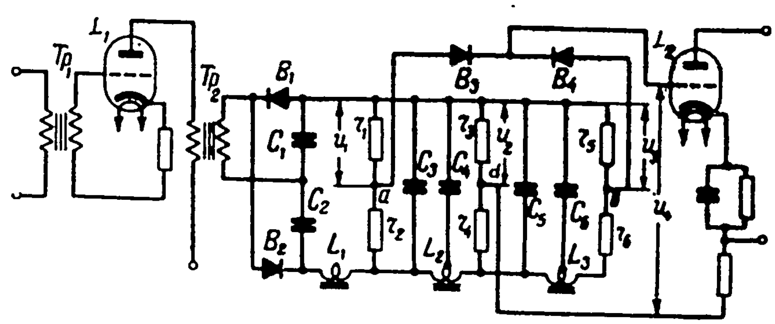


Fig.2

for a v - f telegraphy channel of an apparatus of the BTR50000 type manufactured by the Philips Company (Bibl.1). This circuit performs as follows: A signal is amplified by input stage rectified by rectifiers B_1 and B_2, filtered by filter $L_1 - C_3$, and thereupon passed through circuits $L_2 - C_4 - C_5$ and $L_3 - C_6$. Each of these two circuits delays the rectified signal by one half the time of its establishment $\frac{\pi}{2}$. The rectified pulse is taken from points a, b, and c, and transmitted to the grid of the first tube L_2 of the trigger circuit. If $u_1 < u_3$, then $u_4 = u_2 - u_1$; but if $u_1 < u_3$, then $u_4 = u_2 - u_3$, which is achieved by means of the counter connection of rectifiers B_3 and B_4. At a steady state the voltage of u_2 is twice as high as each of the voltages of u_1 or u_3. The pulse being transmitted through the channel, and the voltages u_1, u_2 and u_3, which arise during that process, are represented by curves a, b, c, and d (Fig.3). The total voltage u_4 is represented by curve e. The trigger circuit is set so that it will fire when the voltage of u_4 (curve e) changes its sign, i.e., during the time moments t_1 and t_2. Obviously, $t_2 - t_1 = T$.

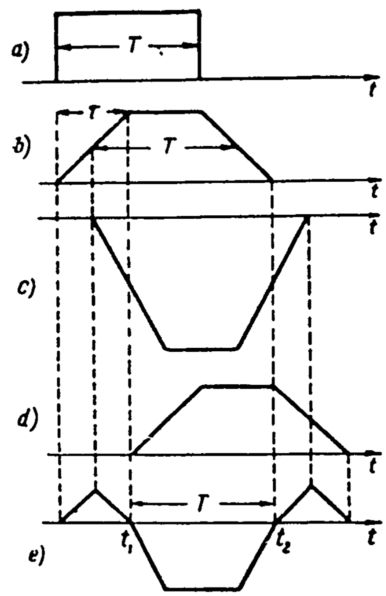


Fig.3

A change in the signal level will entail a directly proportional change in the ordinates of all curves b, c, d, and e, but the distance between the zero values of the curve e will remain the same. In this way, the duration of pulses at the output of the trigger circuit is not affected by the signal level.

Such a presentation of the circuit's performance appears to be theoretical. Actually, a trigger circuit cannot be adjusted for firing at $u_4 = 0$, because then the circuit will tend to misfire. Also it can be easily seen that if the signal voltage changes more than thrice during the transmission of a pause then the trigger circuit will likewise tend to misfire. The prospectus issued by the Philips Company about this apparatus cites the following data concerning pulse distortions at smooth and

STAT

sharp fluctuations of the signal level: at a decrease in the signal level by 0.7 neper below the nominal value, and at ν = 50 bauds, pulse distortions will total less than 18%; and when the level is decreased by more than 0.9 - 1.0 neper , the distortions will total a much higher percentage.

The WT 52/54 apparatus manufactured by the Siemens Halske Company (Bibl.2) operates on a different circuit but, like the above-described circuit, this one is also based on the totaling of rectified pulses one of which is delayed in relation to the other. In the article cited in Bibl.2 this apparatus is described as having the following specifications: at signal-level fluctuations from +0.3 to 0.7 neper or from 0.7 to +0.3 neper (in relation to the nominal value of the signal level) pulse distortions will not exceed 25%, however, at any further increase in the extent of the fluctuations, especially so far as the minus sign is concerned, the distortions will increase sharply.

FM Channels

In the last few years, FM voice-frequency telegraphy systems have found as wide an application as the AM ones. Let us review the behavior of these FM systems at fluctuations in the signal level.

The receiving device of an FM v - f telegraphy channel contains a peak limiter whose performance may be on a practically nonstorage basis; therefore, it should seem that the FM channel would not introduce pulse distortions at fluctuations in the signal level. However, the distortions do take place in this case too. Curve 1, Fig.4, gives the measured characteristic of the maximum pulse distortions in a channel of a v - f telegraphy FM apparatus of the TT 12/16 type at ν = 50 bauds. The measurements were conducted in the following manner: The fluctuation in the signal level was effected by connecting a resistance parallel to the channel path from the transmitting to the receiving filters. To block any influence that the connection of the resistance might exert on the performance of the filters, the points of con-

STAT

nection of the resistance were separated from the filters by a large-attenuation network. The connection and disconnection of the resistance was effected by the contacts of a periodic-performing relay. The frequency of operation of the relay was set at several operations per second and had varied smoothly during the period of observations. The magnitude of the distortion was determined in its maximum possible terms during the prolonged period of observations.

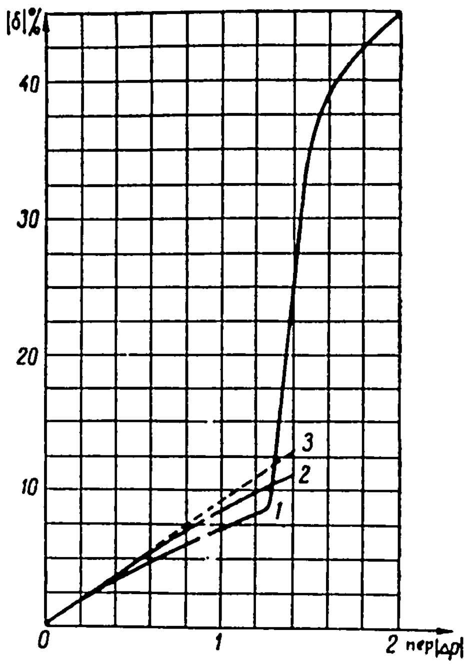


Fig.4

As can be seen from Fig.4, at $\Delta p <$ < 1.3 nepers the distortions are inconsiderable and amount to less than 9 or 10%, but at any further increase in the signal level the distortions increase sharply.

Let us review the process of the appearance of distortions in the FM channel and the ways and means of determining their magnitude.

In the case given, the cause of distortions appears to be the transient process in the receiving filter, as induced by a jump in the signal level. For simplicity's sake let us assume that the filter is an ideal one and that, at modulation, the filter's oscillation frequency changes instantaneously from $\omega_0 - \Delta\omega_0$ to $\omega_0 + \Delta\omega_0$, where ω_0 is the mean channel frequency. At such conditions, and in the absence of a fluctuation in the signal level, the instantaneous value of the signal frequency at the output of the filter is determined (Bibl.3) by the following equations:

$$\omega = \omega_0 + \frac{d\theta}{dt}, \quad (2)$$

$$\theta = \operatorname{arctg} \frac{L \operatorname{tg} \Delta\omega_0 t + K}{\pi}, \quad (3)$$

where

$$L = \mathrm{si}\left(\frac{\Delta\Omega}{2} + \Delta\omega_0\right)t + \mathrm{si}\left(\frac{\Delta\Omega}{2} - \Delta\omega_0\right)t,$$

$$K = \mathrm{ci}\left(\frac{\Delta\Omega}{2} + \Delta\omega_0\right)t + \mathrm{ci}\left(\frac{\Delta\Omega}{2} - \Delta\omega_0\right)t,$$

$$\Delta\Omega = 2\pi\Delta F.$$

Obviously the effect of a fluctuation in the signal level will be maximum when that fluctuation occurs at the same time as a change in the frequency. It can be demonstrated* that at a simultaneous change in frequency and in the level of oscillations at the input of the filter, we will obtain instead of eq.(3):

$$\theta = \mathrm{arctg}\,\frac{(1 + e^{\Delta P})(L \sin \Delta\omega_0 t + K \cos \Delta\omega_0 t) - \pi(1 - e^{\Delta P}) \sin \Delta\omega_0 t}{\pi(1 + e^{\Delta P}) \cos \Delta\omega_0 t - (1 - e^{\Delta P})(L \cos \Delta\omega_0 t - K \cos \Delta\omega_0 t)}. \qquad (4)$$

Figure 5 depicts a graph of the relationship between θ and ΔFt. Curve 1 is plotted pursuant to eq.(3), at a constant oscillation level, while curves 2, 3, 4, and 5, are plotted pursuant to eq.(4), at a decrease in the signal level ($e^{\Delta P}$ = 0.4, 0.2, 0.15, and 0.1). The computations are conducted for $\Delta\Omega = 3\Delta\omega_0$, which approximates the normally accepted ratio of the width of the v - f telegraphy channel to the deviation of the frequency. As can be seen from Fig.5, at a steady signal level the curve θ has its minimum at $\Delta Ft = 0$. At a reduced signal level the minimum shifts to the right; the sharper is the fluctuation the greater is the displacement of the minimum. At an increased signal level these conditions will be reversed; i.e., the sharper is the fluctuation the greater is the displacement of the minimum to the left.

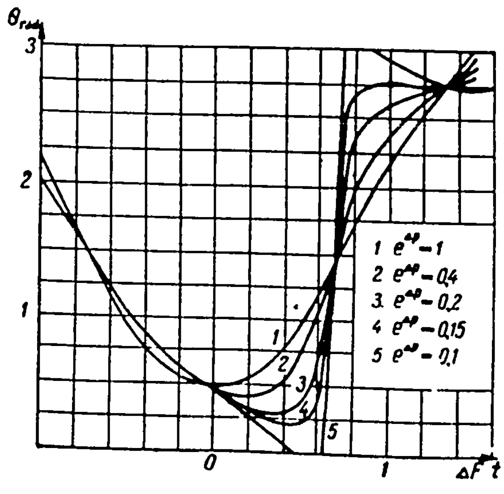


Fig.5

* Vide Appendix 2.

STAT

100

Considering that at the moment when the θ passes through its minimum the frequency of the signal at the output of the filter equals the mean channel frequency and, consequently, the rectified current in the winding of the receiving relay changes its direction at that moment, it is clear that a decrease in the signal level will lead to a retardation of the relay's operating moment, and an increase in that level will lead to a hastening of that moment. It can be also seen from Fig.5 that curves 1, 2 and 3 ($e^{\Delta p} = 1$, 0.4 and 0.2) tend to rise steadily after passing through the minimum, whereas curves 4 and 5 ($e^{\Delta p} = 0.15$ and 0.1) reach a maximum and then a second minimum. This signifies that, beginning at a specific magnitude of the extent of fluctuation, the instantaneous value of the signal's frequency approximates the value of the mean frequency three times in a row, which can lead to either three operations of the relay or to the retardation of the relay's operating moment to approximately the moment of the second minimum. In either case, beginning at a specific magnitude of the Δp, the distortions will increase sharply. As can be seen from Fig.4, this does indeed occur.

In view of the complexity of the whole process, it is hardly possible to find a rating formula for determining distortions at any magnitude of fluctuations in the signal level. Therefore, we will confine ourselves to the range of the relatively small distortions, where the rise in distortions at an increase in fluctuations in the signal level is not so steep. As can be seen from Fig.4, this range comprises fluctuations of up to 1.3 nepers.

At relatively small values of t and Δp, the second member of the denominator of eq.(4) may be ignored, and thereupon:

$$\theta = \operatorname{arctg}\left[\frac{L \operatorname{tg} \Delta\omega_0 t + K}{\pi} + \frac{e^{\Delta p} - 1}{e^{\Delta p} + 1} \operatorname{tg} \Delta\omega_0 t\right].$$

Inasmuch as at low values of t*

---

* Vide Appendix 3

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

$$\operatorname{tg} \Delta\omega_0 t \approx \Delta\omega_0 t,$$

$$\frac{dL}{dt} \approx \Delta\Omega,$$

$$L \approx \Delta\Omega\, t,$$

$$\frac{dK}{dt} \approx -\Delta\Omega \Delta\omega_0 t,$$

therefore

$$\frac{d\theta}{d(\Delta F t)} \approx \left[2\,\Delta F t + \frac{e^{\Delta p}-1}{e^{\Delta p}+1}\right]\frac{\Delta\omega_0}{\Delta F}$$

and θ is at its minimum at

$$\Delta F t \approx 0,5\,\frac{1-e^{\Delta p}}{1+e^{\Delta p}}.$$

In this way, the time of retardation of the relay's operating moment is determined by the requirement

$$t_{ret} \approx 0,5\,\frac{1-e^{\Delta p}}{1+e^{\Delta p}}\,\frac{1}{\Delta F}. \qquad (5)$$

Let us determine distortion as the ratio of the fluctuation in the pulse-restoration moment to the duration of the undistorted elementary pulse $T = \frac{1}{v}$, and we will then obtain:

$$\delta \approx 0,5\,\frac{1-e^{\Delta p}}{1+e^{\Delta p}}\,\frac{v}{\Delta F}\cdot 100\%. \qquad (6)$$

The multiplier $\frac{1-e^{\Delta p}}{1+e^{\Delta p}}$ in the distortion range with which we are concerned, changes almost linearly as depending on the Δp, and therefore this can be written:

$$\delta \approx 0,25\cdot\Delta p\,\frac{v}{\Delta F}\cdot 100\%. \qquad (7)$$

The curves 2 and 3 in Fig.4 yield the rating values of distortions according to eqs.(6) and (7). On comparing these curves with the results of the measurements it can be seen that they are quite satisfactorily commensurate up to Δp =1.3 nepers.

It might be expected that at a gradual increase in the extent of fluctuations

STAT

of the signal level the transition from small to large distortions will be abrupt when the maximum and the second minimum will appear on the curve of θ, just as in the case of $\Delta p \approx 1.9$ nepers. However, as can be seen from curve 1 in Fig.4, the distortions appearing in the range from $\Delta p = 1.3$ nepers to $\Delta p = 1.9$ nepers tend to increase gently and not abruptly even though they increase much more rapidly than at $\Delta p < 1.3$ nepers. This may be explained in the following manner: At an increase in the extent of fluctuations the steepness of the curve of instantaneous frequencies rises steadily beginning with the point of its intersection with ω_o, and at $\Delta p = 1.3$ nepers it becomes much greater than at absence of fluctuation or at a small fluctuation, and, considering that the discriminator has a relatively narrow frequency passband, the increase in the voltage at the discriminator's output will be delayed in comparison with the increase in frequency - which will lead to increased distortions. The small-distortion range can be expanded by expanding the passband of the frequencies transmitted by the discriminator, but this may lead to a rise in distortions caused by external interferences.

Comparison of Distortions in AM and FM Channels

Let us compare the distortions caused by fluctuations in the signal level in AM and FM channels, respectively.

The afore-cited magnitudes of distortions in improved AM systems pertain to channels with receiving filters having a passband width of 90 to 100 cycles, while the width of the passband of the filter of the TT 12/16 type apparatus amounts to approximately 135 cycles. Inasmuch as the value of distortions caused by fluctuations in the signal level is inversely proportional to the width of the passband of the receiving filter, for comparison's sake the values of distortions in the channel of the TT 12/16 apparatus should be increased 1.4 to 1.5 times.

Considering the above and keeping in mind the data cited in Fig.4, we find that an FM channel with a width of 90 - 100 cycles will, at $\Delta p = 0.7$ neper and

STAT

v = 50 bauds, introduce distortions amounting to 8 - 9%, while the channels of the improved AM systems will, in identical conditions, involve 18 - 25% distortion. Moreover, as is to be concluded from the above-cited data, the range of relatively small distortions in FM channels is 0.4 - 0.5 neper wider than in the channels of the improved AM systems.

Appendix 1

Figure 6 depicts in simplified form the curves of the growth of current in the winding of the receiving relay: curve 1 corresponds to the normal level, and curve 2 to a level increased by Δp nepers immediately after the moment of fluctuation in the level. At a current of $\frac{i_0}{2}$ the resultant winding current will equal zero; at a current of $\frac{i_0}{2} + \Delta i$, the relay will be set into operation.

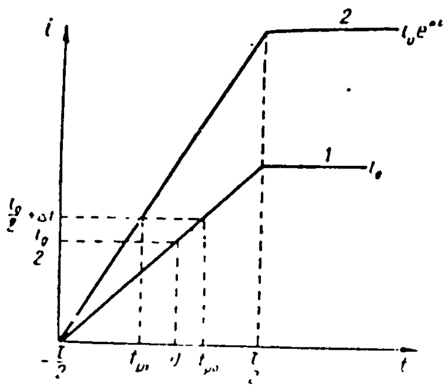


Fig.6

From Fig.6 it can be seen that

$$\frac{i_0}{\frac{i_0}{2} + \Delta i} = \frac{\tau}{\frac{\tau}{2} + t_{pn}},$$

whence

$$t_p = \frac{\tau}{2} \cdot \frac{1}{a},$$

where

$$a = \frac{i}{\Delta i}$$

Further

$$\frac{i_0 e^{\Delta p}}{\frac{i_0}{2} + \Delta i} = \frac{\tau}{\frac{\tau}{2} + t_{p1}},$$

whence

STAT

100

$$t_{p1} = -\frac{\tau}{2}(e^{-\Delta p} - 1) + \frac{\tau}{2} e^{-\Delta p} \frac{1}{a}.$$

On this basis, we obtain

$$\Delta t_1 = t_{p1} - t_{p0} = \frac{\tau}{2}(e^{-\Delta p} - 1)\left(1 + \frac{1}{a}\right).$$

Inasmuch as a fluctuation in the signal level leads to a change in the steepness of the current-rise curve, this will in turn lead to a supplementary change of the relay operating moment, which can be determined on basis of the formula for the time of motion of the relay's reed:

$$t_{dv} = A\sqrt[3]{\frac{\tau}{I_0}},$$

where τ is the time of increase of current in relay winding;

I_0 is the steadied value of current.

At $\Delta p = 0$

$$I_0 = \frac{i_n}{2}.$$

At $\Delta p \neq 0$

$$I_0 = \frac{i_n}{2}(2e^{\Delta p} - 1).$$

Proceeding further, and considering that at low Δp we have $e^{\Delta p} \approx 1 + \Delta p$, we will obtain

$$\Delta t_{dv} = t_{dv}\left(1 - \frac{1}{\sqrt[3]{1 + 2\Delta p}}\right) \approx$$

$$\approx t_{dv}\left(1 - \frac{1}{1 + \frac{2}{3}\Delta p}\right) \approx -\frac{2}{3}\Delta p\, t_{dv}.$$

In this way, the general change in the relay operating moment will be

$$\Delta t = \Delta t_1 + \Delta t_{dv} \approx -\frac{\tau}{2}(1 - e^{-\Delta p})\left(1 + \frac{1}{a}\right) - \frac{2}{3}\Delta p\, t_{dv} =$$

STAT

$$= -\frac{1 - e^{\Delta p}}{2\Delta F}\left(1 + \frac{1}{a}\right) - \frac{2}{3}\Delta p\, t_{dv}.$$

100

On omitting the minus sign we obtain

$$\delta \approx \left[\frac{1-e^{-\Delta p}}{2\Delta F}\left(1+\frac{1}{a}\right)+\frac{2}{3}\,\Delta p\, t_{dv}\right] v\cdot 100\%\,.$$

Appendix 2

It has been demonstrated in Bibl.3 that when the value of the oscillation frequency changes from ω_1 to ω_2 (both values lie within the passband) at the input of the "ideal" filter, the voltage at the output of that filter will, when a unit voltage is fed into the filter's input, change in accordance with the following equation:

$$e_2=\frac{\sin\omega_1 t+\sin\omega_2 t}{2}-\frac{L}{2\pi}(\sin\omega_1 t-\sin\omega_2 t)+$$
$$+\frac{K}{2\pi}(\cos\omega_1 t+\cos\omega_2 t),$$

where L and K have afore-cited values. If the moment of change in frequency coincides with a change in voltage in the equation for $e^{\Delta p}$, the above-cited equation will have this new aspect:

$$e_2=\frac{\sin\omega_1 t+e^{\Delta p}\sin\omega_2 t}{2}-\frac{L}{2\pi}(\sin\omega_1 t-e^{\Delta p}\sin\omega_2 t)+$$
$$+\frac{K}{2\pi}(\cos\omega_1 t+e^{\Delta p}\cos\omega_2 t).$$

Upon making the proper transformations in this last equation, we will obtain

$$e_2=\frac{1}{2}\left[(1+e^{\Delta p})\cos\Delta\omega_0 t-\frac{1-e^{\Delta p}}{\pi}L\cos\Delta\omega_0 t+\right.$$
$$\left.+\frac{1-e^{\Delta p}}{\pi}K\sin\Delta\omega_0 t\right]\sin\Delta\omega_0 t+$$
$$+\frac{1}{2}\left[-(1-e^{\Delta p})\sin\Delta\omega_0 t+\frac{1+e^{\Delta p}}{\pi}L\sin\Delta\omega_0 t+\right.$$
$$\left.+\frac{1+e^{\Delta p}}{\pi}K\cos\Delta\omega_0 t\right]\cos\Delta\omega_0 t,$$

whence

$$\theta=\operatorname{arctg}\frac{(1+e^{\Delta p})(L\sin\Delta\omega_r t+K\cos\Delta\omega_r t)-\pi(1-e^{\Delta p})\sin\Delta\omega_0 t}{\pi(1+e^{\Delta p})\cos\Delta\omega_0 t-(1-e^{\Delta p})(L\cos\Delta\omega_0 t-K\cos\Delta\omega_0 t)}.$$

STAT

Appendix 3

From the equation

$$K = \mathrm{ci}\left(\frac{\Delta\Omega}{2} + \Delta\omega_0\right) t - \mathrm{ci}\left(\frac{\Delta\Omega}{2} - \Delta\omega_0\right) t.$$

It is known that

$$\mathrm{ci}\, x = C + \ln x - \frac{x^2}{2\cdot 2!} + \frac{x^4}{4\cdot 4!} - \ldots,$$

At low values of x the components beginning from the fourth onward may be ignored, and then

$$\mathrm{ci}\, x \approx C + \ln x - \frac{x^2}{2\cdot 2!}.$$

Assuming that

$$x_1 = \left(\frac{\Delta\Omega}{2} + \Delta\omega_0\right) t,$$

$$x_2 = \left(\frac{\Delta\Omega}{2} - \Delta\omega_0\right) t,$$

we will obtain

$$\left(\frac{dK}{dt}\right)_{t\approx 0} \approx -\frac{\left(\frac{\Delta\Omega}{2} + \Delta\omega_0\right)^2 - \left(\frac{\Delta\Omega}{2} - \Delta\omega_0\right)^2}{2}\, t = -\Delta\Omega\, \Delta\omega_0\, t$$

BIBLIOGRAPHY

1 - Description of a BTP50000 AM· V-F Telegraphy System. N.V.Philips Telecommunicale Industrie, Helversum, Netherlands

2. Rudolph, H. - The WT 52/54 V-F Telegraphy System, Fernmeldtechn.Zeitschrift, No.3, 1954

3. Gonorovskiy, I.S. - Frequency Modulation and Its Uses, Svyaz'izdat Publishing House, 1948

Article received by the Editors on 7 January 1956

10

BASIC TYPES OF SWEDISH-PRODUCED DIAL TELEPHONE SYSTEMS WITH CROSSBAR SWITCHES

by

A.D.Kharkevich

This article supplies brief descriptions of the basic types of Swedish-produced urban crossbar-switch dial systems operating on the by-path principle of connection; further, this article describes skeleton diagrams of dial-office and link-connection circuits of selection stages. The article as a whole constitutes a review of foreign periodical press and business-firm prospectuses.

Introduction

The dial telephone systems based on coordinate connectors or crossbar switches as they are called are being introduced on an ever wider scale owing to their numerous advantages.

The development and production of these crossbar-switch dial systems have been evoking special interest in the last few years. The countries which do not manufacture their own equipment for such systems are adapting their production facilities for the manufacture of earlier versions of such systems, developing new versions, or using imported equipment in their telephone networks.

In connection with the development of a domestic Soviet crossbar-switch dial system it is relevant to gain familiarity with foreign experience with the design of such systems. The published Russian-language literature is confined to mentions of the first American-produced crossbar dial systems (Crossbar No.1") (Bibl.1 - 5).

As for the modern crossbar dial systems, and especially the Swedish-produced ones, the related Russian-language literature consists merely of a brief article

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

about the Standard 41 type dial system (Bibl.6) used for the automation of rural telephony in Sweden, and several brief remarks of a general nature in another article (Bibl.7).

A study of the Swedish-produced crossbar systems is of special interest, because Sweden was the first country to have widely applied such systems (Bibl.7). Moreover, Swedish engineers have made very notable contributions to the development of the principles for improved design and construction of such systems.

As early as in 1912 (prior to the invention of the crossbar switch) two Swedish engineers, Betulander and Palmgren, had developed a link system effective when used with low-capacity (mechanical and relay) selectors (Swedish patent No.38,514 issued to the name of Betulander and Palmgren), and subsequently also effective when used with crossbar switches. As formulated by the two Swedes, the principle of primary and secondary selectors (two-step link connection in the selection stage) became the underlying principle of the design of modern crossbar dial systems.

The first application of the crossbar switch was in a direct-connection dial system (Standard 41 type system) (Bibl.6). Such an application of the crossbar switch constituted an efficient utilization of its positive properties with respect to quality of contact and reliability of performance. However, the equipment costs were too elevated to ensure a truly rational use of this type of connector there. The economic and engineering aspects of the problem were resolved only after the crossbar switch, which was first invented by Reynolds, became used jointly with the link system developed by Betulander and Palmgren and operated on the by-path principle of connection.

The widespread introduction of crossbar systems in Sweden is corroborated by the following statistics: As of the present, over 2000 dial offices with crossbar switches are servicing local telephony in Sweden. Approximately one third of all dial telephone sets in Sweden is connected to crossbar systems.

Swedish dial systems are exported to various countries. In the last four

STAT

years (1952 - 1955) the Swedish company Ericsson has manufactured and exported abroad about 200,000 urban, rural and long-distance dial-office systems

The Crossbar Switch (CS)

The modern (1945 type) crossbar switch serving since 1946 to this date as the prototype for the manufacture of crossbar switches for dial telephone systems by the Ericsson Company and the Swedish Telephone and Telegraph Administration is a perfected version of the switch proposed by Betulander and Palmgren as early as in 1919 (Bibl.7).

The outer view of this model, with the casing off, is depicted in Fig.1.

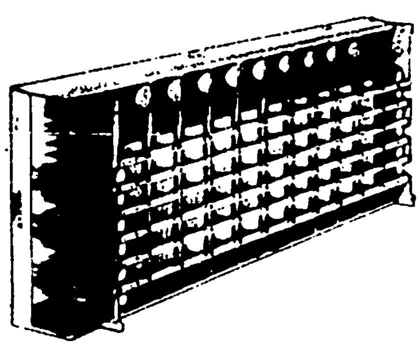


Fig.1

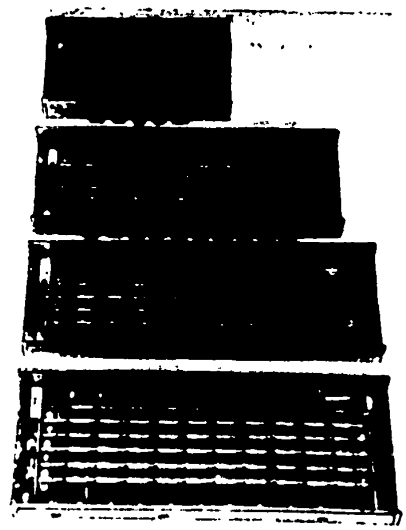


Fig.2

As pointed out in another article (Bibl.8) a dial-system crossbar switch of this design is characterized by its great number of parts of identical form. Aside from a few small exceptions, all parts of this switch can be manufactured by die stamping. The testing and inspection of crossbar parts and assemblies have been wholly automated within the production process. Testing and inspection include: contact-pressure test, clearance test, test of operating and holding current, breakdown test, and so forth.

Table 1 contains data on standard types of crossbar switches manufactured by the Ericsson Company.

Table 1

a)	b)	c)	d)	e) f)	e) g)	e) h)	i)
RVD 100–109	10	5	j)	583	190	136	12,5
RVD 110–119	10	5	k)	535	190	136	12
RVD 130–139	5	5	l)	345	190	136	7,5
RVD 210–219	10	6	m)	583	220	136	13,5

a) Type of crossbar switch ; b) Number of vertical units; c) Number of selecting bars; d) Maximum number of contact strips; e) Dimensions in mm; f) Length; g) Height; h) Thickness; i) Weight in kg; j) 10 vertical units with 10 contact strips each; k) Two vertical units with 10 contact strips each and 8 vertical units with 8 contact strips each; l) Two vertical units with 10 contact strips each and 3 vertical units with 8 contact strips each; m) 10 vertical units with 10 contact strips each

The number of vertical units and of the contact strips within these units can be varied somewhat within the limits of each type. The crossbar switches are adapted for working voltages of 24, 36, 48, and 60 v.

Figure 2 depicts photographs of the four crossbar types named in Table 1.

Basic Types of Swedish-Produced Crossbar Systems

As of the present, the Swedish Telephone and Telegraph Administration and the

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

Ericsson Company are producing several different types of crossbar systems (Bibl.9, 10 10).

The type A-204 is the basic type of crossbar system to be used for the automation of urban telephone networks in Sweden.

The crossbar systems exported to other countries are those of the ARF-10 type designed for the automation of medium- and large-capacity urban telephone networks. Dial offices of the ARF-10 type operate in Denmark, Burma, Indonesia, and Finland. Further, there is the ARF-50 type, designed for joint operation with the Siemens step-by-step dial system, because a crossbar system of this type can be interworked with the Siemens system without intermediary equipment. Such joint offices operate in Finland, Pakistan and Rhodesia. Type ARF-51 has a skeleton diagram similar to that of type ARF-50 but is designed for interworking with step-by-step systems in the dial areas of the British Postal Administration.

The following offices are manufactured for rural use: terminal offices with a capacity of up to 60 numbers (ARK-312); terminal offices with a capacity of 60 to 180 numbers (ARK-314); terminal offices with a capacity of 100 to 1200 numbers (ARK-315); and tandem central offices with a capacity of 100 to 1200 numbers (ARK-335).

The automation of suburban and long-distance telephony is effected by using dial offices of the following types: long-distance, suburban and tandem offices with capacities of 40 - 200 lines (ARM-501) and 100 - 1600 lines (ARM-502), medium- and large-capacity long-distance offices (ARM-20), and others.

Type A-204 Dial System

The crossbar switch used in the development of the A-204 dial system was based on single-step link connection (one high-capacity selector in the connection route within the limits of the selection stage), and on the direct-drive principle (Standard 41). Therefore, subsequent work on the improvement of the A-204 has been prin- STAT cipally oriented toward reducing the bulk of equipment by utilizing the link-system and control ideas suggested by Betulander and Palmgren. The stimulus for the development of this type of crossbar system was, as indicated in an article (Bibl.11), the successful employment of the by-path principle of connection in crossbar systems in the USA.

The A-204 type crossbar system operates on two-step link connection (two low-capacity selectors in the connection route within the limits of the selection stage), with the by-path principle of connection. A hundred-element crossbar switch (number of vertical units n = 10; capacity per vertical unit m = 10) serves as the basic commuting mechanism.

Contrary to the general belief that the by-path principle of connection complicates circuits, dial offices of the A-204 type have a simpler circuit than those of the Standard 41 direct-connection type. The relay used in the A-204 type system is simpler from the viewpoint of control, and therefore the overhead and operating expenses of this system are lower than those of the Standard 41 system.

The A-204 type systems are designed for automation of telephone networks of any capacity upward of 100 numbers. Their battery-supply voltage amounts to 36 v. Such systems have been applied in the telephone networks of many Swedish cities.

<u>The Skeleton Diagram</u> of an A-204 type system with a capacity of up to 10,000 numbers, applicable for a multioffice metropolitan exchange without toll-switching planning, is depicted in Fig.3. Here, the customary nominal designation of selector is represented by a crossbar-switch vertical unit.

This skeleton diagram is designed in conformance with the decadic system (decadic principle) and contains three group selector (GS) stages and one subscriber's stage (SS). The subscriber's stage fulfills the function of a combined finder and connector stage and is used for servicing outgoing and incoming subscriber calls.

The subscriber's stage thus consists of two connection steps. The crossbar switches of the connection step linked to subscriber lines pass a combined flow of

STAT

outgoing and incoming telephone traffic. The vertical units of the crossbar switches
10 of the other connection step are used separately for outgoing and separately for incoming traffic.

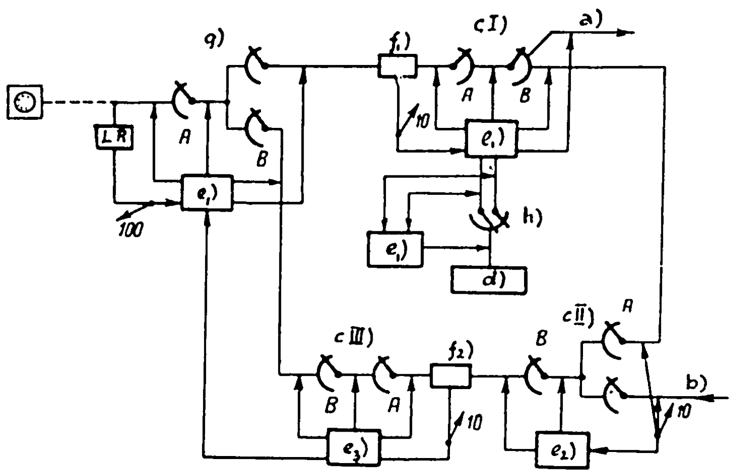


Fig.3

a) Toward other dial offices; b) From other dial offices; c) Group selectors I GS, II GS, III GS; d) Register; e) Marker sets MS_1, MS_2; f) Lineswitch trunk sets LTS_1, LTS_2; g) Subscribers' stage SS; h) Register finder RF

One subscriber's stage unit makes it possible to connect a hundred subscriber lines, and consequently the system operates with groups of hundred lines each. Each of the three group selectors (GS) consists of two connection steps. The outlets of each group-selector stage are divided into 10 routes. Therefore, the third group selector (III GS) stage routes ten 100-line groups, thus forming 1000-line groups. The second group selector or II GS stage routes ten 1000-line groups, thus ensuring the dial office's capacity of up to 10,000 numbers, and the first group selector stage (I GS) connects to local dial offices.

Dial systems of the A-204 type use registers. Each selection stage is provided with markers (MS) for groups of connection routes. The connection of a register to

100

the marker of I GS is effected by means of register finder (RF) and marker MS_p. All markers except MS_A contain crossbar switches in addition to relays. The marker MS_A contains a relay only.

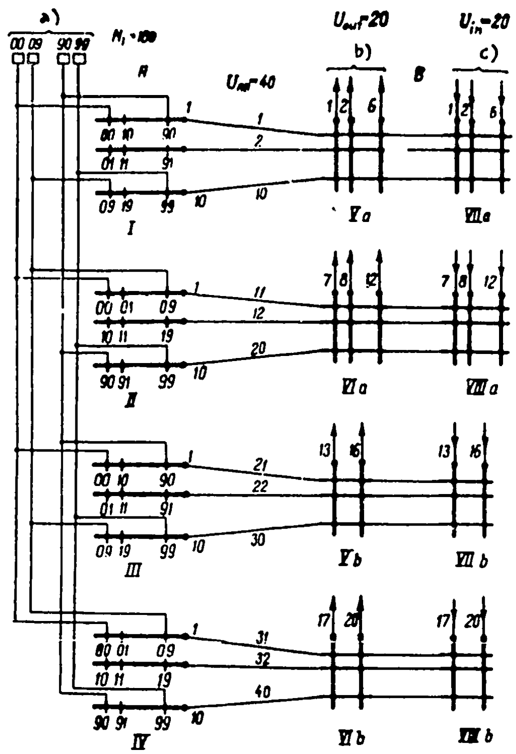


Fig.4

a) Lines; b) Toward I GS; c) From III GS

The lineswitch trunk set LTS_1 controls the outgoing part of the connection route and participates in the connection of register to trunk when a number is being dialed. The incoming lineswitch trunk set LTS_2 is designed for effecting the power supply and for sending calls and "busy" signal.

Considering that the two connection steps of the subscriber's stage unit, used in handling incoming calls, cannot ensure low losses during pulse action of subscriber line, the finding of the subscriber line is carried out simultaneously through the subscriber's stage and the third group selector (III GS) stage. For this purpose, the markers MS_A and MS_3 of the corresponding selection stages are interlinked by control circuits. The crossbar switches of I and II GS commutate four wires, and the switches of the subscriber's selection stage and III GS commutate five wires.

A subscriber's-stage 100-line unit has 20 trunks for outgoing and 20 for incoming traffic. Each route of a group selector unit has 20 outlets to the subsequent selection stage. The connection routes (trunks) between the selection stages are connected stepwise except for the trunks between III GS and subscriber's stage. Intermediate distributing frames can be used for this purpose.

The transmission of a number from the register to the marker may be effected by pulses analogously to the (decadic-system) pulse dialing, but by means of coding. Consequently, provision is made for two types of markers. Coded transmission of pulses ensures higher speed, which may be of major importance on larger telephone networks. Pulse dialing makes possible interworking with step-by-step dial systems without having to resort to intermediary equipment. In small rural dial offices the markers receive numbers directly from subscribers and hence registers are unnecessary there.

To reduce the costs and labor involved in servicing dial offices, the markers are endowed with certain testing and fault-detecting functions. In the event of detection of a fault, the concerned marker connects to a recording device and effects the recording of the failure.

<u>The Subscriber's Stage (SS)</u> consists of a varying number of units in dependence on the local capacity of the system. An SS unit contains, in turn, a set of crossbar switches creating, in a 100-line group, 20 routes toward I GS for outgoing subscriber traffic load and 20 routes from III GS for incoming load. There are eight 100-element crossbar switches in a unit. The link circuit of such a unit is shown in Fig.4.

The crossbar switches CS I to CSI to CS IV form the first SS connection step

STAT

(Step A). Subscriber lines are connected to the contact bank of these switches. Inasmuch as each switch (I to IV) in this connection step has 10 vertical units with a capacity of 10 outlets each (in Fig.4 vertical units are designated by heavy black lines with thickened ends), each switch contains an entire hundred lines. The order of sequence of the connection lines to switches CS II and CS IV differs from that

a)

b)

1	2	3	...	9	10
00	01	02	...	08	09
10	11	12	...	18	19
20	21	22	...	28	29
...	...	...		...	...
80	81	82	...	88	89
90	91	92	...	98	99

c)

1	2	3	...	9	10
00	10	20	...	80	90
01	11	21	...	81	91
02	12	22	...	82	92
...	...	...		...	...
08	18	28	...	88	98
09	19	29	...	89	99

Fig.5

a) Vertical units; b) CS I and CS III; c) CS II and CS IV

used in switches CS I and CS III. Accordingly, in switches CS I and CS III the lines are so connected as to cause the contact bank of each vertical unit to contain 10 lines with identical unit digits, while in switches CS II and CS IV the lines are so connected as to cause the contact bank of each vertical unit to contain 10 lines with identical decadic digits.

Such a connection of subscriber lines, termed transposed connection, is a form of grading. Its application facilitates a more equal distribution of traffic load among connection routes, with the traffic loads of separate groups of tens being intershifted (the load on connection routes occupied consecutively for the second, third, and fourth time, corresponds better to Poisson's distribution than in the event of separate trunk groups for every ten lines). Figure 5 depicts the panel for connection of subscriber lines to the first crossbar-switch group (CS I and CS III) and the second crossbar-switch group (CS II and CS IV). If it is considered that each line has four connection routes accessible between the A and B connection steps, the circuit for the grading of connection

STAT

routes corresponding to a transposed system of connection will have the form shown in Fig.6.

The crossbar switches CS V - CS VIII (Fig.4) form the second connection step

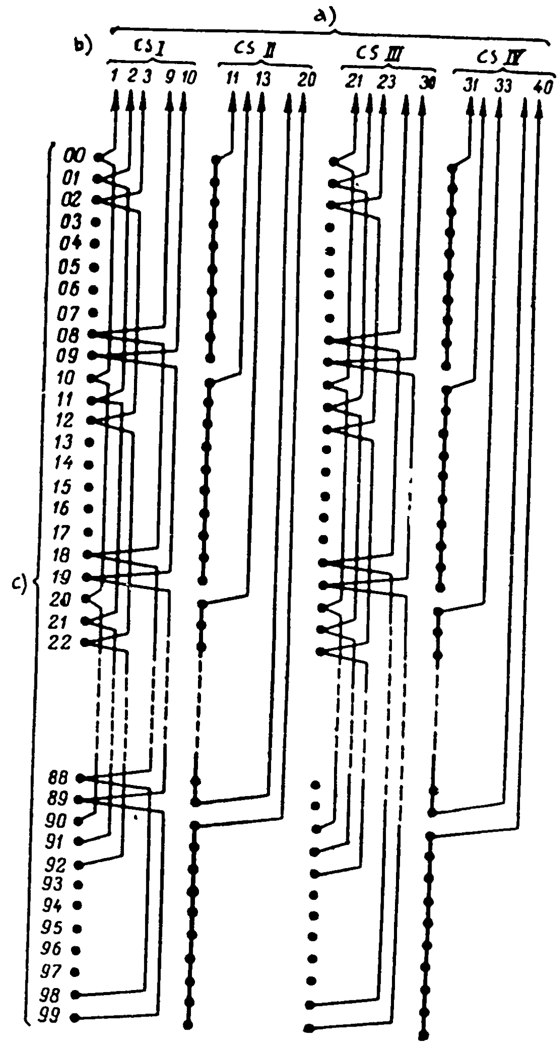


Fig.6

a) Trunks; b) CS I to IV; c) Subscriber lines

(Stage B) of an SS unit. Each crossbar switch in this group is divided into two

STAT

parts, having four and six vertical units respectively. In Fig.4 these parts are denoted by letters a and b. The twenty vertical units of crossbar switches CS V and CS VI are used for outgoing traffic from a 100-line group, while a like number of vertical units in switches CS VII and CS VIII is used for servicing incoming traffic. The contact banks of switches CS V to CS VIII are arranged parallelwise, as shown in Fig.4, and form 40 connection routes to the 40 vertical units of the first connection step A (CS I - IV). The connection routes service both the incoming and

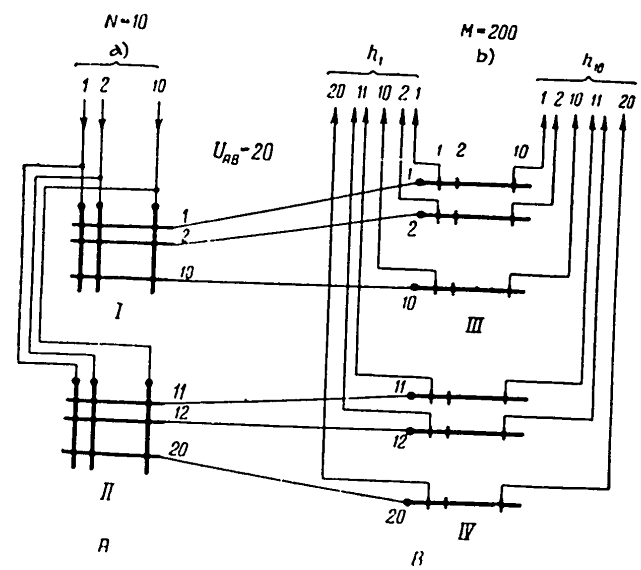


Fig.7

a) Inlets; b) Outlets

the outgoing traffic. Although they are divided into four trunk groups of 10 routes each, any one of these trunk groups can be used to identical extent for connecting to the entire subscriber group.

In outgoing and incoming traffic the finding of an available connection route is effected in a fixed order of sequence. First, a connection route is found in Trunk Group No.1 connected to CS I; second, a connection route is found in Trunk

Group No.2 connected to CS II, and so forth. At such an order of sequence of the occupation of connection routes, the routes of the first two trunk groups pass a large traffic load and therefore their servicing is ensured by the provision of a large number of trunks connecting to I GS and from III GS.

The outgoing trunks to I GS are crossed by like trunks of the other SS units so as to provide 50 - 70 trunks outgoing to I GS for a 1000-line group (10 SS units).

The above-described SS unit of the A-204 dial system is designed for a traffic load of 6.7 erlangs (i.e., 240 calls lasting 100 sec), at losses of $p = 0.002$. The establishment of incoming connection through the connector unit serves to test a line for the "busy" tone.

<u>The Group Selector Stage (GS)</u> consists of separate GS units. A GS unit, whose link diagram is depicted in Fig.7, consists of four 100-element crossbar switches. Such a unit has 10 inlets and 200 outlets and, if holding is ignored, it can link 10 selectors (for instance, selectors of the "Strowger 32-a" type) into 200 outlets. Such a unit is serviced by a master marker, and the number of simultaneous connections can reach 10. Such a unit is used in all GS stages.

The vertical units of switches I and II [first(A) connection step of GS unit] are linked by trunks with the preceding GS unit and, inasmuch as the vertical units of these two switches are connected in parallel pairs (Fig.7), the 20 vertical units of these switches form 10 inlets to the GS unit. The contact banks of the said two switches are in parallel and form 20 outlets toward the second connection step (Step B) of the GS unit. The vertical units of the crossbar switches III and STAT connect to connection routes, while the contact banks of these switches are not in parallel, and form 200 outlets. These outlets are divided into 10 routes ($h_1 - h_{10}$) of 20 outlets each. The outlets of separate groups in the GS cross each other in all GS stages save the last, thus forming stepwise connection to the necessary number of trunks in the subsequent GS stage.

According to the nominal designations used by the Ericsson Company and in the

Swedish crossbar-switch literature, the circuit in Fig.7 could be represented in the form shown in Fig.8. Such a representation is less graphic, but despite its compactness it has the advantage of showing not only the design principle but also the actual location of vertical units in individual crossbar switches. In accordance with the Swedish designations, in Fig.8 each vertical unit is denoted by a circle with a dashline attached and the vertical units of a single crossbar switch are comprised in a rectangle. The in-parallel arrangement of the contact banks of several vertical units is denoted by the identical direction of the dashlines of these units. The combination of two vertical units to form a single doubled-capacity selector is indicated by connecting these units by means of a dashline.

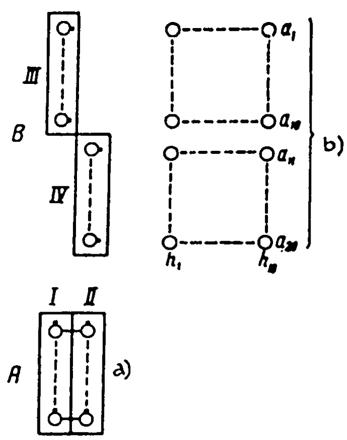


Fig.8

a) 10 inlets; b) 200 outlets

From Fig.7 it can be seen that all the 20 outlets in each direction in a GS unit are available for the first call only. After that first call appears and occupies a connection route, the number of available routes for the second call in the subsequent selection stage will be reduced by one in each of the 10 routes and will equal 19. When a tenth call occupies the last inlet of the GS unit, the number of available outlets will be 11.

In this way, the circuit of a GS unit in a dial system of the A-204 type ensures a greater utilization of outlets toward the next selection stage than at an availability of ten, and a smaller such utilization that at an availability of 20.

STAT

Type ARF-10 Dial System

Dial systems of the ARF-10 type are designed for the automation of medium- and large capacity urban telephone networks. As in dial systems of the A-204 type, the ARF-10's operate on the principle of multistage linking and by-path connection.

The basic commutating mechanism of the ARF-10 is a 200-element crossbar switch ($n = 10$, $m = 10$). Its high qualities ensure low operating expenses (Bibl.12). According to the official data of the Swedish Telephone and Telegraph Administration (Bibl.12), the operating expenses on crossbar dial offices amount to 0.5 working hour

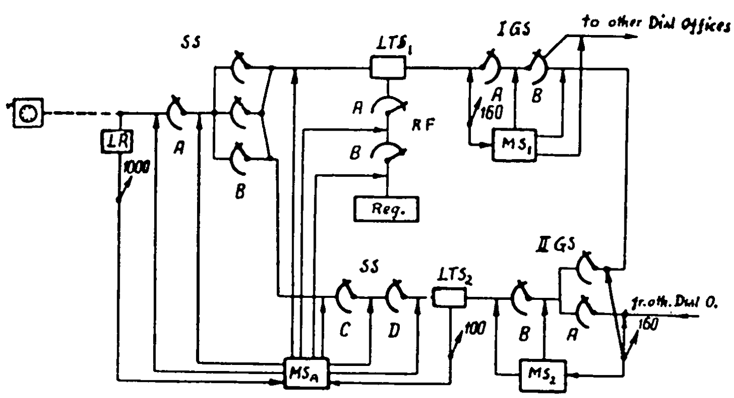


Fig.9

per number per year, aside from the operating expenses of distribution frames and power sources. It is also recorded that a 10,000-number office in the city of Malmo (Sweden) is serviced by shifts of two operators each.

Figure 9 illustrates a skeleton diagram of a dial office of the ARF-10 type, with a capacity of up to 10,000 numbers within a multioffice city exchange without toll-switching planning. The diagram contains two GS stages and one SS stage which, as in A-204, carries out the function of a combined finder-connector stage.

The GS stages are subdivided into two connection steps each, while the SS stage consists of four connection steps. Subscriber lines are combined into 1000-line groups in the SS stage. As per the diagram in Fig.9, the traffic load outgoing from a 1000-line group traverses two connection steps, A and B, while the incoming traffic load traverses all the four connection steps A, B, C, and D.

In step B the incoming and outgoing traffic loads become partially combined by the vertical units of that step. In step A the two traffic loads become completely combined.

While the contact bank of III GS services 10 routes, the stages II GS and SS form a ten-thousand office group, and each route of stage I GS contains lines leading toward the 10,000 group.

Each selection stage is provided with a marker MS. The marker MS_A of the SS stage establishes the outgoing connections of the 1000-line group and controls the connection of the register to a trunk.

The outgoing lineswitch trunk set LTS_1 contains the battery supply relay of the calling subscriber, the register-connecting relay, and the signal relay. The incoming lineswitch trunk set LTS_2 contains the battery supply relay of the called subscriber, and the relay for connecting marker MS_A. The connection of the register to LTS_1 is effected by means of a register-finder stage consisting of two connection steps A and B.

The connection routes outgoing to the other dial offices connect to the contact bank of the B connection step of I GS. The connection routes incoming from the other dial offices connect to the vertical units of the A step of II GS.

The transmission of the digits of a number from the register to the markers is effected by means of d.c. pulses of varying polarity transmitted by speech-wire circuits. The average time of transmission of one digit amounts to 35 milliseconds.

STAT

When two subscribers talk over two different dial offices of the ARF-10 type, their connection in the terminal (incoming) office is effected under the control of the register of the outgoing office. In this case, the number is transmitted by a code. But when the calling or called party is connected to a step-by-step or rotary-

10

switching dial office, the coded pulses are replaced by their corresponding direct dialing pulses or revertive pulses. For this purpose a relay for converting to the needed type of pulse is applied in telephone networks with dial offices of divers types, or, if such relay is lacking, special intermediary equipment is installed there.

At incoming traffic from offices of another dial system, the incoming traffic ARF-10 office is equipped with intermediary registers for converting the pulses received from the outgoing dial offices into a code that can be processed by the ARF-10, and thus carrying out a successful connection of the incoming calls.

The ARF-10 type dial system is particularly applicable in long-distance telephony.

The system is provided with special control equipment which tests the efficiency of performance of the control devices and records the traffic loads.

The commutating equipment of ARF-10 type dial offices is located in bays of ten crossbar switches each. Height of bay: 2900 mm.

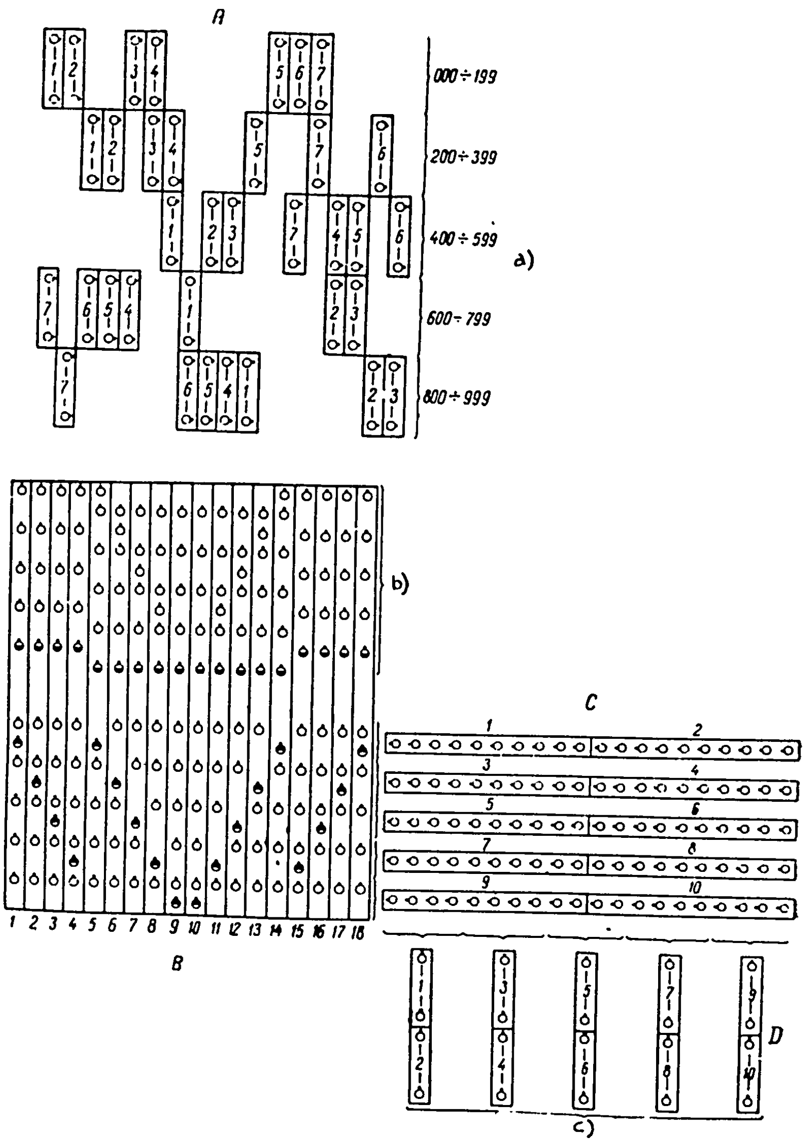
The space needed to accommodate the equipment depends, naturally, on the form of the switch room, traffic load, number of connection routes, and other factors. According to bibliographical references (12, 15 and 16), at the typical arrangement of equipment in a 10,000-line office of a multi-office city exchange without toll-switching planning (as per the diagram shown in Fig.9), the switch room space amounts to 235 m^2 (18.5 × 12.7 m). Here, all the equipment is arranged in 16 rows having a maximum length of 10,026 mm, a maximum width of 350 mm, and a maximum height of 2900 mm.

STAT

<u>The Subscriber's Stage</u> consists of 1000-line SS units. Figure 10 depicts the link diagram of a 1000-line unit designed for servicing of equally high incoming and outgoing traffic loads amounting to 59 erlangs each, and at losses of $p = 0.002$.

The set of crossbar switches included in the SS unit creates – in accordance with the link diagram shown in Fig.10 – for the 1000-line group, 98 trunks toward

10C



STAT

Fig.10

a) Subscriber lines; b) 98 outlets to IGS; c) 100 inlets to II GS

I GS for outgoing traffic and 100 trunks from II GS for incoming traffic. The SS unit consists of four connection steps. Subscriber lines connect to the contact bank of the crossbar switches of the first step (Step A).

When the lines are connected the 1000-line group becomes subdivided into five 200-line groups each of which is serviced by 70 CS vertical units in Step A. In this way, at a specific traffic load of $y = 0.059$ erlangs per line, the number of CS vertical units K in Step A servicing a group of 20 lines amounts to 7. The total number of vertical units in Step A - and consequently of connection routes between Steps A and B - amounts to 350 ($V_{AB} = 350$).

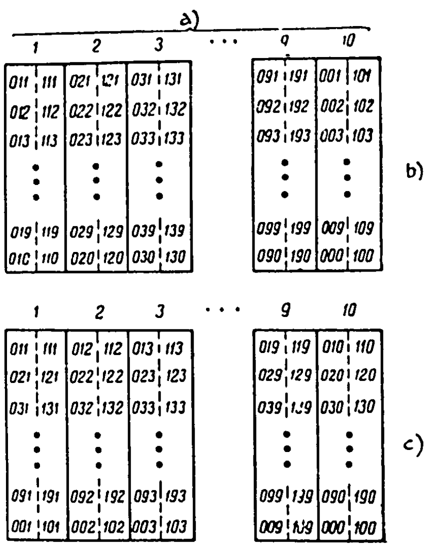


Fig.11

a) Vertical units; b) CS I - IV;

c) CS V - VIII

The system operates on transposed connection of lines in accordance with the panel depicted in Fig.11.

The second connection step (Step B) of the SS unit makes it possible to connect to the vertical units of the crossbar switches of this step 98 outgoing trunks from I GS and 100 incoming traffic connection routes between Steps B and C. A part of the vertical units of Step B is (as indicated in Fig.10 by blackening the upper or lower part of some of the circles designating the CS vertical units) used for both incoming and outgoing traffic. Steps C and D are designed for connecting to Step B the hundred trunks outgoing from II GS. The transposed connection effected by these steps ensures normal losses ($p = 0.002$) in the process of the pulse-action finding of specific lines by incoming traffic.

STAT

The SS unit consists of seventy three 200-element crossbar switches. Step A contains 35 switches, Step B - 18 switches, and Steps C and D - 10 switches each.

The number of vertical units in crossbar switches and, hence, the number of crossbar switches per se in each connection step of the SS unit, and the number of connection routes between the steps, all depend on the traffic load at a given quality of service. The related literature (Bibl.12) contains data on the number of vertical units and connection routes as depending on traffic load at p = 0.002 and mutually equal incoming and outgoing traffic loads (Table 2).

Table 2

a)	b)	c)				d)	
Y_{sel}	k	A	B	C	D	LTS	LTS
46	6	300	150	80	80	80	80
59	7	350	180	100	100	98	100
73	8	400	200	120	120	100	120
101	10	500	300	160	160	160	160

a) Outgoing traffic load per 1000-line group, in erlangs; b) Number of vertical units per 20 lines; c) Number of vertical units per 1000-line group in each connection step; d) Number of lineswitch trunk sets per 1000-line group

Let us now explore the group selector (GS stages).

Fundamentally speaking, all GS stages are identical. The so-called GS unit represents the element of a GS stage from the diagrammatic and design viewpoint. A GS unit, whose link diagram is depicted in Fig.12, has 80 inlets and 400 outlets which may be divided by 10 or 20 routes into 40 or 20 outlets per route. Figure 12 pertains to the latter instance, i.e., 20 routes (h = 20). The GS unit contains two connection steps. Trunks from the preceding group selector connect to the 80 vertical units of the first step (Step A) of the GS unit, and these units are, in turn, linked by inside connection routes to the 120 vertical units of step B. The contact

STAT

bank of the Step-B crossbar switches forms the outlets to the subsequent group selector stage.

At the presence of 20 outlets for each route, the minimum availability of the outlets of each route amounts to 7, and the maximum - to 20. The mean coefficient of expansion for the GS stage amounts to = 1.5, and thus 120 connection routes are provided within the GS stage for the 80 inlets.

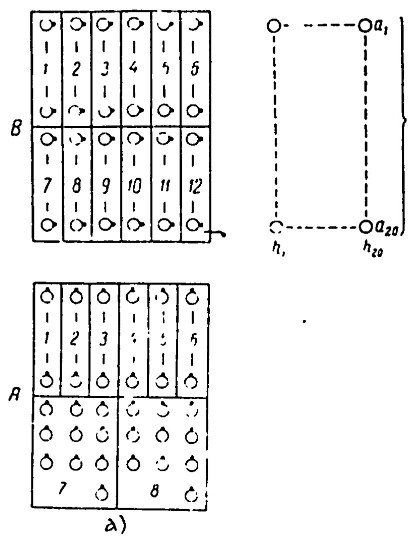


Fig.12

a) 80 inlets; b) 400 outlets

Type ARF-50 Dial System

Dial offices of this system type are designed for interworking with urban dial offices of the step-by-step system. This is because the ARF-50 dial offices require no intermediary equipment for such interworking. Further, dial offices of this system operate with registers (distributed according to selection stage) and on the by-path principle of connection. The dial-office battery voltage amounts to 60 volts. As of 1950 the ARF-50 system has been servicing the dial areas of the city of Helsinki (Bibl.17).

A skeleton diagram of an ARF-50 type dial office with a capacity of 10,000 numbers, operating in a multi-office city exchange without toll-switching planning, is depicted in Fig.13.

As can be seen from Fig.13, the skeleton diagram of this dial office contains one subscriber's stage and two group selector stages. Dial offices of this type operate with 1000-line groups whose number is determined by the capacity of the SS unit.

The subscriber's stage SS consists of four connection steps A,B, C, and D, as in dial offices of the ARF-10 type (Fig.10), and differs from the latter solely by

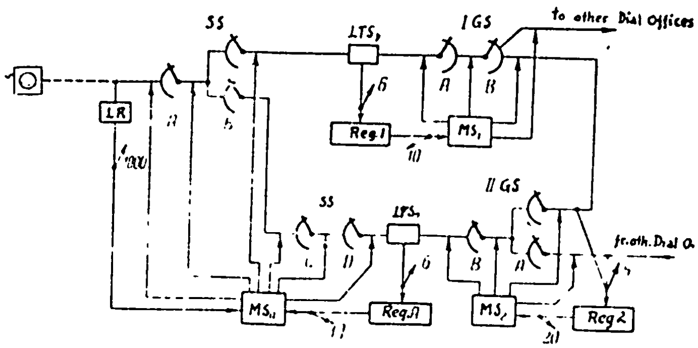


Fig.13

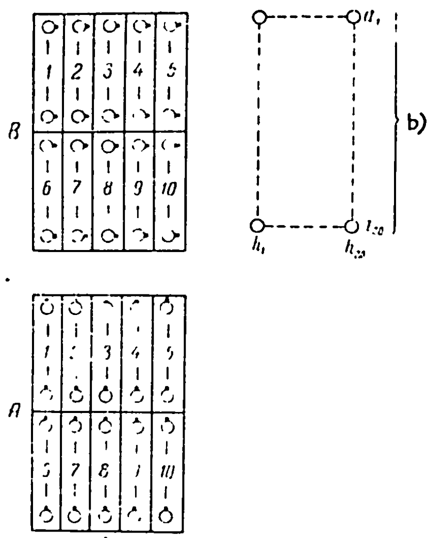
a) Subscriber's stage SS; b) Lineswitch trunk set LTS; c) I, II and III GS's; d) Toward other dial offices; e) From other dial offices; f) Register; g) Marker MS

its number of outgoing and incoming trunks. In the skeleton diagram in Fig.13 the number of outgoing trunks is the same as that of the incoming ones and equals 60.

Likewise, just as in ARF-10, the ARF-50 subscriber's stage unit uses only two connection steps A and B, for outgoing traffic but uses all four steps A, B, C, and D, for incoming traffic. Step A handles both incoming and outgoing traffic. In the here-described instance, the SS unit contains not only the crossbar switches forming the four connection steps but also a control marker MS_A for the unit as a whole, and it also contains 10 registers (Reg.A), 60 lineswitch trunk sets LTS_1 for outgoing trunks, and as many lineswitch trunk sets LTS_2 for incoming trunks.

Each register services six incoming trunks. A register can service only one trunk at a time. The duration of the operation of marker MS_A at incoming connection amounts to about 350 milliseconds. The connections are effected in a transposing manner just as in ARF-10.

The GS stages have their own registers and markers. The link diagram of the I GS unit is the same as in dial offices of the ARF-10 type. A register services six trunks, while a marker MS_1 services 10 registers. The register of I GS can receive one or two digits of a subscriber number and transmit them immediately after registration to marker MS_1 which establishes a connection route toward II GS.



a)

Fig.14

a) 100 inlets; b) 400 outlets

The II GS unit is designed as shown in Fig.14 and has 100 inlets and 400 outlets which, in the presence of 10 routes, are broken down into 10 groups of 40 outlets each (in Fig.14 the distribution is among 20 routes). The register of this GS stage services five trunks, and 20 such registers are serviced by a common marker MS_2.

CONCLUSIONS

Knowledge of the related literature, some of which is cited in the appended bibliography, supplies grounds for assuming that the Swedish-produced crossbar dial systems are completely up-to-date from the engineering viewpoint. These systems should be studied in detail for the purpose of incorporating the best solutions into the design of a domestic Soviet crossbar dial system.

Dial offices of the A-204 type are suitable for use in low-congestion dial areas where it is expedient to have low-capacity stations and use branch dial offices. The ARF-10 type is expedient for construction of medium- and large-capacity offices with 1000 and more numbers. The employment of offices of the ARF-50 type is expedient in dial areas serviced by offices of the step-by-step system because no intermediary equipment is required there for interworking.

A domestic Soviet crossbar dial system should, in all likelihood, incorporate the engineering features of divers types of dial systems, because the conditions of operation of this system vary so extensively throughout the great territory of the Soviet Union that telephone networks must be constructed economically as depending on their capacity in a given area. Moreover, it would be necessary to ensure the interworking of the crossbar system with step-by-step and rotary-switching systems on some telephone networks, aside from the networks where the crossbar system is to operate alone.

BIBLIOGRAPHY

1. Markhay,Ye.V. and Babitskiy, I.A. - Dial Offices. Part 1, Svyaz'izdat Publishing House, 1944, pp.207-215
2. Karmazov, M.G. - Dial Telephony. Svyaz'izdat, 1947, pp.192-211
3. Markhay,Ye.V. and Babitskiy, I.A. - Dial Telephony, Svyaz'izdat, 1950, pp.325-334
4. Babitskiy, I.A. - The Crossbar Dial System, Vestnik svyazi (Elektrosvyaz'), No.3, 1945, pp.22, 23; No.4, 1945, pp.19, 20
5. Karmazov, M.G. - Crossbar Dial Systems, Tekhnika Svyazi Compendium Svyaz'izdat, July 1947, pp.3-37
6. Farafonov, L.S. - Standard-41 Automatic Telephone Systems, Vestnik svyazi, No.2, 1956
7. Kharkevich, A.D. - Developmental Trends of Dial System Engineering, Elektrosvyaz', No.3, 1956
8. Wiberg,Ye.A. - The Crossbar Switch - Its Design and Manufacture, Ericsson Review, No.2, 1949
9. L.M.Ericsson - Automatic Telephone Exchanges with Crossbar Switches. Reference

STAT

[redacted]

List, March 1955

10. – Scientific Research Institute for Urban and Rural Telephony, Information about Urban and Rural Telephony Techniques. No.1, 1956

11. Rost, H.F., Modee, G., Ek, F., and Strandlund, H. – A New Common Control Crossbar Automatic Telephone System of the Swedish Telephone and Telegraph Administration, Trans. AIEE, Vol.70, Part II, 1951

12. – L.M.Ericsson's Automatic Telephone Exchanges with Crossbar Switches. System ARF-10, Stockholm, July 1953

13. Berglund, C. – Cut-Over to Crossbar System in Aarhus, Ericsson Review, No.3, 1953

14. Borregaard, N. – Conversion of the Aarhus Telephone Exchanges. Ericsson Review, No.3, 1953

15. Stenbaek, S. – Choice of Automatic Telephone System for Copenhagen and District. Ericsson Review, No.4, 1953

16. Johansen, V. – The Copenhagen New Automatic Telephone System with Crossbar Switches. Ericsson Review, No.4, 1953

17. Karlsson, S.A. – The L.M.Ericsson Crossbar Switch System in Helsinki. Ericsson Review, No.4, 1950

Article received by Editors on 7 June 1956

STAT

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

FOREIGN PRESS NOTES

A SYSTEM FOR MULTIPLEXING THE CIRCUITS OF REGIONAL TELEPHONE NETWORKS

The Bell System Company has developed a system for multiplexing regional telephone circuits, and termed it "P1". The related equipment can be used on cable and aerial (copper, metallic and bimetallic) subscriber lines. The system is so designed as to form one to four channels, each having a width of 200 to 3000 cycles, in a single circuit. A spectrum of 8 - 100 kc is used at four-channel operation.

The equipment is based on amplitude modulation, and transmission comprises both the currents of both sidebands and the carrier-frequency current. It is possible to design the linear spectrum of the system in three different versions.

The first version provides the possibility for a gradual increase in the number of channels up to four. In this case, the spectra of a single channel corresponding to both directions of transmission are arranged in a row. The distance between carrier frequencies amounts to 12 kc (values of carrier frequencies: 12, 24, 36, ..., 96 kc). In this version, the system operates without through repeaters.

The two other versions of spectrum structure are used whenever it is necessary to include through repeaters (which may amount to up to four). In this case, the spectra of the four channels running in one direction are arranged in the lower part of the 8 - 100 kc range, and the spectra of the four channels' other direction are arranged in the upper part of that range. Here, the carrier frequencies can be spaced either as in the first version or at a displacement of 6 kc upward (with the number of channels being thereby reduced to three).

STAT

A characteristic trait of the P1 apparatus is its lack of electron tubes - their functions are fulfilled by transistor devices. Another special feature is the application of compandor devices.

Among the important elements of the system are the two- and three-stage ampli-

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

fiers operating on n-p-n type transistors. The two-stage amplifiers are employed in terminal equipment as HF (transmission and reception) and LF (compandor) amplifiers. The three-stage amplifier is a basic element of the tandem office. All amplifiers perform with negative feedback.

The P1 apparatus operates with one conversion of frequency. The converters used consist of ring and bridge circuits comprising silicon diodes. The carrier-frequency generators are assembled on the basis of two transistors, one serving as the generator proper and the other as the amplifier. The compandor device incorporates silicon diodes.

The tandem office contains two (one for each direction of transmission) three-stage repeaters, directional filters, equalizers and, in some cases, AGC devices. With regard to the transmission from central office to a subscriber, the AGC utilizes the total power of the carrier-frequency currents; in an opposite direction, the carrier-frequency currents are transmitted only during the period of channel operation (so as to reduce the power expenditures) - therefore, pilot-frequency current is transmitted in that direction.

The maximum attenuation of upper-frequency currents in all channels that can be compensated by the P1 apparatus amounts to approximately 30 db. The highest sideband transmission level is assumed at - 8 db, and the carrier-frequency level - at 12 db higher.

A subscriber can be called from the central office by using currents with the frequencies of 1150, 1750 and 2500 cycles. By combining these frequencies it is possible to call up each of the four subscribers who can be connected to the line.

The power-supply equipment of the apparatus includes a full-wave rectifier based on a bridge circuit containing germanium or silicon diodes and supplied from a 125-volt, 65-cycle network, and it also includes a buffer battery. The latter ensures normal operation of the system during breakdowns in the power network for a period of several days. The application of transistors has made it possible to pro-

STAT

duce the whole apparatus on a very economical basis. This is also facilitated by the circumstance that the compandor, the carrier-frequency generator, and the transmission amplifier, comprised within the subscriber set, are fed with power only when the subscriber removes the handset. A subscriber telephone set consumes about 0.6 v during the period of "silence" and about 2 v during the actual conversation. The terminal equipment also consumes 2 v.

Designwise the apparatus is constructed in the form of a series of blocks with plug-type connectors. The application of transistors and printed circuits has made possible a substantial reduction in the dimensions and weight of the apparatus. The subscriber telephone set is designed in the form of two small cast-aluminum boxes which can be attached to a pole.

Communication and Electronics, No.24, 1956,
pp. 188-214

EXPERIMENTAL RADIO COMMUNICATION BY MEANS OF FORWARD SCATTER IONOSPHERIC PROPAGATION OF METER WAVES, IN THE USA

The recently discovered propagation of meter waves by means of scattered reflection from the ionosphere has made possible a considerable expansion of the frequency range of ionospheric propagation for radio communication purposes.

Ionospheric scattering of radio waves in the direction of their propagation makes it possible to effect radio communications in the 25 - 65 mc frequency band at distances of approximately 900 to 1900 km.

It has been found that scattered reflection of radio waves ensures more reliable radio communication at times of disturbances in the ionosphere and, especially, in the polar zones where the ionosphere is more often disturbed and where, therefore, short-wave radio communication is not so very reliable. Strengthening of scattered signals can occur during short-term disturbances of the ionosphere caused by solar flares, and this makes it possible to continue the interrupted short-wave radio com-

STAT

munication by means of meter waves.

As is known, meter waves usually penetrate through all ionospheric layers; however, owing to the irregularity of the lower part of the E layer, only a very small part of the energy of these waves is retained, scattered, and reflected to the earth in the direction of the directed propagation of radio waves.

According to measurements conducted along an experimental route from Anchorage to Barrow (Alaska), the gain of the scattered signal in the receiver is accompanied by an increase in magnetic activity in the midpoint of the experimental path, i.e., in the point of the reflection of waves from the ionosphere (above Fairbanks). That point is also marked by the greatest disturbance of the ionosphere to occur along the experimental path. This is favorable to meter-wave radio communication whenever short-wave communication is disrupted.

The applied meter-wave frequency range and the distance between the points of communication appear to be the principal factors restricting a wider utilization of ionospheric propagation of radio waves. The experiments conducted have revealed that signal power declines when frequency is increased above 60 mc. At too low a frequency, on the other hand, for example, at a frequency of less than 30 mc, the normal ionospheric propagation is also present, and this leads to interference between stations.

The distance of transmission is determined by the altitude of the ionosphere and the extent of the angle of incidence of radio waves in relation to the ionosphere, i.e., angle between the direction of wave propagation and the tangent to the ionosphere at the scattering point. Observations indicate that the scattering power of the signal declines sharply at an increase in the angle of incidence of radio waves - which occurs when the wave path is shortened. At a distance of less than 900 km this angle becomes large, the scattering of energy becomes small, and the signal becomes too weak and insufficient to effect radio communication.

STAT

At an increase in distance, signal power becomes sufficient. However, this occurs only so long as the part of the ionosphere in which wave reflection takes place remains "visible" to the transmitter. Increasing the antenna height will not increase greatly the range of action. However, at transmission from elevated shores of the ocean the range of action covered 2200 km.

The experiments were conducted along the following routes: Cedar Rapids (Iowa) - Sterling (Virginia), 1230 km; Fargo (North Dakota) - Churchill (Canada), Anchorage-Barrow (Alaska), St. John's (Newfoundland) - Terceira (The Azores), 2250 km; Cedar Rapids - Bermuda; and also Labrador - Greenland, and Maine - North Greenland. The last-named route has recently extended from Goose Bay (Labrador Peninsula) via Narsarssuak (Greenland) to Reykjavik (Iceland).

Considering the smallness of the scattered energy, the experiments were conducted with 40-kw transmitters, especially designed high-sensitivity receivers, large rhombic antennas and, for comparison's sake, Yagi arrays.

In 1951 the U.S. Air Force became interested in the use of scattered ionospheric propagation for radio communications in the Arctic. As of 1953 four-channel radio-teletype systems were set in operation for multiplex traffic. After a brief experimental period these lines were transferred for use by the Air Force. During the first year of their operation the lines were used 91% of all the time. Failures of radio waves to effect communication in these conditions totaled about one percent.

(Wire and Radio Communications,
Vol.74, No.3, 1956)

STAT

FIELD EXPERIMENTS WITH RURAL TELEPHONE INSTALLATIONS

As is known, considerable attention has been lately devoted to the question of utilizing solar batteries as power sources for rural telephone plants (see for example: Telecommunication Reports of 3 October 1555, Tele-Tech of August 1955, Electronics of February 1956, and Radio of July 1955). Some additional information on this problem is supplied below.

The first field experiments with telephone systems operated on transistors and solar batteries constructed by Bell Laboratories were conducted in Americus, Georgia, and yielded satisfactory results. The solar battery was installed in October of 1955 on only a part of the tested equipment. The purpose of the conducted experiments with supplanting conventional batteries by solar ones was to clarify the degree of usefulness of solar batteries as basic power sources for telephone plants, chiefly in the localities without other readily available power resources.

The conducted tests of the structure and use of solar batteries as devices for converting sunlight into electric power have met all expectations, technically speaking. The experiments revealed that from the viewpoint of dependability and technical efficiency of performance, the solar battery, which is usually mounted on a pole, can be utilized for ensuring the power supply of rural telephone plants. However, a widespread introduction of this battery is hampered by the high price of the chemically pure silicon used in its manufacture. Therefore, so long as that material remains expensive, it is more economical to use the existing normal current sources. Accordingly, Bell Laboratories has no plans for continuing its experiments with solar batteries in the immediate future.

As for transistors, they are of great importance to the development of media of communication. The experiments with new equipment based on transistors have corroborated the belief that these triodes are crucial to the development of future telephone systems.

The equipment tested in Americus contained about 275 transistors. Tests of this equipment have demonstrated its considerable technological superiority over the equipment based on electron tubes. Such a situation expands the opportunities for increasing the use of short-hand rural telephone lines and thus makes it possible to improve the servicing of rural population in the less populated regions.

STAT

A special discharger designed for protecting the low-power transistors from electric overvoltages, has been tested for the first time in the experimental installations of telephone devices. This discharger, termed "silicon-aluminum junction-type diode", when used together with the existing standard types of dischargers, protects transistors from lightning damage.

Bell Lab. Rec., No.6, 1956

PARABOLIC ANTENNA FOR STUDIES OF PROPAGATION BY MEANS OF SCATTER

The last few years have been marked by intensive studies of the feasibility of transmitting signals beyond the limits of the horizon. The experiments conducted indicate that a considerable part of microwave energy can be transmitted over the curved earth's surface, beyond the limits of the horizon. The mechanism making this possible has not yet been fully revealed, but the assumption is that scatter caused by irregularities of the atmosphere plays an important role here.

To study the nature of propagation by scattering, an American firm, Bell Telephone Laboratories, has constructed a 60-feet high microwave aluminum antenna at its laboratory in Holmdel, New Jersey. The compact surface of this antenna has the form of a paraboloid constructed with an accuracy of up to 3/15". The weight of the parabolloid amounts to 5.5 tons. The antenna together with its supporting beam can withstand a wind of 100 mph. Although designed to withstand a 1-inch thick load of snow or ice, this antenna has in practice withstood a 5-inch thick load of ice, and a wind of 150 mph.

The antenna is designed for operation on the frequencies of 460 - 4000 mc, but it has also been tested on a frequency of 9400 mc.

Calibrated pyramidal-form horns were used as the standard.

Antenna gain amounted to 37 db on a 460-mc frequency, 55 db on a 3890-mc frequency, and 61 db on a 9400-mc frequency.

Tests of the antenna on the highest frequency have demonstrated its good qualities not only with respect to its gain and beam width but also with respect to

STAT

clarity of its field pattern.

(Radio Television News , Vol.56,
No.10, 1956, p.102)

STAT

AUTHOR'S CERTIFICATES

Class 21a^1, 34$_{40}$. No.103595. A.A.Ionov. A Stereoscopic Television Method.

Class 21a^2, 18$_0$. No.103606. Ye.Ya.Yerstaf'yev. A Phototransistor Amplifier-Converter.

Class 21a^2, 18$_{08}$. No.103677. P.G.Tager. A Method for Pulsating Amplification of Electric Signals.

Class 21a^2, 34$_{01}$. No.103676. M.A.Sapozhkov and A.K.Lidikh. A Telephone Set for the Common-Battery System.

A proposal for joint use of the electromagnetic microphone and of an amplifier powered through speech wires from telephone-office common battery, for the purpose of increasing the power output of the microphone, expanding the opportunities for correction of the frequency response of the transmission ratio, and increasing the stability of the transmission level, in common-battery system telephone sets.

Class 21a^2, 36$_{02}$. No.103535. Ya.Yu.Ginzburg. A Call Distributor for Group Telephone Traffic Centers.

Class 21a^2, 36$_{02}$. No.103661. Yu.R.Gints. A Device for Effecting Both-Way Group Telephone Traffic Among Several Four-Wire Circuits.

Class 21a^3, 49$_{20}$. No.103427. A.N.Yuzhakov. A Device for Alarm Signaling through Telephone Lines.

Class 21a^4, 8$_{01}$. No.103658. G.B.Il'in. Three-Phase Vacuum-Tube Oscillator.

A proposal for using the feedback circuits of some tubes of the three-phase vacuum-tube oscillator formed by three electrically coupled oscillators to obtain driving voltage serving to effect a definite directing of phase rotation, low deviations of interphase angles at unbalanced loads, and elimination of autonomous oscillation of each arm.

STAT

Class 21a^4, 10. No.103594. Ye.G.Bronnikova and L.Z.Rusakov. Piezoelectric Resonator.

Class $21a^4$, 39. No.103621. L.Ye.Leykhter. Phase Magnetic Demodulator.

Class $21a^4$, 46_{01}. No.103500. A.V.Sokolov. Relay Stations of Radio Relay Lines.

A proposal for using a combined amplifying system to amplify the signals transmitted in both directions on various frequencies in the relay stations of radio relay lines, for the purpose of reducing the equipment used. The signals at the input and output of such an amplifying system would be separated by means of filters.

Class $21a^4$, 72_{04}. No.103460. M.A.Shkud, F.A.Gel'man, K.M.Ryabov, E.N.Kuchuk, O.I.Pakhomov, A.M.Ivashkin, and V.I.Fedorenko. An Antenna Switch.

Antenna switch for shifting current phases by 180° for the purpose of simplifying the design, and increasing the homogeneity and electric stability of feeder lines. The proposal is that the switch be constructed in the form of a section of a transposed four-wire feeder whose wires resemble tubular rods and whose fixed input contacts, to which the said rods are attached, are displaced 45° in relation to output contacts so that a 45° turn of the rods entails a 180° change of the phase of the current.

Class 21d, 25. No.103578. V.Ye.Savchenko. A Device for Measuring Insulation Resistance.

Class 21e, 11_{20}. No.103447. B.A.Barskiy and V.A.Volosevich. A Device for Harmonic Analysis.

Class 21e, 28_{02}. No.103417. N.A.Pikulev. A Device for Simultaneous Recording of Several Processes on the Screen of an Electron Oscillograph.

Class 21e, 29_{01}. No.103519. V.N.Nikol'skiy. An AC Ammeter.

Class 21e, 31. No.103622. R.S.Medvedeva. A Method for Measuring the Conversion Transconductance of Radio Tubes by the Zero Frequency Method.

A proposal for using the zero frequency method to measure the conversion transconductance of radio tubes by means of two synchronously acting switches, one to shift the phase of the signal-grid voltage and the other to alter the direction of the flow of plate current through the device, which may be directly measured in

STAT

units of transconductance conversion.

FOREIGN PATENTS

Editor's Note: In publishing here brief mentions of some foreign patents concerning television technology the Editors wish to add for the information of the readers that photostats of patents (in the original language, with designs) are available upon applying person or mailing c.o.d. orders to the All-Union Technical Patent Library of the Committee for Inventions and Discoveries, Council of Ministers of the USSR.

Address of the library: Moscow-Center, proyezd Serova, 4, pod'yezd 7a. Phone B8-64-52, for information.

Photostats of patents can be also obtained through the Purchasing Office of the All-Union Institute for Scientific and Technical Information, serving the USSR Academy of Sciences and the Gostekhnika State Engineering Institution.

Address of the Institute: Moscow, D-219, Baltiyskiy pos., d.42 b. Phone D 7-00-10, Ext.51.

Patent, German Federal Republic. Cl.21a, 32/22. No.759 354, 11.08.55. A Television Pickup Tube. (Opta Radio Akt. Ges.) (Einrichtung mit Bildfaengerroehre).

A proposal for an iconoscope-type tube device at which the electron beam is directed perpendicularly to the surface of the mosaic. The axis of the electron gun is directed parallel to the mosaic plane, but after scanning the electron beam turns perpendicularly toward it by means of an electromagnetic or electrostatic device. The application of such a tube makes it possible to dispense with the devices for compensating keystone distortions in ordinary iconoscopes. The above-described procedure is also suitable for receiving and, especially, projection tubes. Here the location of the electron gun at a side of the screen makes it possible to utilize its brighter-illuminated side; a further increase in screen efficiency is achieved by mounting the luminophor on a reflecting surface.

U.S.Patent, Cl. 178-7:5, No.2, 717,920, 13,09.55. Avins Jack (Radio Corp. of

STAT

America). Noise Cancellation Circuit.

A circuit designed for improving the pulse-noiseproof feature of synchro circuits and AGC in television receivers. The noiseproof feature is ensured by connecting a noise inverter stage between the input and output of the video amplifier. The normally-closed inverter stage opens when the noise exceeds the synchro pulse level. Here, the noise voltage on the video-amplifier load decreases (or vanishes) because the inverter-amplified noise voltage which is opposed in phase to the noise voltage amplified by the video amplifier, develops on the said load and cancels the latter voltage.

The amplitude of pulse noise may either be smaller than the magnitude of the cut-off bias of the inverter stage or may exceed slightly or considerably the latter, depending on the extent of opening of the inverter tube. In the former case, the inverter does not operate; however, the energy of pulse noise is relatively small and therefore the signal-to-noise ratio in the synchro channel is not much affected by that noise. In the latter case, the inverter tube should have a high grid through if the inverter is to have a maximum efficiency; further, when the tube is opened to its fullest extent, the inverter efficacy is maximal and cancels noise completely.

The noiseproof feature of the AGC circuit is additionally increased by connecting the AGC tube grid to the plate circuit of the inverter.

Australian Patent No.165,582, 27.10.55. (Radio Corp. of America). Color Television Amplifier.

A spurious capacitance coupling exists between the plates of various color components in the multicolor transmission tube. To eliminate this coupling, negative feedback of capacitance character is applied in the input amplifiers of the video signal of each color component to reduce the influence of spurious capacitances on a wide frequency range.

U.S.Patent, Cl. 178-5:4, No.2,725,418, 29.11.55. Sziklai, George (Radio Corp. of America). Color Television Receiving System.

STAT

Proposal for a receiving device for color television systems of the point-contact successive-transmission type; a receiving device that can be applied together with reproducing devices consisting of a single three-color kinescope with one or three beams, or of three separate kinescopes. The receiving device ensures the performance of the color-reproducing devices synchronously with the emission of color signals. For this purpose, use is made of a generator of three sinusoidal oscillations with a 120°-phase shift, controlling the three signal dissectors which, in turn, control three pilot lamps. These lamps release the necessary electron beam (or one beam at the required moments of time) at the moment of the emission of corresponding color signals. Further, a single video amplifier is used for all three color video signals.

U.S.Patent, Cl. 178-5:4, No.2,706,217, 12.04.55. Roland N. Rhodes and Allen A.Barco (Radio Corp. of America). Color Television Control Apparatus.

Proposal for a method of emitting the useful product of modulation out of the overall signal of several modulators of the color subcarrier in color television systems with single-frequency-band transmission of the brightness signal and of the subcarrier signal containing color information in the form of amplitude and phase modulation.

Specific instance: obtainment of a color subcarrier signal by means of amplitude modulation of three oscillations with subcarrier frequency and at a 120°-phase shift. In this case, when the output signals of the three modulators are combined, the subcarrier is suppressed. In order to suppress the modulating signal itself it is proposed to supply a reverse-polarity modulating signal of a corresponding amplitude to the overall output of the modulators. This makes unnecessary the application of balanced modulators and filters.

Canadian Patent No.515,333, 02.08.55. Cetsworth, Albert III. A Television Receiver with Sound Reception by Means of Beats between Carrier Frequencies (Zenith Radio Corp.). (Intercarrier Television Receiver).

Proposal for a circuit for separating the audio and video signals in a television receiver in which beats between carrier frequencies are used for reception of audio signals. The signal is conveyed from the detector output to a separation pentode; and a resistance whose part is capacitance-shunted and a resonance circuit tuned to a frequency equal to the difference between carriers are (the resistance and the resonance circuit) series-connected to the plate circuit of that pentode. The signal enters the audio channel from the resonance circuit, which is inductively coupled to the plate circuit, and it enters the video amplifier - from the resistance. A capacitor is connected between the point of connection of the video amplifier and the second circuit, and this capacitor is used for introducing the differential-frequency signal at an amplitude and phase sufficient to compensate the component of the differential frequency falling from resistance into the video amplifier.

West German Patent, Cl. 21a, 33/10, No.933,871, 06.10.55. Barthelemy, R. (Co. pour la Fabrication des Compteurs et Materiel d'Usines a Gaz). A Device for the Operation of Television Pickup Tubes. (Anordnung zum Betrieb von Fernsehzerlegern.)

To simplify the design of television equipment and eliminate the low-frequency noise usually appearing in the video-signal amplifying path, it is proposed that HF oscillations be modulated by video signals directly in the pickup tube. This may be done only in tubes performing with beams of low-velocity electrons, for instance, in the super-opticon. The HF generator of frequency f is connected subsequently together with a source of constant and negative (in relation to cathode) voltage delivered to the tube's Wehnelt cylinder. In this way, the electron beam becomes modulated by high frequency. The HF voltage, modulated by video signals, is taken from the signal plate by means of a wideband HF transformer tuned to frequency f.

U.S.Patent, Cl. 178-5:8, No.2,706,218, 15.04.55. Wooten, W.A. System for Television-Program Film Recording and Record Reproduction.

Proposal for a system of film-recording of programs being televised. In customary program-transmission conditions it is necessary to use 3 to 5 television cameras

STAT

positioned in various sides of the scene and switched on during transmission time. Each of these cameras can be operated jointly with a movie camera, and the resulting film record of the television program is compiled from fragments shot by individual cameras, which is both complicated and expensive. The proposed system suggests such a joint operation of television and movie cameras but without the factors that make it so complicated and expensive. The moments of the switching of cameras automatically punch holes into the control film tape which moves in time with the working movie camera. The negative tapes shot by the several movie cameras appear and then are processed through the synchronized printing heads which are so designed as to print a single positive tape. The control tape serves to control the printing of film on the positive tape in closely corresponding relation to the changes of cameras during the shooting of the scene. A complete film record of the program can thus be obtained in a simpler manner.

The device for punching holes into control tape consists of die punches and solenoids operated by push-button control from the program producer's desk. Sound recording is likewise ensured by concurrent operation of movie cameras and sound recorders, with changes in cameras signaled by strobing of points on the sound tape. Electrical circuits and appropriate devices (solenoids, control drum, etc.) ensure the automation of the process.

STAT

BOOKS TO BE ISSUED IN 1957

Concise information is cited below concerning the plans of some publishing houses for the publication, in 1957, of technical literature that would be, in the Editor's opinion, of interest to the readers of Elektrosvyaz'.

Sovetskoye Radio Publishing House

Arenberg, A.G. - Propagation of Decimeter and Microwaves. Edition: 10,000 copies. Price: 16 rubles. To be issued in first quarter of 1957

Neyman, M.S. - A Course in Radio Transmission Devices. Edition: 25,000 copies. Price: 6 rubles. To be issued in second quarter of 1957

Rizkin, A.A. - Fundamentals of the Theory of Amplifying Circuits. Third Edition: 25,000 copies. Price: 10 rubles. To be issued in last quarter of 1957

Mehl and Gerhard - Decimeter-Wave Engineering (translated from the German, under the direction of N.K.Svistov). Edition: 10,000 copies. Price: 18 rubles 50 kopecks. To be issued in third quarter of 1957

Beck, A. - Electron Tubes: Theory and Design (translated from the English, under the direction of L.A.Kotomina). Edition: 10,000 copies. Price: 18 rubles 50 kopecks. To be issued in fourth quarter of 1957

Lo, Endres, Zawels, Waldhauer, and Cheng - Fundamentals of Transistor Electronics (translated from the English, under the direction of E.I.Gal'perin). Edition: 10,000 copies. Price: 27 rubles. To be issued in third quarter

- Problems of Long-Distance Ultrashort-Wave Communications, edited by V.I.Siforov. Edition: 10,000 copies. Price 15 rubles. To be issued in first quarter.

Levin, B.R. - The Random Process Theory and Its Applications in Radio Engineering. Edition: 10,000 copies. Price: 12 rubles 50 kopecks. To be issued in first quarter.

STAT

10(

Poletayev, I.A. - The Signal: Concerning Some Cybernetic Concepts. Edition: 50,000 copies. Price 7 rubles. To be issued in fourth quarter

Wiener, N. - Cybernetics. Edition: 10,000 copies. Price: 9 rubles. To be issued in third quarter.

Govorkov, V.A. and Kupalyan, S.D. - Exercises on the Electromagnetic Field Theory. Edition: 10,000 copies. Price: 11 rubles. To be issued in second quarter.

Shestakov, M.F. - Bibliographical Index of Literature on Undergraduate and Graduate Designing of Radio Transmission Devices. Edition: 10,000 copies. Price: 1 ruble 50 kopecks. To be issued in first quarter.

Fedotov, Ya.A. - Instead of the Radio Tube (Technology in the Sixth Five-Year Plan Series). Edition: 50,000 copies. Price: 1 ruble. To be issued in first quarter.

Konev, Yu.I. - Transistors in Automatic Control Devices. Edition: 10,000 copies. Price: 5 rubles. To be issued in first quarter.

State Publishing House for Foreign Literature

Goldman, S. - Information Theory (translated from the English). Price: 15 rubles 50 kopecks. To be issued in the first quarter.

Rayt, D. - Transistors (translated from the English). Price: 8 rubles. To be issued in third quarter.

Shea, R. - Transistor Audio Amplifiers (translated from the English). Price: 12 rubles 50 kopecks. To be issued in second quarter.

Kayver, M. - Fundamentals of Color Television (translated from the English). Price: 13 rubles 50 kopecks. To be issued in first quarter.

- Third International London Symposium on the Information Theory (translated from the English). Price: 19 rubles 50 kopecks. To be issued in second quarter.

State Publishing House for Theoretical Engineering Literature

Kharkevich, A.A. - Theoretical Foundations of Radio Communications. Edition: 50,000 copies. Price: 6 rubles 40 kopecks. To be issued in second quarter.

Parfent'yev, A.N. and Pussep, L.A. - Physical Foundations of Magnetic Recording of Sound. Edition: 10,000 copies. Price: 11 rubles. To be issued in third quarter.

Sheftel', I.T. - Thermosensitive Transistor Resistances. Edition: 10,000 copies. Price: 3 rubles. To be issued in fourth quarter.

German-Prozorova, L.P. and Vinogradova, N.I. - English Russian Radio Engineering Dictionary. Edition: 50,000 copies. Price 15 rubles 50 kopecks. To be issued in third quarter.

The dictionary contains about 20,000 terms concerning radio engineering and broadcasting, television, receiving and transmitting devices, and electronics.

Geyler, L.B. and Dozorov, N.I. - English Russian Dictionary of Electrical Engineering. Second (enlarged) edition: 35,000 copies. Price: 23 rubles 50 kopecks. To be issued in first quarter.

This dictionary contains about 40,000 terms concerning: Power generation and industry, construction of electrical machinery and instruments, wire communications, and radio.

Team of authors: German-Russian Dictionary of Electrical Engineering. Edition: 35,000 copies. Price: 21 rubles 50 kopecks. To be issued in third quarter

This dictionary contains about 35,000 terms concerning power stations, electrical machinery and instrument construction, other branches of industrial applications of power, wire and radio communications, and automation.

STAT

NEW BOOKS

Radovskiy, M.I. - Publishing House of the USSR Academy of Sciences, 1956. Edition: 10,000 copies. 207 pages. Price: 3 rubles 10 kopecks.

Declassified in Part - Sanitized Copy Approved for Release @ 50-Yr 2013/09/06 : CIA-RDP81-01043R002500080007-5

A bibliographical sketch ("Scientific-Popular Library").

Tsyan' Syue-Sen' - Engineering Cybernetics. Translated from the English by M.Z.Litwin-Sedoy. Edited by A.A.Fel'dbaum. Izdatel'stvo inostrannoy literatury Publishing House, 1956. 462 pages. Price in hard cover: 17 rubles 5 kopecks.

Gubanov, A.I. - Theory of the Rectifying Action of Transistors. Gostekhizdat Publishing House, 1956. 348 pages. Edition: 10,000 copies. Price in hard cover: 11 rubles 15 kopecks.

- Electrophysical Properties of Germanium and Silicon. A compendium of translations, edited by A.V.Zhdanov. Sovetskoye Radio Publishing House, 1956. 392 pages. Price in hard cover: 17 rubles.

Bonch-Bruyevich, M.A. - Collected Works. Publishing House of the USSR Academy of Sciences, 1956. 526 pages. Edition: 3500 copies. Price in hard cover: 30 rubles 70 kopecks.

Malkin, I.G. - Some Tasks of the Theory of Nonlinear Oscillations. Gostekhizdat Publishing House, 1956. 491 pages. Edition: 6000 copies. Price in hard cover: 16 rubles.

Kaplanov, M.R. and Levin, V.A. - Automatic Frequency Control. Second (enlarged) edition. Gosenergoizdat Publishing House, 1956. 200 pages. Edition: 12,000 copies. Price in hard cover: 11 rubles 50 kopecks.

Ivanov, A.B. and Sosnovkin, L.N. - Superhigh -Frequency Pulse Transmitters. Sovetskoye Radio Publishing House, 1956. 616 pages. Price in hard cover: 13 rubles 85 kopecks.

Andreyevskiy, M.N. - Design of Generators of Decimeter and Microwaves. Oborongiz Publishers, 1956. 132 pages. Hard-cover price: 5r 35 k.

Volin, M.L. - IF Amplifiers. Third, enlarged, edition. Sovetskoye Radio Publishing House, 1956. 232 pages. Price in hard cover: 7 rubles 20 kopecks.

- New Radio Broadcasting Literature. A Compendium, Svyaz'izdat Publishing House, 1956, 108 pages + 5 insets. Edition: 12,500 copies. Price: 3 rubles 60 kopecks.

Gertsenshteyn, B.Ya. and Savina, N.A. - Fundamentals of the Theory and Computing of Wire Broadcasting Lines. Svyaz'izdat Publishing House, 1956. 372 pages. Edition: 10,000 copies. Price in hard cover: 12 rubles.

Brandt, A.A. - Techniques for Assembling and Adjustment of Radio Circuits. Moscow University Press, 1956. 247 pages. Edition: 25,000 copies. Price in hard cover: 6 rubles.

Vostroknutov, N.G. - Technology of the Measurements of Electrical and Magnetic Values. Second, revised edition. Gosenergoizdat Publishing House, 1956. 440 pages. Edition: 15,000 copies. Price in hard cover: 9 rubles 75 kopecks.

A textbook for the secondary schools of engineering which instruct workers in measurement techniques.

Nazarov, M.V. - Channel Capacity of Multiplex Traffic Systems. Svyaz'izdat Publishing House, 1956. 66 pages. Edition: 10,000 copies. Price: 1 ruble 85 kopecks.

A brochure, one in the series of "Lessons in Communications Engineering".

Vitenberg, M.I. - Computations of Electromagnetic Relays used in Automation and Communications. Second, revised and enlarged edition. Gosenergoizdat Publishing House, 1956. 464 pages. Edition: 10,000 copies. Price in hard cover: 14 rubles 50 kopecks.

- New Dial Telephone Systems. A compendium of translated articles, edited by L.S.Farafonov, Svyaz'izdat Publishing House, 1956. 100 pages. Edition: 5500 copies. Price: 5 rubles 70 kopecks.

Babitskiy, I.A. - Rating of Step-by-Step Dial Systems. Svyaz'izdat Publishing House, 1956. 32 pages. Edition: 7000 copies. Price: 80 kopecks.

A brochure in the series of "Lessons in Communications Engineering

Kuznetsov,Ye.K. - Telephone Sets. Svyaz'izdat, 1956. 292 pages. Edition: 5000

STAT

copies. Price in hard cover: 10 rubles 10 kopecks.

Koptev, I.V. - Theory of Aerial Lines. Svyaz'izdat, 1956. 282 pages. Edition: 10,000 copies. Price in hard cover: 7 rubles.

A student manual for institutes of electrical communications engineering.

Grodnev, I.I., Efimov, I.Ye. and Marimont, L.B. - Electrocommunication Lines. Voyenizdat, 1956. 504 pages. Price in hard cover: 11 rubles 65 kopecks.

A textbook for military communications schools.

Buchenkov, A.N. - Radio Electronics: Its Achievements and Perspectives of Development. Review of literature, for guidance purposes, 1956. 16 pages. Edition: 20,000 copies. Price: 30 kopecks.

Levshina, O.N. - Transistors and Their Applications in Science and Engineering. A review of literature, for guidance purposes, 1956. 14 pages. Edition: 20,000 copies. Price: 30 kopecks.

The above two brochures are the first issued in a series of bibliographic reviews for guidance purposes initiated under the overall name of "Novosti tekhniki" Engineering News by the V.I.Lenin USSR State Library in collaboration with the Central Polytechnical Library.

STAT