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THE NOISEPROOF FEATURE OF RECEIVERS WITH END RECOVERY TIME

by

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An equation is discussed for determining the degree of probability of signal reception at the instant of time t in a receiver with end recovery time. With this, the given degree of probability of appearance of a signal in the receiver input is computed. A solution for the equation is found for the case when the intensity of interference is low. This solution permits a determination of the relation between the number of correctly received signals and the number of spurious signals in the receiver with end recovery time.

Introduction

We will discuss the pulse communication line which has been subjected to the effects of interference. Pulse signals are transmitted along the line. To lessen the harmful effects of interference, the receiver may be cut off at the time t after reception of the next pulse. If the signals are transmitted periodically, the presence of end recovery time t reduces the probability of emergent false signals during the interval between transmissions. At the same time, the probability of receiving the transmitted signal is diminished due to the fact that the receiver being cut off at the instant of its arrival. In the receiver with end recovery time, the ratio of the number of the received correct signals to the spurious signals is larger than the corresponding ratio in a conventional receiver.

If the pulse receiver is not cut off artificially, it will have end recovery time. This time remains quite small in the vast majority of cases and can be disregarded in practice. However, when using a relaxation oscillator at the receiver

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output or when receiving strong signals which overload the receiver, the end recovery time must be taken into account.

It is of interest to define the probability $p(t)dt$ of receiving a signal in the time interval from t to $t + dt$, considering as given the probability $\pi(t)dt$ of the arrival of the signal at the receiver input during the time interval from t to $t + dt$. Below, the arrival of a signal at the receiver input, during the time interval adjacent to $t + dt$, will be considered an independent event. This corresponds to an approximation when "the infinitely small" interval dt is denoted as the time interval of the pulse duration series, and $\pi(t)dt$ as the probability of the appearance of a signal for an interval of time approaching the point t.

Equation for the Degree of Probability of Signal Reception

We will now consider the equation for the degree of probability $p(t)$ of signal reception in the receiver with end recovery time τ. The probability of signal reception in the interval $t - t + dt$ is equal to the probability of signal reception at the receiver input in the interval $\pi(t)dt$, multiplied by the probability that the receiver is in recovery condition. It is obvious that the receiver will be closed if the signal is received at any of the instants of time ξ, lying in the interval $t - \tau < \xi < t$. The probability that the signal is received for the time $t - \tau$ to t is equal to $\int_{t-\tau}^{t} p(\xi)d\xi$. Therefore the probability that the receiver is in recovery condition at the instant t is equal to $1 - \int_{t-\tau}^{t} p(\xi)d\xi$. It is obvious that the probability $p(t)dt$ may be represented as the product $\pi(t)dt\,[1 - \int_{t-\tau}^{t} p(\xi)d\xi]$.

Equating both expressions and reducing by dt, we get the desired equation

$$p(t) = \pi(t)\left[1 - \int_{t-\tau}^{t} p(\xi)\,d\xi\right]. \qquad (1)$$

If the signal received is transmitted periodically at fixed period T and interference, then the degree of probability of signal appearance at the receiver output $\pi(t)$ has the constant value π for all instants of time when the signal is not being transmitted and is rejected during the instants of time $t = kT$ when the signal is being transmitted. These rejects correspond to the fact that, at these instants, the probability of signal appearance at the receiver input increases sharply. If $\delta(t)$ denotes the delta function, determined by the conditions

$$\delta(t) = \begin{Bmatrix} 0 & t \neq 0 \\ \infty & t = 0 \end{Bmatrix}, \int_{-\infty}^{+\infty} \delta(t)\,dt = 1,$$

then, for $\pi(t)$ in the case under consideration, the following analysis may be written:

$$\pi(t) = \pi + \sum_k q\delta(t - kT), \qquad (2)$$

where q is the probability of the appearance of a transmitted signal at the receiver input in the presence of interference. Thus the determination of the degree of probability of signal reception at the instant of time t, if a periodic pulse signal is transmitted and the receiver is affected by a fixed interference, reduces to the derivation of eq.(1) when $\pi(t)$ is given by eq.(2). If reception begins long before the instant of time under consideration, the solution of eq.(1) is hardly distinguishable from the fixed derivation, i.e., from the periodic solution of eq.(1), with $\pi(t)$ as given by $-\infty < t < +\infty$. In the following, only the periodic solution of the equation will be considered.

Solution of the Equation for Low-Intensity Interference

The solution of the equation under ordinary circumstances is far from simple. In this connection, the case below which the intensity of interference is low will

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be considered. Interference may be considered low if $\pi\tau \ll 1$, i.e., if the probability of a false signal arriving at the time τ is much less than unity. It is natural to limit the calculation to the case where the recovery time of the receiver τ is less than the period of signal transmission T, since otherwise, even without interference, not all of the transmitted signals could be picked up by the receiver.

We will represent the degree of probability $\pi(t)$ as $\pi(t) = \pi + \pi_o$, where $\pi_o = \sum_k q\,\delta(t - kT)$, and the degree of probability $p(t)$ represents the analysis $p(t) = p_o + p_1 + p_2 + \ldots$, where p_n is a member of the series n conforming to $\pi\tau$. If these terms in eq.(1) are substituted and the members of one finite series equated, then a system of equations is obtained which permits the determination of all p_n:

$$\left.\begin{aligned}
p_0 &= \pi_0 \left| 1 - \int_{t-\tau}^{t} p_0(\xi)\,d\xi \right| \\
p_1 &= \pi \left[1 - \int_{t-\tau}^{t} p_0(\xi)\,d\xi \right] - \pi_0 \int_{t-\tau}^{t} p_1(\xi)\,d\xi \\
p_2 &= -\pi \int_{t-\tau}^{t} p_1(\xi)\,d\xi - \pi_0 \int_{t-\tau}^{t} p_2(\xi)\,d\xi \\
&\cdots\cdots\cdots\cdots\cdots\cdots \\
p_n &= -\pi \int_{t-\tau}^{t} p_{n-1}(\xi)\,d\xi - \pi_0 \int_{t-\tau}^{t} p_n(\xi)\,d\xi.
\end{aligned}\right| \qquad (3)$$

The first equation of the system (3) corresponds to zero approximation, i.e., to the case where π equals zero. Only at $t \neq kT$ $\pi_o = 0$ does it follow that $p_o = 0$ at $t \neq kT$. At the points $t = kT\,\pi_o = q\,\delta(t - kT)_{t = kT}$, and the interval $\int_{kT-\tau}^{kT} p_o(\xi)d\xi = 0$, since it takes into its scope where $p_o(\xi) = 0$ $(\xi \neq kt)$. From this, it is obvious that $p_o = \pi_o$.

A study of the second equation of the system (3) indicates clearly that it corresponds, in first approximation, to $\pi\tau$. In following this through as $p_o = \pi_o$, this

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takes the form of $p_1(t)=\pi\left[1-\int\limits_{t-\tau}^{t}\pi_0(\xi)\,d\xi\right]-\pi_0\int\limits_{t-\tau}^{t}p_1(\xi)\,d\xi$, at $t \neq kT$, $\pi_0 = 0$ and, consequently,

$$p_1(t)=\pi\left[1-\int\limits_{t-\tau}^{t}\pi_0(\xi)\,d\xi\right]=\pi-\sum_k \pi q \; direct\left(\frac{t-kT-\frac{\tau}{2}}{\tau}\right), \tag{4}$$

where, in conformity with the symbols introduced previously (Bibl.1), we use

$$direct\;(t)=\begin{Bmatrix}1 & |t|<\frac{1}{2}\\ 0 & |t|>\frac{1}{2}\end{Bmatrix}.$$

For the points $t = kT$, we may now write

$$p_1(kT)=\pi\left[1-\int\limits_{kT-\tau}^{kT}\pi_0(\xi)\,d\xi\right]-q\delta(t-kT)_{t=kT}\int\limits_{kT-\tau}^{kT}p_1(\xi)\,d\xi$$

or, neglecting the first term on the right-hand side versus infinity, as the value of the second term and considering eq.(4),

$$p_1(kT)=-q\left[\pi\tau-\pi q\int\limits_{kT-\tau}^{kT} direct\left(\frac{\xi-kT-\frac{\tau}{2}}{\tau}\right)d\xi\right]\delta(t-kT)_{t=kT}.$$

Let $\tau > \frac{T}{2}$; then,

$$dtp_1(kT)=-q\left[\pi\tau-\pi q(2\tau-T)\right]=-q\pi\left[T-(2q-1)\tau\right].$$

Finally we obtain for $p_1(t)$

$$p_1(t)=\pi-\sum_k \pi q \; direct\left(\frac{t-kT-\frac{\tau}{2}}{\tau}\right)-\sum_k \pi q\left[T-(2q-1)\tau\right]\delta(t-kT). \tag{5}$$

The following approximations may be derived in an analogous manner. It can be

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demonstrated that, if $\pi\tau < 1$, then the series $p(t) = p_0 + p_1 + p_2 + \dots$ converges uniformly and, consequently, represents the solution of eq.(1). We are limiting the calculation to the solution in first approximation which, according to the discussed principles, has the form

$$p_0 + p_1 = \sum_k q\{1 - \pi[T - (2q-1)\tau]\}\,\delta(t - kT) + \pi - \sum_k \pi q \;\mathrm{direct}\left(\frac{t - kT - \frac{\tau}{2}}{\tau}\right). \tag{6}$$

The Relationship of the Number of Correct Signals Received to the Number of Spurious Signals

The electronic receiver admits all signals arriving at its input. Therefore, if the degree of probability of the arrival of signals at its input is given in eq.(2), then the relation of the number of correctly received signals to the number of spurious signal is equal to $\frac{q}{\pi T}$.

In the receiver with end recovery time, the degree of probability of signal reception, in first approximation, is described by eq.(6). The probability of reception of a transmitted signal, according to this formula, equals

$$q\{1 - \pi[T - (2q-1)\tau]\}.$$

The mean number of spurious signals picked up by the receiver for a period T, is equal to

$$\int_{kT-T+0}^{kT-0} (p_0 + p_1)\,d\xi = \pi\tau(1-q) + \pi(T - \tau) = \pi T\left[1 - q\frac{\tau}{T}\right].$$

Whence the desired relationship in the receiver with end recovery time will equal

$$\frac{q}{\pi T}\,\frac{\{1 - \pi[T - (2q-1)\tau]\}}{\left[1 - q\frac{\tau}{T}\right]}$$

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and is distinguished from the corresponding relationship in the electronic receiver by the common multiple

$$d = \frac{\{1 - \pi[T - (2q - 1)\tau]\}}{\left|1 - q\frac{\tau}{T}\right|}.$$

For example, if q = 0.95, τ = 0.8 T, and πT = 0.1, then d = 4. The mean number of correct signals in the electronic receiver in the case under discussion is 95%, whereas in the receiver with end recovery time, it is $q\{1 - \pi[T - (2q - 1)\tau]\}$, i.e., 93%. It is clear from this that, in the given case, the end recovery time results in practically no change in the number of correctly received signals, but leads to a quadruple reduction in the number of spurious signals.

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GENERALIZED ANALYSIS OF AMPLIFIER STAGES

by

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Generalized systems and generalized matrixes for various types of tube and semiconductor stages are presented. The application of the theory to the analysis of equivalent systems of semiconductor stages at high frequencies is demonstrated.

Generalized Systems and Generalized Matrixes

The analysis of amplifier stages (tubes or semiconductors) in a linear hookup is ordinarily made by constructing an equivalent corresponding system in the form of a linearly operating quadripole, containing dependent sources of current or voltage. For the analysis of complex multistage systems, the matrix method is especially well suited.

In addition to this, it is desirable to obtain a separate stage which will be adaptable to stages of various types. For example, in the case of stages in electron tubes, it is desirable to have a generalized matrix, uniformly suited to stages with a common cathode, grid, or anode; and in the case of stages in semiconductor triodes, to stages with a common emitter, base, or collector.

This can be easily attained if the generalized equivalent systems of amplifier stages with double dependent sources are introduced into the types presented by the author (Bibl.1). One of the possible variants of the generalized equivalent circuit of the tube stage is given in Fig.1, and the semiconductor stage in Fig.2. The transition from the generalized stage systems to the concrete types has been carried out in conformity with the data of Tables 1 and 2.

The generalized equivalent systems permit the corresponding generalized matrixes

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to be obtained. Thus, the system of equations for the voltage nodes of the circuit in Fig.1 directly determines the matrix [y] of this system, whereas the system of equations for the current circuit of the diagram in Fig.2 is the generalized matrix [z] of a semiconductor stage. The generalized matrixes of other types are obtained from these matrixes by the known conversion formulas (Bibl.2).

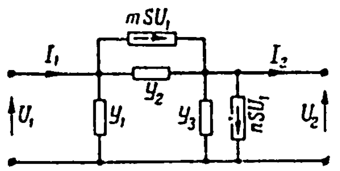

Fig.1

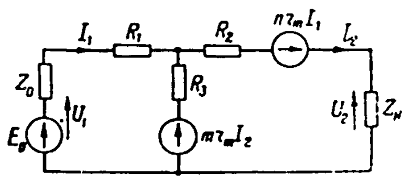

Fig.2

The generalized matrixes of tube or semiconductor stages obtained in this manner are given in Tables 3 and 4. To obtain the matrixes of the concrete stages it is sufficient to substitute, for the corresponding generalized matrix, the values conforming to Tables 1 and 2.

According to the known matrixes of the individual stages, the matrixes of the complex systems are determined. These permit obtaining all values which characterize the operation of the system.

If a source with an emf of E_o and an internal resistance of Z_o operates at the input of the system, while the output of the system is closed at the resistance Z_n (Fig.3), then the amplification of the system through the voltage K_i, the amplification of the system through the current K_j, the input resistance of the system Z_{in} and its output resistance Z_{out} can easily be found according to the formula presented in Table 5 (Bibl.2).

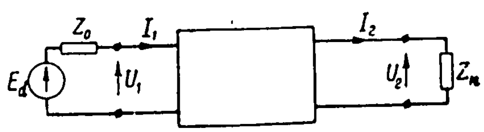

Fig.3

In compiling the Tables for positively directed currents and voltages, the directions shown in the diagrams were assumed, whereas the matrix specifications correspond to the following matrix equations:

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = [y] \begin{bmatrix} U_1 \\ U_2 \end{bmatrix}; \quad \begin{bmatrix} U_1 \\ U_2 \end{bmatrix} = [z] \begin{bmatrix} I_1 \\ I_2 \end{bmatrix};$$

$$\begin{bmatrix} U_1 \\ I_2 \end{bmatrix} = [d] \begin{bmatrix} U_2 \\ I_1 \end{bmatrix}; \quad \begin{bmatrix} U_2 \\ I_1 \end{bmatrix} = [f] \begin{bmatrix} U_1 \\ I_2 \end{bmatrix}; \quad \begin{bmatrix} U_1 \\ I_1 \end{bmatrix} = [a] \begin{bmatrix} U_2 \\ I_2 \end{bmatrix}.$$

Table 1

Generalized Matrixes of the Tube Stage

	a)	Y_1	Y_2	Y_3	m	n
1	b)	$Y_{g\kappa}$	Y_{ag}	$Y_{a\kappa}$	0	1
2	c)	$Y_{g\kappa}$	$Y_{a\kappa}$	Y_{ag}	1	0
3	d)	Y_{ag}	$Y_{g\kappa}$	$Y_{a\kappa} + S$	0	—1

a) Stage denomination; b) With common cathode; c) With common grid; d) With common anode

Table 2

Generalized Matrixes of the Semiconductor Stage

	a)	R_1	R_2	R_3	m	n
1	b)	r_v	$r_\kappa - r_m$	r_e	0	—1
2	c)	r_e	r_κ	r_v	0	1
3	d)	r_v	r_e	r_κ	1	0

a) Stage denomination; b) With common emitter; c) With common base; d) With common collector

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Table 3

Generalized Matrixes of the Pi-Network

$[y]$	$\begin{bmatrix} Y_1+Y_2+mS & -Y_2 \\ (m-n)S+Y_2 & -(Y_2+Y_3) \end{bmatrix}$
$[z]$	$\frac{1}{\lvert y \rvert}\begin{bmatrix} -(Y_2+Y_3) & Y_2 \\ (n-m)S-Y_2 & Y_1+Y_2+mS \end{bmatrix}$
$[d]$	$\frac{1}{Y_1+Y_2+mS}\begin{bmatrix} Y_2 & 1 \\ \lvert y \rvert & (m-n)S+Y_2 \end{bmatrix}$
$[f]$	$\frac{-1}{Y_2+Y_3}\begin{bmatrix} (n-m)S-Y_2 & 1 \\ \lvert y \rvert & -Y_2 \end{bmatrix}$
$[a]$	$\frac{1}{(m-n)S+Y_2}\begin{bmatrix} Y_2+Y_3 & 1 \\ -\lvert y \rvert & Y_1+Y_2+mS \end{bmatrix}.$

$$\lvert y \rvert = [Y_1(Y_2+Y_3)+nSY_2+(Y_2+mS)Y_3] = $$
$$= [Y_{ag}(S+Y_{gk}+Y_{ak})+Y_{gk}Y_{ak}].$$

Table 4

Generalized Matrix of the T-Network

$[z]$	$\begin{bmatrix} R_1+R_3 & -R_3+mr_m \\ R_3+nr_m & -R_2-R_3+mr_m \end{bmatrix}$
$[y]$	$\frac{1}{\lvert z \rvert}\begin{bmatrix} -R_2-R_3+mr_m & R_3-mr_m \\ -R_3-nr_m & R_1+R_3 \end{bmatrix}$
$[d]$	$\frac{-1}{R_2+R_3-mr_m}\begin{bmatrix} -R_3+mr_m & \lvert z \rvert \\ 1 & -R_3-nr_m \end{bmatrix}$
$[f]$	$\frac{1}{R_1+R_3}\begin{bmatrix} R_3+nr_m & \lvert z \rvert \\ 1 & R_3-mr_m \end{bmatrix}$
$[a]$	$\frac{1}{R_3+nr_m}\begin{bmatrix} R_1+R_3 & -\lvert z \rvert \\ 1 & R_2+R_3-mr_m \end{bmatrix}$

$$-\lvert z \rvert = [R_2(R_1+R_3)+R_1R_3-r_m(mR_1+nR_3)] = $$
$$= [r_e(r_v+r_k)+r_v(r_k-r_m)].$$

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Table 5

Basic Ratios

$[M]$	$K_u = \frac{U_2}{E_0}$	$K_i = \frac{I_2}{I_1}$	Z_{in}	Z_{out}
$[z]$	$\frac{Z_{21}Z_n}{H_z}$	$\frac{Z_{21}}{Z_n - Z_{22}}$	$\frac{Z_{11}Z_n - \|z\|}{Z_n - Z_{22}}$	$-\frac{Z_{22}Z_0 + \|z\|}{Z_0 + Z_{11}}$
$[y]$	$\frac{Y_{21}Z_n}{H_y}$	$\frac{Y_{21}}{-\|y\| Z_n + Y_{11}}$	$\frac{Y_{22}Z_n - 1}{\|y\|Z_n - Y_{11}}$	$-\frac{Y_{11}Z_0 + 1}{\|y\| Z_0 + Y_{22}}$
$[d]$	$\frac{d_{22}Z_n}{H_d}$	$\frac{d_{22}}{-d_{21}Z_n + 1}$	$\frac{\|d\| Z_n + d_{12}}{-d_{21}Z_n + 1}$	$\frac{Z_0 + d_{12}}{-d_{21}Z_0 + \|d\|}$
$[f]$	$\frac{f_{11}Z_n}{H_f}$	$\frac{f_{11}}{f_{21}Z_n + \|f\|}$	$\frac{Z_n - f_{12}}{f_{21}Z_n + \|f\|}$	$\frac{\|f\| Z_0 - f_{12}}{f_{21}Z_0 + 1}$
$[a]$	$\frac{Z_n}{H_a}$	$\frac{1}{a_{21}Z_n + a_{22}}$	$\frac{a_{11}Z_n + a_{22}}{a_{21}Z_n + a_{21}}$	$\frac{a_{22}Z_0 + a_{12}}{a_{21}Z_0 + a_{11}}$

$$H_z = -Z_{22}Z_0 + Z_{11}Z_n + Z_0Z_n - |z|; \quad H_f = |f|Z_0 + Z_n + f_{21}Z_0Z_n - f_{12};$$
$$H_y = Y_{11}Z_0 - Y_{22}Z_n - |y| Z_0Z_n + 1;$$
$$H_d = Z_0 + |d| Z_n - d_{21}Z_0Z_n + d_{12};$$
$$H_a = a_{22}Z_0 + a_{11}Z_n + a_{21}Z_0Z_n + a_{12}.$$

Semiconductor Stage in the High-Frequency Field

We will demonstrate the procedure of applying the method to an analysis of the equivalent semiconductor stage at high frequencies. For the semiconductor stage with a common emitter, the equivalent circuit, adapted for a wide range of frequencies, is illustrated in Fig.4 (Bibl.3; 4). This system may easily be represented in the form of Fig.5, since the latter is so designated as to be indistinguishable from the equivalent circuit of a tube stage with a common cathode.

To find the parameters of the layout in Fig.5, all that is necessary is to set up the matrix [y] of this schematic and compare it with the matrix [y] of the schematic in Fig.4. This leads to the following results:

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$$\left.\begin{aligned}
Y_{gk} &= \frac{Y_{ve}}{1 + r_v (Y_{ve} + Y_{vn})} \\
Y_{ag} &= \frac{Y_{vn}}{1 + r_v (Y_{ve} + Y_{en})} \\
Y_{an} &= Y_{ne} + \frac{r_v Y_{vn} (g_m + Y_{ve})}{1 + r_v (Y_{ve} + Y_{vn})} \\
S &= \frac{g_m}{1 + r_v (Y_{ve} + Y_{vn})}
\end{aligned}\right| \qquad (1)$$

As demonstrated above, the system in Fig.5 does not differ from the tube stage layout with a common cathode. Neither is there any difference between the equivalent circuits of semiconductor stages with a common base and the tube stage with a

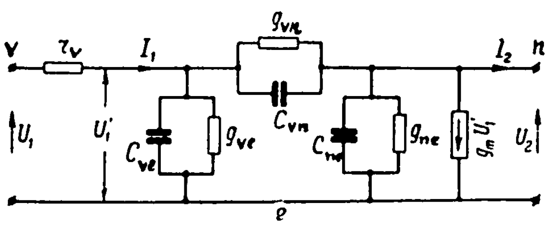

Fig.4

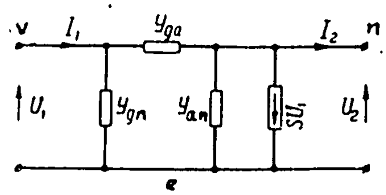

Fig.5

common grid, or between the equivalent circuits of semiconductor stages with a common collector and the tube stage with a common plate.

Therefore, the generalized matrixes of the semiconductor stage at high currents can be obtained from Tables 1 and 3 by means of eq.(1), if one bears in mind the above correspondence between tube and semiconductor stages.

If required, it is simple to convert from the schematic in Fig.5 to the usual equivalent semiconductor stage T-network. The corresponding conversion formulas have the form

$$Z_v = \frac{Y_{an}}{-|y|}; \qquad Z_e = \frac{Y_{ag}}{-|y|}; \qquad Z_n = \frac{Y_{gn} + S}{-|y|}; \qquad Z_m = \frac{S}{-|y|}, \qquad (2)$$

whereupon

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$$y = Y_{ag}(S + Y_{gn} + Y_{an}) + Y_{gn}Y_{an}.$$

In the indicated manner, as an example, we can define the matrix [d] of the semiconductor stage with a common base in the high-frequency range. For this, it is necessary to take the common matrix [d] from Table 3, substitute there the numerical values from the second line in Table 1 (since the system with a common base corresponds to the one with a common grid), and then convert to the parameter of the semiconductor triode according to eq.(1).

The matrix [d] of the common base system obtained in this way has the form

$$[d] = \frac{1}{Y_{gn} + Y_{an} + S}\begin{bmatrix} Y_{an} & 1 \\ y & S + Y_{an} \end{bmatrix}. \quad (3)$$

It is known that the element d_{12} of the matrix [d] represents the input resistance of the stage in the short-circuit of the output, while the element d_{22} is the stage amplification according to the current of this system. Therefore,

$$(Z_{in})_0 = \frac{1}{Y_{gn} + Y_{an} + S}; \quad (4)$$

$$(K_i)_0 = \frac{S + Y_{ak}}{Y_{gn} + Y_{an} + S}. \quad (5)$$

Since the amplification according to the current in the short-circuit hookup with a common base is equal to α, eq.(5) represents the complex value α for the high-frequency range.

For example, for the triode, having the following data (Bibl.3):

$$r_v = 200 \text{ ohm}; \; g_{ve} = 10^{-3} \text{ ohm}; \; g_{..} = 4 \cdot 10^{-2} \text{ megohm}$$

$$g_{vn} = 2 \cdot 10^{-7} \text{ megohm}; \; g \;\; = 10^{-5} \text{ ohm};$$

$$C_{ve} = 3000 \; \mu\mu f; \; C_{vn} = 30 \; \mu\mu f.$$

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The Z parameters for the high-frequency field and the approximated expressions for α at high frequencies, in accordance with eqs.(2) and (5) are as follows:

$$r_V = 750 \text{ ohm}; \; r_e = 11 \text{ ohm};$$

$$r_n = 2{,}26 \cdot 10^6 \text{ ohm}; \; r_m = 2{,}2 \cdot 10^6 \text{ ohm}$$

$$\alpha = \frac{0{,}975}{1 + j\omega \cdot 0{,}73 \cdot 10^{-7}} \cdot$$

In accordance with the latter expression, the limiting frequency according to α equals

$$f_\alpha = 2{,}2 \cdot 10^6 \text{ cycles}$$

APPENDIX

Semiconductor Triode h-Parameters

The foreign literature makes extensive use of the matrix [h] in place of the matrix accepted by us [d]

$$\begin{vmatrix} U_1 \\ I_2 \end{vmatrix} = |h| \begin{vmatrix} I_1 \\ U_2 \end{vmatrix}.$$

In addition, for a positive output current, counterclockwise direction is used. Between the matrix elements [d] and [h], the following relations take place:

$$h_{11} = d_{12}, \; h_{12} = d_{11}; \; h_{21} = -d_{22}; \; h_{22} = -d_{21}.$$

Inasmuch as the h-parameters of the semiconductor triode are easily measured and have a simple physical concept, they are frequently assigned as primary parameters. It is therefore useful to establish the connection between the h- and r-parameters of the semiconductor triode.

Frequently, the h-parameters are gaged for either the common base or the common emitter system, but not for the system with a common collector. In order not to duplicate the computation, it is convenient to use the generalized circuit in Fig.2, substituting here m = 0, i.e., excluding the common collector system.

Expressing the matrix [z] through the elements of the matrix [h] and representing it at the same time as the generalized matrix [z] of the schematic in Fig.2 when m = 0, we obtain the following matrix equation

$$[z] = \frac{-1}{h_{22}} \begin{bmatrix} -|h| & h_{12} \\ h_{21} & 1 \end{bmatrix} = \begin{bmatrix} R_1 + R_3 & -R_3 \\ R_3 + nr_m & -(R_2 + R_3) \end{bmatrix}.$$

Equating the corresponding elements, we get

$$R_1 = \frac{|h| - h_{12}}{h_{22}}; \quad R_2 = \frac{1 - h_{12}}{h_{22}};$$
$$R_3 = \frac{h_{12}}{h_{22}}; \quad nr_m = -\frac{h_{21} + h_{12}}{h_{22}}.$$

The transition from the generalized resistance R to r (parameters of the crystal triode) is accomplished according to the data given in Table 2, with the calculation of the type of stage, in short, given as h-parameters.

Assume, for example, that the h-parameters, calculated for the common base system, have the following values:

$$h_{11} = 40 \text{ ohm}; \; h_{12} = 2{,}5 \cdot 10^{-4};$$
$$h_{21} = -0{,}92; \; h_{22} = 1 \cdot 10^{-6} \text{ ohm}$$

Using the data given in Table 2 for the common base system and the formulas presented above, we get

$$R_1 = r_e = 20 \text{ ohm}$$
$$R_2 = r_n = 10^6 \text{ ohm}$$
$$R_3 = r_v = 250 \text{ ohm}$$
$$nr_m = r_m = 0{,}92 \cdot 10^6 \text{ ohm}$$

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DETERMINATION OF THE BASIC PARAMETERS OF MULTICHANNEL RADIO RELAY SYSTEM APPARATUS

by

S.V.Borodich

The method for determining the optimum values for the basic parameters of multichannel radio relay systems with frequency division multiplex and frequency modulation is discussed. The optimum quantitative correlations are established between the individual noise components in the telephone channel.

Introduction

In designing multichannel radio-relay systems, the designer must primarily determine its basic parameters, based on the fact that the equipment must supply the required quality of the coupling, and that its practical execution has to be connected with minimum technical difficulties. In other words, the designer must determine the optimum values of the basic parameters of the equipment, including: power of the transmitter; antenna amplification factor; receiver noise factor; frequency swing of the transmitter within the channel; maximum amplification of the tandem office; parameters, determining the nonlinearity of the line route (linearity of the modulator, demodulator, and the phase characteristic of the channel). Several of these parameters may be selected on the basis of existing technical possibilities; however, in order to make a proper choice, the designer must know how any change in one parameter will affect the others.

The present article is an attempt to establish the method of determining optimum values of the basic parameters of the equipment. In summary, this article has been included as a supplement to the new recommendation to the MKKR [International

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Consultant Committee for Radio (Bibl.1)], to determine the allowable noise intensity at the end of a hypothetical standard circuit for the radio-relay communications systems. The recommendation was accepted by the Eighth Plenary Meeting of the MKKR in September 1956.

Initial Data for the Solution of the Problem

The basic quality factor of communication in a telephone channel is its noise level. All values higher than the parameter of the equipment immediately influence the quantity of noise in the channel, whereas the remaining characteristics of the telephone channel, such as stability of overall circuit attenuation, frequency and amplitude characteristics, transit time of the signal, proofing against audible crosstalk, channel stability, and so on, do not depend on these parameters. The noise volume in the channel depends also on the length of line, its construction, and the conditions of radio-wave propagation. Consequently, the allowable noise volume serves as a guide for the designers at the end of a hypothetical standard circuit, which possesses a determined length and structure.

New recommendations by the MKKR, adopted at the Eighth Plenary Meeting, define two hypothetical standard circuits for radio-relay systems and give the permissible noise level at the ends of these circuits (Bibl.2).

The first hypothetical standard circuit is designated for radio-relay systems with a capacity of 12 - 60 telephone channels. It has a length of 2500 km and contains three pairs of individual, six pairs of group, and six pairs of supergroup converters, as well as six pairs of radio modulators and demodulators, respectively. This circuit is divided into six sections of equal length, contained between the stations with signal demodulation.

The second hypothetical standard circuit, designated for radio-relay systems, with a capacity of more than 60 channels, has the same length and number of individual and group converters, but contains nine pairs of supergroup converters and

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nine pairs of radio modulators and demodulators. It is divided into nine sections of equal length, contained between the stations with signal demodulation and modulation.

The common psophometric noise power at the relative zero level in any telephone channel, at the end of both the first and the second hypothetical standard circuits, is determined by the following time values (Bibl.1):

a) Average power per hour of greatest usage; in the absence of fading, this may not exceed 5000 μμw;

b) Average power at any hour, not to exceed 7500 μμw.

These values make no allowance for the noise of the multiplexing equipment whose power, in conformity with the MKKF recommendation, does not exceed 2500 μμw.

It should be mentioned that this recommendation, strictly speaking, is not adequate. At the time of strong signal fading in the course of short time intervals, the noise volume may grow so much that the connection is broken despite the fact that the average noise volume per hour does not rise above the permissible values. The recommendation, therefore, indicates that it is essential to determine the manner in which the allowable noise level should be assigned in the course of short time intervals; however, at present, it is impossible to give a precise answer to this question.

The above-mentioned recommendations by the MKKR which define the hypothetical standard circuit and the allowable noise volume, may be used as initial data for determining the equipment parameters.

Optimum Ratios of Noise Components in Telephone Channels

It was shown in another paper (Bibl.3), that the psophometric noise magnitude in the telephone channel of radio-relay systems (without considering the multiplexing equipment noise), at the point of zero relative level, equals

$$P_{-} = P_{tn} + P_{nn2} + P_{nn3} = A\frac{1}{x} + Bx + Cx^2 \ \mu\mu w, \qquad (1)$$

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i.e., represents the total intensity of thermal noise, $R_{tn} = A \frac{1}{x}$ the noise power of nonlinear crossings of the second order $R_{tn_2} = Bx$, and the noise power of nonlinear crossings of the third order $R_{nn_3} = Cx^2$. The products of nonlinearity larger than the highest order are not taken into account in view of their relative smallness. The coefficients A, B, C are determined by the basic parameters of the equipment, and also by the length and structure of the system. The dimensionless value x is proportional to the square of the frequency swing to the channel

$$x = \left(\frac{\Delta f_k}{\Delta f_1}\right)^2, \qquad (2)$$

where Δf_k is the effective value of the frequency swing to the channel, when sending in the channel at the zero relative level, a sine current of 1 mw and 800 cycles;

Δf_1 is some unit frequency swing, whose selection is somewhat arbitrary (i.e., 1 mc).

The recommendation by the MKKR specifies the allowable mean value of the noise level in the course of any hour. Besides, in conformity with this recommendation, sections of the hypothetical standard circuit have a homogeneous form, and the entire allowable noise power for the entire circuit can be divided equally among the parts. In this way, it is possible to consider only one section of the hypothetical standard circuit, contained between two stations with signal modulation and demodulation which, while not detracting from the communal feature, does somewhat simplify the problem.

Bearing this in mind, we can apply eq.(1) to determining the average, for any hour (including the hour of greatest charge), of the noise power in the telephone channel of one section of a hypothetical standard circuit

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$$P_{n1} = A_1 \frac{1}{x} + B_1 x + C_1 x^2. \qquad (3)$$

The coefficients A_1, B_1 and C_1 for one section of the circuit are determined

(Bibl.3) by the following expressions:

$$A_1 = A_0 \bar{S}_{1\,hr}, \tag{4}$$

$$A_0 = \frac{4n\Delta F_k k_n^2}{P_{tr}} \left(\frac{F_k}{\Delta f_1}\right)^2 \frac{16\pi^2 R_0^2 e^{2\alpha l_0}}{\lambda^2 G_A^2}, \tag{5}$$

where n is the coefficient of receiver noise;

P_{tr} is the power of the transmitter (in μμw);

ΔF_k is the bandwidth of the channel (in cycles);

k_n is the psophometric coefficient;

F_k is the mean frequency of the channel in the linear spectrum;

λ is the wavelength;

G_A is the antenna amplification factor (relative to the unipole);

R_0 is the length of one stretch of the line;

α is the transmission line fading, in nepers per unit length;

l_0 is the entire length of the feeders in one section.

Since the sections of the hypothetical standard circuit are divided into m_1 unit parts, then

$$\bar{S}_{1\,hr} = \sum_{i=1}^{m_1} \left(\overline{\frac{1}{V_i^2}}\right)_{hr}, \tag{6}$$

where V_i is the field attenuation factor of free space in the i – M sector.

The value $\left(\frac{1}{V_1^2}\right)_{hr}$ is the mean for one hour of accidental quantity $\frac{1}{V_1^2}$, inasmuch as eq.(3) determines the average for one hour of noise volume.

Further,

$$B_1 = \frac{\Delta F_k k_n^2 \cdot 10^9}{\Delta F} e^{4b_{mean}} y_2(\sigma_k) \left[4e^{-2b_{s1}} + (2\pi F_k)^2 \Delta f_1^2 \frac{1}{2} \gamma_1^2\right], \tag{7}$$

$$C_1 = \frac{\Delta F_k k_n^2\, 10^9}{\Delta F} e^{6b_{mean}} y_3(\sigma_k) \left[24e^{-2b_{s1}} + (2\pi F_k)^2 \Delta f_1^4 \frac{2}{3} \gamma_2^2\right], \tag{8}$$

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where ΔF is the width of the line spectrum;

b_{mean} is the equality (in nepers) between the mean power level of all channels and the measured level of one channel*;

$y_2(\sigma_k)$, $y_3(\sigma_k)$ are the functional values of the distribution of spectral density of the nonlinear products of the second and third order at a frequency F_k of the given channel;

b_{21}, b_{31} are the nonlinear fading (in nepers) at double- and triple-frequency harmonics for stations with modulation and demodulation signals (the circuit consisting of modulator, demodulator, and group amplifier), measured for unit frequency variation Δf_1;

γ_1, γ_2 are the resolution factors (for first and second degrees of separation) of the characteristic of the group signal time in the channel of one section of the hypothetical standard circuit, engaging the m_1 station.

Equations (7) and (8) disregard the nonlinear products formed in the antenna feeders as a consequence of the energy reflected from the mismatched load. If the feeders are comparatively short, these products will mostly be nonlinear products of the second order and may be considered as adding the following value to the right-hand side of eq.(7):

$$B_{f1} = \frac{\Delta F_k k_n^2 \cdot 10^9}{\Delta F} e^{4b_{mean}} 2(2\pi F_k)^2 (2\pi \Delta f_1)^2 \sum_{i=1}^{2m_1} \tau_{fi}^4 r_{1i}^2 r_{2i}^2, \qquad (9)$$

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*The Third Research Committee of the MKKF recommends that this value be considered as equal to $b_{mean} = -1.72 + \frac{1}{2} \ln N$ when there is a sufficient number of channels N. The value -1.72 neper represents the mean level (per hour of greatest charge) of the speaking currents power in the channel at the relative zero level, determined on the basis of measurements in several countries. Unfortunately, no suitable value has been determined for our domestic communications systems.

where $2m_1$ is total number of feeders;

τ_{f1} is the group wave transit time in the i - M feeder;

r_{1i}, r_{2i} are the reflectance at the ends of the i^{th} feeder.

Equation (9) is valid, provided that

$$4\pi\Delta f_k e^{b_{mean}}\tau_{fi} \leq 0,5. \qquad (10)$$

Otherwise the calculation of the nonlinear products, originating in the feeders, in eq.(3) would become extraordinarily complicated.

It is apparent from eqs.(2), (5), (7), and (8) that the values x, A_o, B_1, C_1 are determined by the basic parameters of the equipment, whereas the value $A_1 = A = A_o\bar{S}_{1\ hr}$ depends also on the radio-wave propagation conditions. Therefore, the problem which we presented at the beginning of this article may be defined as follows: It is required to determine the values A_o, B_1, C_1, and x from the condition that the noise volume in the channel at the end of the hypothetical standard circuit is equal to the allowed value P_{nd} = 7500 μμw.

For one section of the hypothetical standard circuit, this condition corresponds to the equation

$$\frac{P_{nd}}{\nu} = P_{nd_1} = A_0\bar{S}_{1\ hr}\frac{1}{x} + B_1 x + C_1 x^2, \qquad (11)$$

where is the number of sections in the hypothetical circuit (ν = 6 at N = 12 ÷ 60 and ν = 9 at N > 60).

It is obvious that eq.(11) alone is not adequate for determining the four unknowns A_o, B_1, C_1, and x. We will, therefore, present a second condition, guaranteeing a minimum of technical difficulty in the practical construction of the equipment. For this, we assume that the value x, determined by the frequency variation, is the optimum value and that the noise volume will thus be minimal. Differentiating eq.(11) by x and adjusting the result to zero, we obtain the second equation

$$0 = -A_0\bar{S}_{1\ hr}\frac{1}{x^2} + B_1 + 2C_1 x. \qquad (12)$$

The two equations (11) and (12) are still insufficient for solving the problem; however they will yield

$$\left.\begin{aligned} B_1 x &= 2P_{nd_1} - 3A_0\bar{S}_{1\,hr}\frac{1}{x} \\ C_1 x^2 &= 2A_0\bar{S}_{1\,hr}\frac{1}{x} - P_{nd_1} \end{aligned}\right\}, \tag{13}$$

which implies

$$\left.\begin{aligned} P_{nn_2} &= 2P_{nd_1} - 3P_{tn} \\ P_{nn_3} &= 2P_{tn} - P_{nd_1} \end{aligned}\right\}. \tag{13a}$$

Apparently, in the real system, the power of nonlinear noise of the second order, as well as the power of nonlinear noise of the third order, cannot be equal to zero and still be less than zero. Bearing this in mind, we obtain from eq.(13a) the third condition

$$\frac{1}{2}P_{nd_1} < P_{tn} < \frac{2}{3}P_{nd_1}. \tag{14}$$

We further assume that the thermal noise volume equals

$$P_{tn} = A_0\bar{S}_{1\,hr}\frac{1}{x} = \xi P_{nd_1},$$

where the value ξ, in conformity with eq.(14), lies within the limits of

$$\frac{1}{2} < \xi < \frac{2}{3}. \tag{15}$$

Then from eqs.(13) and (13a) we find the ratios between the components of the noise in the telephone channel

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$$P_{tn} = A_0\bar{S}_{1\,hr}\frac{1}{x} = \xi P_{nd_1}, \tag{16}$$

$$P_{nn_2} = B_1 x = 3P_{nd_1}\left(\frac{2}{3} - \xi\right), \tag{17}$$

$$P_{nn_3} = C_1 x^2 = 2P_{nd_1}\left(\xi - \frac{1}{2}\right); \tag{18}$$

where ε satisfies the condition (15).

The ratios (16), (17), and (18) are optimum since they are derived from the condition that the minimum possible noise volume in the channel of the radio-relay system is determinable by the parameters A_o, $\bar{S}_{1\ hr}$, B_1, C_1 and x, equal to the allowable value P_{nd_1}.

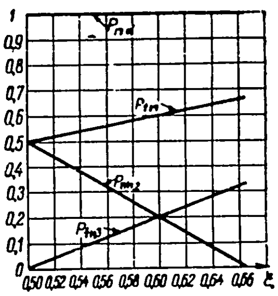

Fig.1

To illustrate the ratios obtained, the noise components are plotted in Fig.1 as a function of the value ε, in which the allowable noise level is taken as unity.

Determination of the Basic Parameters of the Equipment

We note that, of the four values A_o, B_1, C_1, x, determining the basic parameters of the equipment, only one value, namely x, can be considered known, since the frequency deviation of the channel has been fixed by the specifications established at the Eighth Plenary Meeting of the MKKR in September 1956. To insure international contact of the radio-relay lines, the MKKR recommends the following effective frequency variation values of the channel Δf_k, for systems with 24 to 600 channels (Bibl.4).

Number of Telephone Channels in the System	Effective Value of Frequency Variation in the channel Δf_k (kc)
24	35
60	50, 100, 200
120	50, 100, 200
240	200
600	200

Therefore the three equations (16), (17), and (18) obtained above are entirely adequate for determining the three unknowns A_0, B_1, and C_1.

From this, the following equations are obtained:

$$A_0 = P_{nd_1} \frac{\xi x}{\overline{S}_{1\ hr}}, \tag{19}$$

$$B_1 = P_{nd_1} \frac{(2 - 3\xi)}{x}, \tag{20}$$

$$C_1 = P_{nd_1} \frac{(2\xi - 1)}{x^2}. \tag{21}$$

In these expressions, the value ξ is not accurately determined; it is known only that it must satisfy the inequation (15). In selecting the value ξ, the designer has an opportunity of changing the numerical ratio of the noise components, i.e., to increase one of the parameters (B_1 or C_1) at the expense of the other, if it is expected that nonlinearity of the second or third order will be predominant.

As can be seen from eq.(19), the parameter A_0 also depends on the nonlinearity of the value $\overline{S}_{1\ hr}$, expressing the conditions of the crossing signal for all m_1 parts of one section of a hypothetical standard circuit. It is known that the attenuation factor V_1, in every part of the system, is an accidental value. Thus, the sum

$$S_1 = \sum_{i=1}^{m_1} \frac{1}{V_i^2},$$

is also a determinant of the volume of thermal noise and likewise is an accidental value. In order to find the probability distribution for S_1, it is necessary to know not only the laws of distribution probability for all V_i, but also the ratio of these accidental values, since they are not independent. This is obviously a very complicated task.

However, the allowable noise volume is given as an average per hour. Consequently, as already mentioned, we must consider the mean value of the sum $\overline{S}_{1\ hr}$ per hour, which is also an accidental value.

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Treating the problem in this manner greatly simplifies the task, since the mean value of the sum is equal only to the sum of the mean values of the individual components

$$\overline{S}_{1\,hr} = \sum_{i=1}^{m_1} \overline{\left(\frac{1}{V_i^2}\right)}_{hr}$$

and it is not required to know the correlations between V_i in the various parts. Nevertheless, to determine the probability distribution of the value $\overline{S}_{1\,hr}$ the laws of distribution for the values $\overline{\left(\frac{1}{V_i^2}\right)}_{hr}$ in all sections must be known. It should be mentioned that the statistical probability distribution curves for the values of V_i in the various sections, published in numerous papers, cannot be used for determining $\overline{\left(\frac{1}{V_i^2}\right)}_{hr}$, since they do not present the statistics on an hour basis and for a long period of time. Thus, a corresponding statistical revision of existing material and a compilation of new experimental data is needed on the probability distribution of the value $\overline{\left(\frac{1}{V_i^2}\right)}_{hr}$ or, directly, the values of mean thermal noise volume in the channels of various sections of the system.

The problem may be solved by approximation, if the probability distribution of the value $\overline{\left(\frac{1}{V_i^2}\right)}_{hr}$ for some average or even for the less efficient part of the system is known. In this case, the wanted value of the sum

$$\overline{S}_{1\,hr} \approx m_1 \overline{\left(\frac{1}{V_{mean}^2}\right)}_{hr}, \qquad (22)$$

where the value $\overline{\left(\frac{1}{V_{mean}^2}\right)}_{hr}$, is so selected that the accidental value $\overline{\left(\frac{1}{V_i^2}\right)}_{hr}$ in the given part may exceed the one selected in the course of a fixed but small percentage of the time.

The recommendation by the MKKR specifies the allowable mean noise volume in the

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channel in the course of any hour. The optimum correlations listed above eqs.(16), (17), and (18) between the noise components and the basic parameters of the equipment (19), (20), (21), whose calculation is based on these correlations, refer to the hour of greatest load, when the nonlinear noise level is at its highest. Therefore, the value $\bar{S}_{1\,hr}$, essential for the computation, must be specified for the daylight hours, when, as experience shows, the signal fading is usually not great. At night and in the early morning when the fading is usually greater, the line load is small, nonlinear noise is almost absent and the rise in thermal noise above the optimum value is completely permissible.

The supplements to the recommendation by the MKKR contains examples for typical probability distribution curves of thermal noise in the sections of the system, and also examples of the reckoning from which it follows that, even under adverse conditions, the value is $\left(\overline{\frac{1}{V^2}}_{\text{mean}}\right)_{hr} = 2 - 3\ (3 - 5)$ db, on the average, in each part of a line of 2500 km.

These considerations are also contained in the same recommendation so that a comparison of two suggested values will give the permissible noise volume: for an hour of greatest load in the absence of fading, and for any given hour.

Actually, if fading is absent, then $\bar{S}_{1\,hr} = m_1$ and eq.(11), determining the noise level, may be written as

$$\frac{P_{nd_0}}{\nu} = P_{nd_{10}} = A_0 m_1 \frac{1}{x} + B_1 x + C_1 x^2,$$

where $P_{nd_0} = 5000$ μμw, in accordance with the recommendation.

If eqs.(16), (17), and (18) are substituted here, bearing eq.(22) in mind, then it is possible to obtain quite readily

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$$\frac{\bar{S}_{1\,hr}}{m_1} = \left(\overline{\frac{1}{V^2_{\text{mean}}}}\right)_{hr} = \frac{\xi}{\frac{P_{nd_1}}{P_{nd}} - 1 + \xi}.$$

With a change of ε in the interval (15), the value $\left(\overline{\frac{1}{V2}}\right)_{\text{mean hr}}$ will vary from 3 to 2, if $\frac{P_{ndo}}{P_{nd}} = \frac{5000}{7500}$.

However, this does not infer that the value $\overline{S}_{1\,hr}$ should be chosen in all cases to be equal to $\overline{S}_{1\,hr} = (2 - 3)m_1$. It is obvious that the value $\overline{S}_{1\,hr}$ depends on the usable wave range and has to be determined by the principles of statistical measurement by various means.

Computing the values A_o, B_1, and C_1 according to eqs.(19), (20), and (21) it is easy to determine the basic parameters of the equipment from eqs.(5), (7), (8), and (9). Then the designer still has some freedom in choosing the values of the remaining parameters.

Thus, eq.(5) will yield the value which can be called the equivalent power

$$P_e = \frac{P_{tr}\, G_A^2}{n\, e^{2\alpha l_0}} = \frac{4\Delta F_k k_n^2 16\pi^2 R_0^2}{\lambda^2 A_0^2}\left(\frac{F_k}{\Delta f_1}\right)^2. \qquad (23)$$

The designer has the choice of selecting the values of the individual parameters n, P_{tr}, G_A, resulting from the existing technical possibilities of their realization under the condition that the value of P_e, which is determined by eq.(23), remains unchanged.

Exactly in the same way, the values b_{21} and γ_1, b_{31} and γ_2, determining the allowable nonlinearity of the route, cannot be found in a simple operation from eqs.(7) and (8), using the computed values B_1 and C_1. However, by fixing the values of b_{21} and b_{31} it is possible to find γ_1 and γ_2, respectively. In the process of drafting the equipment further, in the basic computation of the characteristics of the individual junctures, a change may be required in any of the selected values; however there must also be a corresponding change in its dependent values such that the values B_1 and C_1 remain constant.

To simplify the calculations, it is recommended to plot the functions of γ_1 on b_{21} and γ_2 on b_{31}, for the found values of B_1 and C_1.

As examples, Figs.2, 3, and 4 show the graphs, calculated according to the formulas given, by means of which the basic parameters of the equipment can be determined.

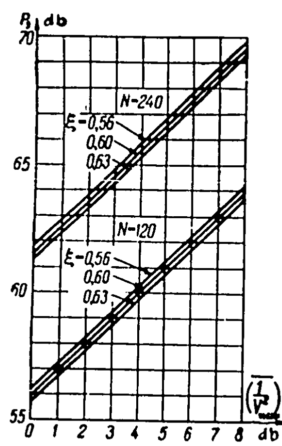

Fig.2

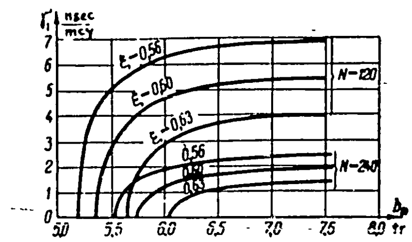

Fig.3

These graphs are calculated for two radio-relay systems with 120 and 240 telephone channels, having the following data:

$\Delta F_k = 3.1$ kc; $k_n = 0.75$; $\Delta f_k = 200$ kc; $\Delta f_1 =$ $= 707$ kc (amplitude equals 1 mc); $v = 9$; $m_1 = 6$; $R_o =$ $= 43.6$ km; $\lambda = 7.5$ cm; $b_{mean} = 1.72 + \frac{1}{2} \ln N$;

1) The system with 120 channels: $N = 120$; $F_k =$ $= 552$ kc; $\Delta F = 492$ kc; $y_2(\sigma_k) = y_3(\sigma_k) = 0.45$; $b_{mean} =$ $= 0.67$ nepers;

2) The system with 240 channels: $N = 240$; $F_k =$ $= 1052$ kc; $\Delta F = 992$ kc; $y_2(\sigma_k) = y_3(\sigma_k) = 0.5$; $b_{mean} =$ $= 1.02$ neper.

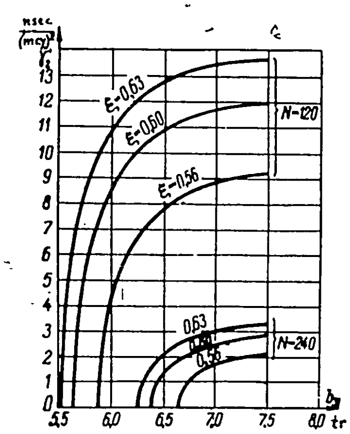

Fig.4

In Fig.2, the functions of the equivalent power P_e are shown, reflected in decibels proportional to 1 watt, dependent on the mean value of signal fading in each section $\left(\frac{1}{\overline{V2}_{mean}}\right)_{hr}$. In Figs.3 and 4, the functions of γ_1 at b_{21} and γ_2 at b_{31} are shown, from which it becomes quite clear that an increase in attenuation of the nonlinearities b_{21} and b_{31} above the determined values is of no meaning, since γ_1 and γ_2 remain practically constant.

If the nonlinear products, arising in the feeders, increase further, then the value γ_1', determinable from the graph in Fig.3, will equal

$$\gamma_1' = \sqrt{\gamma_1^2 + \frac{32\pi^2 m_1 \tau_f^4 r_1^2 r_2^2}{y_2(\sigma_k)}}.$$

This expression, derived from eqs.(7) and (9), is correct for calculating the inequation (10).

The values γ_1 and γ_2 are the coefficients of the group time characteristics of the signal transit over the track of the entire section of the system, consisting of m_1 stations. To find the corresponding coefficients γ_{1i} and γ_{2i} for one station, it is necessary to know the combination law of these coefficients. This problem is considered in some detail in a previous paper (Bibl.3).

Maximum Amplification of the Tandem Office

The maximum amplification of the tandem office is determined primarily by the need for stability of the connection. If, as the result of deep fading, the signal level at the receiver input of any station drops below the threshold of the amplitude limiter, a drop in the power of the signal at the station output will result, accompanied by a sharp rise in noise, which may result in a disruption of the connection.

If we assume that an interruption of this type in connection in a line consisting of m parts represents δ% of the time, then a nonsimultaneous deep fading in the various parts, will lead to disruptions $\frac{\delta}{m}$% of the time.

Let us assume that in $\frac{\delta}{m}$% of the time the fading factor V in the section of the system must be smaller than a value V_{min}. Then the total attenuation in the section of the connection will be

$$\mu < \frac{16\pi^2 R_0^2 e^{2\alpha l_0}}{\lambda^2 G_A^2 V_{min}^2}$$

during $(100 - \frac{\delta}{m})\%$ of the time.

Obviously then, to insure a stable connection over $(100 - \frac{\delta}{m})\%$ of the time, the maximum amplification of the intermediary station should be equal to that fading

$$\mu_{max} = \frac{16\pi^2 R_0^2 e^{2\alpha l_0}}{\lambda^2 G_A^2 V_{min}^2}. \qquad (24)$$

The value V_{min} is derived from the statistical probability distribution curve of fading in the various sections of the system.

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AN INVESTIGATION OF THE SELF-OSCILLATING SYSTEM IN AN OSCILLATOR WITH A SEMICONDUCTOR JUNCTION-TYPE TRIODE

by

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Phase relations are examined in the semiconductor junction-type triode, and a conclusion is drawn as to phase balance in the self-oscillator. It is shown that phase balance failure may prevent the generation of self-oscillation at higher frequencies. Systems are proposed with phase correction, permitting a substantial boost in the critical frequency. An analysis is given of oscillation characteristics and recommendations are advanced, relating to the selection of the optimum system of self-oscillation.

An Investigation of Phase Relations

In self-oscillators, not only the energy indexes have a decisive significance but also the phase relationships. In the investigation of oscillators with independent excitation and semiconductor junction-type triodes, the phase relations were not considered (Bibl.1).

In the steadily working system of the self-oscillator both phase and amplitude balance are attained. It is known from the theory of self-oscillation that the condition for phase balance is

$$\varphi_s + \varphi_k + \varphi_a = 2\pi n, \quad (1)$$

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where φ_s is the phase shift in the triode (the phase angle of average transconductance);

φ_k is the phase shift in the feedback circuit;

φ_a is the phase shift in the circuit (in the load);

$n = 1, 2, 3 \ldots$, where n is a natural number series.

The phase shift in the semiconductor triode (or tetrode) differs basically from the phase shift in an electron tube. In the semiconductor triode (in the following, the tetrode can be substituted everywhere too) the phase shift is determined by the effect of the current flowing through the base and the junctions between emitter and base and base and collector.

If the investigation is restricted to no-voltage and critical operation, the low permeability of semiconductor triodes will make it permissible to neglect the collector voltage response, in first approximation.

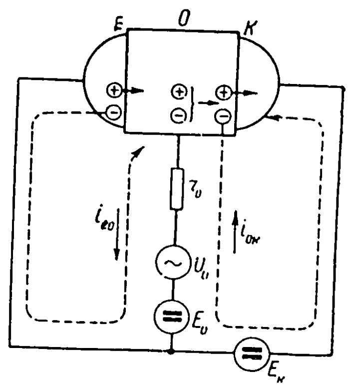

Fig.1

Under these conditions, the equivalent triode system must be derived from an analysis of the currents in the system shown in Fig.1.

Under the effect of the excitation voltage u_o in the circuit, the current begins to flow. A voltage drop occurs in the pedestal resistance, as well as in the electric wires (this is denoted as the resistance r_o). The remainder of the voltage will operate at the emitter-base junction.

Assume that, under the voltage effect at the emitter-base junction

$$u_{eo} = U_{eo} \sin \omega t$$

the following current begins to flow in this circuit:

$$i_{eo} = I_{eo} \sin \omega t.$$

Then, the following current begins to flow in the base-collector circuit;

$$i_{ok} = I_{ok} \sin (\omega t - \varphi_{dr}).$$

The phase angle φ_{dr} is determined by the drift time of the minority carriers (from the emitter-base junction to the base-collector junction) and by the frequency

$$\varphi_{dr} = \omega t_{dr}.$$

The current in the base circuit, generated by the excitation source, will be equal to the difference in currents at the junction (Fig.1)

$$i_0 = i_{e0} - i_{0k} = I_{e0} \sin \omega t - I_{0k} \sin(\omega t - \varphi_{dr}). \qquad (2)$$

In the general case,

$$I_{e0} \neq I_{0k}.$$

However, at a known error it can be permitted that $I_{eo} = I_{ok} = I_k$ if the recombination of the minority carrier with the base is neglected; besides, it is assumed that the slope of the curve of both currents is the same.

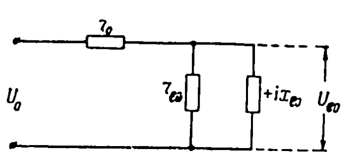

Fig.2

Thus, if it is allowed that $I_{eo} = I_{ok} = I_k$, then eq.(2) will yield

$$i_0 = 2I_k \sin\frac{\varphi_{dr}}{2} \cos\left(\omega t - \frac{\varphi_{dr}}{2}\right), \qquad (3)$$

or

$$i_0 = I_k \sin\varphi_{dr} \cos\omega t + 2I_k \sin^2\frac{\varphi_{dr}}{2} \sin\omega t. \qquad (4)$$

The equivalent system, shown in Fig.2, corresponds to the existence of these two currents.

The equivalent resistances are determined by the conditions

$$\left.\begin{aligned} 2I_k \sin\frac{\varphi_{dr}}{2} &= \frac{U_{e0}}{r_0} \\ \text{and} \quad I_k \sin\varphi_{dr} &= -\frac{U_{e0}}{x_{e0}} \end{aligned}\right\}. \qquad (5)$$

From the conditions of the equivalent circuit of Fig.2 it follows that

$$\left.\begin{aligned} r_{eo} &= \frac{U_{eo}}{2I_k \sin^2 \frac{\varphi_{dr}}{2}} \\ -x_{eo} &= \frac{U_{eo}}{I_k \sin \varphi_{dr}} \end{aligned}\right\} \qquad (6)$$

The ratio $\frac{I_k}{U_{eo}}$ represents the mean transconductance for the variable component of the collector current $I_k \approx S_{mean} U_{eo}$. At the very lowest frequencies, the voltage at the junction U_{eo} is practically equal to the voltage of the excitation source u_o. Therefore, the value of S_{mean} should be determined at the lowest frequencies – as $S_{mean} = \frac{I_k}{u_o}$. In Class A operation, the value S_{mean} is equal to the steepness of the statistical characteristic $S = \frac{dI_k}{dE_o}$ when $E_k \doteq$ const.

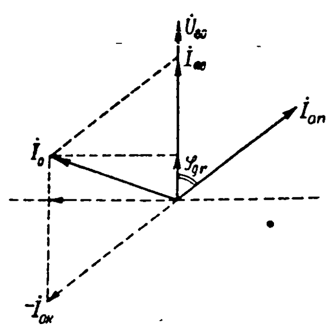

Fig.3

On computation of the stated values, we have

and

$$\left.\begin{aligned} r_{eo} &= \frac{1}{2S_{mean} \sin^2 \frac{\varphi_{dr}}{2}} \\ -x_{eo} &= \frac{1}{S_{mean} \sin \varphi_{dr}} \end{aligned}\right\} \qquad (7)$$

At $0 \leq \varphi_{dr} \leq \pi$, the reactive component will be $x_{eo} < 0$; under the conditions assumed by us this implies that this has the character of capacitance.

This case is well illustrated by the discussed vector diagram (Fig.3). The lagging vector of the current $\dot{I}_{ok}$ corresponds to the vector leading by the phase current $\dot{I}_o$.

It is easy to show that the angle of drift $\pi \leq \varphi_{dr} \leq 2\pi$ corresponds to the reactive resistance of an inductive character ($x_{eo} > 0$). The resistance is $r_{eo} > 0$ at $0 \leq \varphi_{dr} < 2\pi$. The general character of change in the functions r_{eo}, $x_{eo} = \varphi(\varphi_{dr})$ is shown in Fig.4.

Let us determine the equivalent input capacitance C_{eo}. Since, in this case,

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$i x_{eo} = - i \frac{1}{\omega C_{eo}}$, then from eq.(6) we obtain

$$C_{eo} = \frac{I_k \sin \varphi_{dr}}{U_{eo}\omega}. \qquad (8)$$

Equation (8) may be presented as

$$C_{eo} = \frac{I_k t_{dr}}{U_{eo}} \frac{\sin \varphi_{dr}}{\varphi_{dr}} = \frac{Q}{U_{eo}} \frac{\sin \varphi_{dr}}{\varphi_{dr}}; \qquad (9)$$

where $Q = I_k t_{dr}$ is the charge of the minority carriers of the current in the base of the triode.

The capacitive nature of the triode input resistance is well illustrated by eq.(9). The co-factor $\frac{\sin \varphi_{dr}}{\varphi_{dr}}$ testifies to the fact that the value of the equivalent capacitance depends on the angle of drift (Fig.5). The equivalent inductance may be found from eq.(6) under the condition that $i x_{eo} = i\omega L_{eo}$ or

$$L_{eo} = - \frac{U_{eo}}{\omega I_k \sin \varphi_{dr}}. \qquad (10)$$

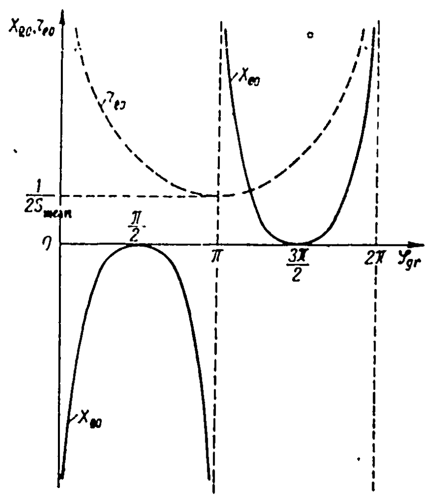

Fig.4

It should not be forgotten that, when $\pi \le \mu_{dr} \le 2\pi$, the co-factor of the denominator is $\sin \varphi_{dr} < 0$ and, consequently, in eq.(10) $L_{eo} > 0$ is obtained.

Turning again to the system in Fig.2, we find the correlation between the voltage at the junction U_{eo} and the voltage of the excitation source U_o. The voltage at the junction equals

$$\dot{U}_{eo} = \frac{\dot{U}_0}{\dot{Z}} \frac{i x_{eo} r_{eo}}{r_{eo} + i x_{eo}} = \dot{U}_0 K e^{i \varphi_{eo}},$$

where

$$\dot{Z} = r_0 + \frac{i x_{eo} r_{eo}}{r_{eo} + i x_{eo}}.$$

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After several transformations, the following equation is obtained:

$$K = \frac{1}{\sqrt{\beta^2 + \sigma^2}} \tag{11}$$

and

$$\varphi_{eo} = \operatorname{arc\,tg} \frac{\sigma}{\beta}, \tag{12}$$

where

$$\sigma = \frac{r_o}{x_{eo}}, \tag{13}$$

$$\beta = \frac{r_o + r_{eo}}{r_{eo}}. \tag{14}$$

Equations (11) and (12), at drift angles of $0 \leq \varphi_{dr} \leq \pi$ and irrespective of the active resistance r_{eo} ($r_{eo} \geq r_o$), yield the previously known fuctions (Bibl.1):

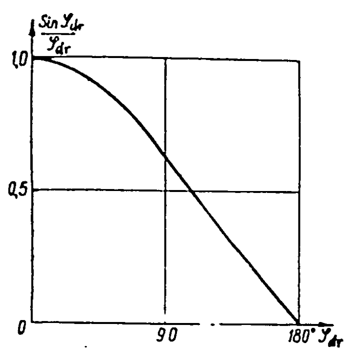

Fig.5

$$K = \frac{1}{\sqrt{1 + (\omega C_{eo} r_o)^2}} \tag{15}$$

and

$$\varphi_{eo} = -\operatorname{arc\,tg} \omega C_{eo} r_o. \tag{16}$$

It is important to establish how K and φ_{eo} change on any variation in φ_{dr} within the limits of 0 to 2π. Using eqs.(7), (13), and (14) we get

$$\sigma = \frac{r_o}{x_{eo}} = -r_o S_{mean} \sin \varphi_{dr}, \tag{17}$$

$$\beta = \frac{r_o + r_{eo}}{r_{eo}} = 1 + 2 S_{mean} r_o \sin^2 \frac{\varphi_{dr}}{2}. \tag{18}$$

The values of interest to us are listed in Table 1.

The greatest interest is represented by the deduction that the angle φ_{eo} changes within the limits of $\pm \frac{\pi}{4}$, since, as indicated in Table 1, $-1 \leq \left(\frac{\sigma}{\beta}\right)_{max} \leq +1$

[see eq.(12)].

Table 1

φ_{dr}	β	σ	$\frac{\sigma}{\beta}$	φ_{eo}	K
0	1	0	0	0	1
$\frac{\pi}{2}$	$1+r_0S_{mean}$	$-r_0S_{mean}$	$-\frac{r_0S_{mean}}{1+r_0S_{mean}}$	$-\operatorname{arctg}\frac{r_0S_{mean}}{1+r_0S_{mean}}$	$\frac{1}{\sqrt{(1+r_0S_{mean})^2+(r_0S_{mean})^2}}$
π	$1+2r_0S_{mean}$	0	0	0	$\frac{1}{1+2r_0S_{mean}}$
$\frac{3\pi}{2}$	$1+r_0S_{mean}$	r_0S_{mean}	$\frac{r_0S_{mean}}{1+r_0S_{mean}}$	$\operatorname{arctg}\frac{r_0S_{mean}}{1+r_0S_{mean}}$	$\frac{1}{\sqrt{(1+r_0S_{mean})^2+(r_0S_{mean})^2}}$
2π	1	0	0	0	1

Typical slopes of the curves $\varphi_{eo} = \varphi(\varphi_{dr})$ and $K = \varphi(\varphi_{dr})$ are illustrated in Fig.6.

At the beginning of the investigation, it was mentioned that the current in the base-collector circuit lags in phase behind the current in the emitter-base circuit by an angle of $\varphi_{dr} = \omega t_{dr}$. Consequently, the overall phase shift between the voltage of the excitation source $\dot{U}_0$ and the current in the collector circuit I_k equals

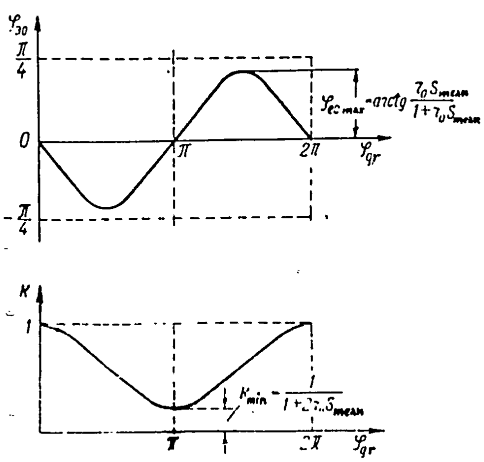

Fig.6

$$\varphi_S = \varphi_{eo} + \varphi_{dr}. \qquad (19)$$

In principle, the angle of drift can be found analytically. In the given case, we discovered a method for determining the angle φ_{dr} by experimental means. To calculate φ_{dr}, the schematic depicted in Fig.7 can be used. In this diagram, under the condition that $R_0 \gg Z_{eo}$, the current of the emitter will be in phase with the source voltage; the alternating voltage in the col-

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lector will be exactly in antiphase with the current I_k.

If, the horizontal plates of the oscillotron, are fed with the voltage U_o and the vertical plates with the voltage $U_k = I_k R_n$, then the phase shift φ_{dr} may be determined by the slope of the axis of an ellipse.

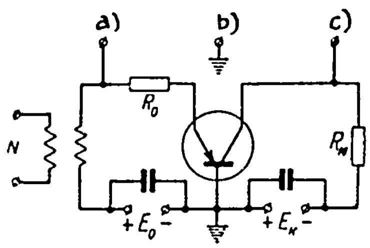

Fig.7

a) Horizontal; b) To the oscillograph; c) Vertical

The results of the measurement of $\varphi_{dr} = \varphi(f)$ for the triode PZB are presented in Fig.8.

The changes are realized in the linear amplification system with a collector current $I_{ko} \approx 30$ ma. The resistance R_o (Fig.7) will have a value of 1 k-ohm and the resistance R_n = = 300 ohms. The driving voltages and the loads from the terminals were fed directly to the plates of the oscillotrons (oscillograph EO-4). The specific arrangement of

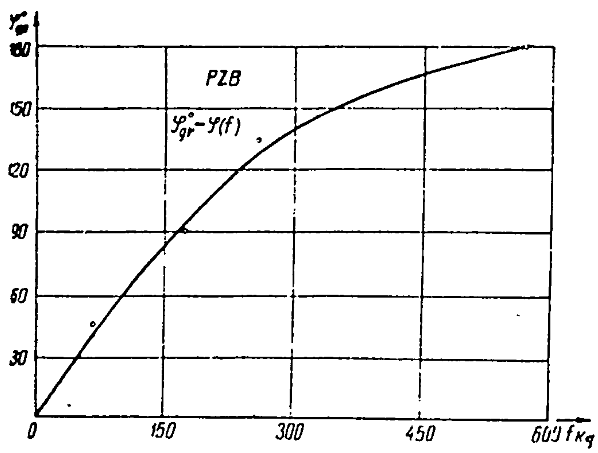

Fig.8

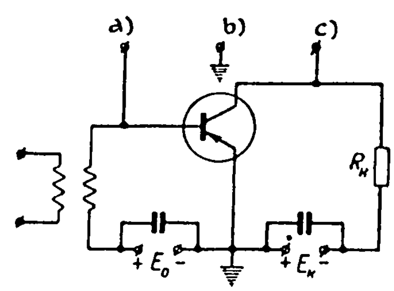

Fig.9

a) Vertical; b) To the oscillograph; c) Horizontal

the ellipses permitted determination of four frequencies at which the angles of phase shift (φ_{dr}) were equal to 0, 45, 90, 135, and 180°. An oscillator of the type I-100 served as the oscillating source.

The total angle $\varphi_s = \varphi_{eo} + \varphi_{dr}$ can be determined from the diagram in Fig.9. In this diagram, the overall phase shift $\varphi_s = \varphi_{eo} + \varphi_{dr}$ is equal to the shift in phase between the excitation voltage $\dot{U}_o$ and the current of the collector I_k (or the voltage under load $\dot{U}_n = I_k R_n$).

This phase shift is easily determined by means of the oscillotrons (see the test for calculating φ_{dr}), or the voltages are given directly at the vertical and horizontal tube plates, passing through the amplifier.

The result of the variations in the angle φ_s for the triode PZB are given in Fig.10. The measurements are made by the very same method used in the previous experiment. It is significant to compare the measured angles φ_{dr} and φ_s at a frequency of approx. 170 kc (Figs.8 and 10). Here we have the angle $\varphi_{dr} \approx 90^o$ and the angle $\varphi_s \approx 135^o$, which, generally, agrees quite well with the theoretical conclusions (see Fig.6, where $\varphi_{eo} < 45^o$ when $\varphi_{dr} = 90^o$).

At the lowest frequencies, the angle φ_{eo} is quite small, and consequently the values of the angles φ_{dr} and φ_s converge (see Figs.8 and 10, for f < 50 kc).

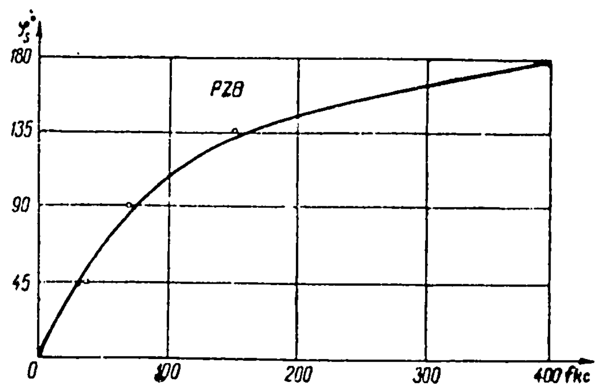

Fig.10

At the highest frequencies (f > > 400 kc) the angle φ_{eo} is greatly reduced (Fig.8) and the values of the angles φ_{dr} and φ_s again converge (Figs.8 and 10).

The total phase shift $\varphi_s = \varphi_{eo} + \varphi_{dr}$ is the shift in phase between the voltage of the excitation source U_o and the first harmonic of the collector current I_{k1}. Applicable to the self-oscillator layout, this phase shift is depicted in Fig.11.

In the known systems of vacuum-tube self-oscillators, the phase shift between the currents $\dot{I}_{a1}$ and the voltage under load (circuit) $\dot{U}$ usually tends toward zero. This indicates that the frequency of the generated oscillations and the natural frequency of the circuit practically coincide. From the view-point of the required

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frequency stability and energy indexes, this system is optimal. However, it is known that, if the sum of the phase angles of the triode φ_s and the feedback φ_k do not equal zero, a separation between the natural and oscillating frequencies will occur. The sign and value of the separation are determined by the equation

$$\varphi_a = \varphi_s + \varphi_k, \qquad (20)$$

where φ_a is the load phase angle.

In principle, the angle φ_a may vary within the limits of $\pm 90^\circ$; however, in practice by virtue of a considerable reduction of the equivalent resistance of the circuit outside of the limits of its zone of passage, the angle φ_a changes within considerably smaller limits. In other words, if it is required to satisfy the phase balance at greater φ_a, then the oscillation collapses due to a disrupted amplitude balance.

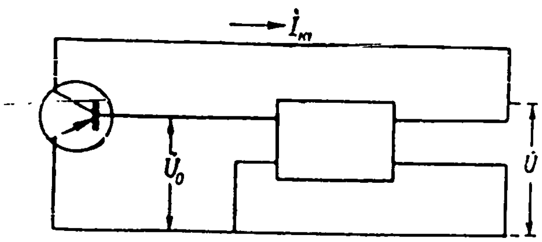

Fig.11

It follows from this that the sum of the angles φ_s and φ_k must also be close to zero [see eq.(20)]. The variations in the angle φ_s for the semiconductor triode were already examined.

In the known systems of single-circuit self-oscillators, the phase angle of the feedback factor can be made to approach 0° or 180°. In two-circuit and three-circuit self-oscillators, the phase of the feedback factor, at loose coupling between the circuits, may vary within wider limits due to a variation in natural frequency of one of the circuits (in the two-circuit system) or of two of the circuits (in the three-circuit setup).

From the above statement it follows that the values $\varphi_s + \varphi_k$ approaching 90° are more disadvantageous, since it is hardly possible to compensate for such a shift in phase [see eq.(20)] in the common self-oscillator. Consequently, in self-oscillators

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with a semiconductor triode, one may expect oscillation failure at frequencies to which $\varphi_s + \varphi_k = 90^\circ$ corresponds.

As mentioned above, in ordinary self-oscillation systems, the value $\varphi_k = 0$ can be considered as the first approximation. Then, the approach of φ_s to 90° denotes the limiting frequency of self-oscillation, if the characteristic $\varphi_s = \varphi(f)$ is used.

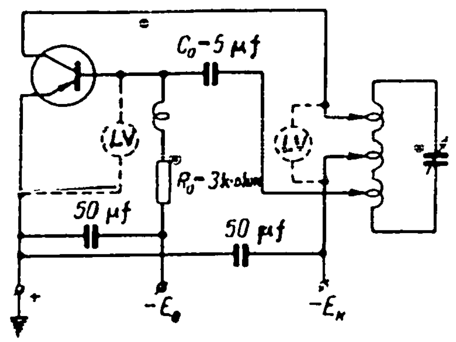

Fig.12

Obviously, an oscillation failure may occur due to an amplitude imbalance. However, we will disregard this point for the time being.

To put all the above material to a practical test, a self-oscillator was selected with an inductive Hartley circuit (Fig.12) in a triode of the PZB type. For this, the earlier characteristic $\varphi_s = \varphi(f)$ (Fig.10) was taken.

Careful measuring showed that the highest cutoff frequency of the oscillator f_{np} is 90 kc. A study of the curve $\varphi_s = \varphi(f)$ shows that the angle $\varphi_s = 90^\circ$, is correlated by a frequency of, for example, 70 kc. The coincidence of the results should be considered satisfactory, if the error in measuring the phase φ_s is taken into consideration according to our accepted method (by placing the large elliptical axis at the tube grid). Besides this, the capacitance C_0 of the separating capacitor (see below) could affect the result.

Investigation of Phase Correction Systems

For phase relation changes in the self-oscillation system, some reactance should be included in the base circuit (Fig.13).

If the following designates

$$\frac{\dot{U}_{en}}{\dot{U}_0} = K e^{i\varphi'_{en}},$$

then it is not difficult to obtain the following expression for the phase angle:

$$\varphi'_{eo} = \operatorname{arc\,tg} \frac{r_o r_{eo} - x_o x_{eo}}{r_{eo} x_o + x_{eo} r_o + x_{eo} r_{eo}}, \tag{21}$$

where r_o, r_{eo}, x_o, x_{eo}, see Fig.13.

Let us examine the variation limits of the angle φ'_{eo} in the case where $0 \leq \varphi_{dr} \leq \pi$. In this case, we have $x_{eo} \leq 0$ (of a capacitive nature). The nature of the value changes $x_{eo} = \varphi(\varphi_{dr})$, $r_{eo} = \varphi(\varphi_{dr})$ is shown in Fig.4.

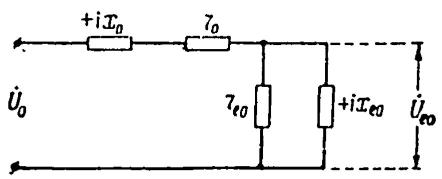

Fig.13

Let there be some auxiliary capacitance included in the base circuit, i.e., let $x_o < 0$. Then, the denominator in eq.(21) will always be less than zero. The subtrahend in the numerator of eq.(21) will be greater than zero ($x_o x_{eo} > 0$). Therefore, at

$$r_o r_{eo} - x_o x_{eo} > 0 \tag{22}$$

the phase angle φ'_{eo} will be optimal, i.e., $\dot{U}_{eo}$ will lag in phase behind $\dot{U}_o$; and at

$$r_o r_{eo} - x_o x_{eo} < 0$$

the phase angle φ'_{eo} will be positive, i.e., $\dot{U}_{eo}$ will lead $\dot{U}_o$ in phase. The first case corresponds to the large capacitance C_o (Fig.12), and the second to the small capacitance. The best results are obtained with a low capacitance C_o.

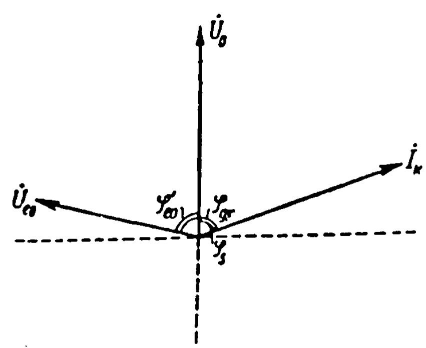

Fig.14

Actually, in this case, x_o will be large, and x_{eo} will also be greatly increased, inasmuch as, at the higher frequencies, we have $\varphi_{dr} \to \pi$ (see Fig.4). The value r_{eo}, on the other hand, will tend toward the minimum.

Therefore, the fractional numerator in eqs.(21) will increase rapidly, while the denominator will rise more slowly. The absolute value of the fraction will increase with decreasing C_o, and a boost in frequency (increase in $\dot{\varphi}_{dr}$), i.e., an advance in $\dot{U}_{eo}$ relative to $\dot{U}_o$, will also increase the value. In Fig.14 the possibility of maintaining phase balance ($\varphi_s < 90^o$), even at angles φ_{dr} approaching 180°, is demonstrated. As a consequence, a considerably more favorable phase balance can be expected in the system with a low capacitance C_o. As a result, higher frequencies can be generated.

Actually, with a capacitor of C_o = 1000 μμf at the triode PZB, steady self-oscillations took place, going as low as a frequency of 300 kc. This limit frequency trebled the cutoff frequency of 90 kc, obtained with the separation capacitor C_o = = 5 μ f when the condition of phase correction was not satisfied.

It would be quite interesting to analyze the experimental data given in Table 2. This Table lists the self-oscillating frequencies and the smallest corresponding feedback factors, insuring operation of the system close to oscillation decay. Besides this, the Table gives the values of the capacitance C_o, supplying the required phase correction. The values of C_o are not critical. A reduction in C_o is a condition for raising the cutoff frequency of self-oscillations.

Table 2

Triode PZB

a)	$K_{oc} = \frac{U_{om}}{U_m}$	C_o, μμf
60	0,0025	10 000
205	0,033	10 000
238	0,0016	5 000
288	0,015	1 000
300	0,038	1 000

a) Frequency of Self-Oscillation, in kc

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A frequency of 205 kc should be considered a limit in this respect for C_o = = 10,000 μμf; a frequency of 238 kc, for C_o = 5000 μμf; and a frequency of 300 kc, for C_o = 1000 μμf. As preliminary data, a further decrease in C_o does not permit an increase in the cutoff frequency of self-oscillation.

The voltages U_{om} and U_m are measured by vacuum-tube voltmeters VKS-7b, installed as shown in Fig.12. Lower values of the feedback factor, at C_o = 5 μf and C_o = 1 μf, and also the corresponding cutoff frequencies are shown in Table 3.

Table 3

Triode PZB

a)	$K = \frac{U_{om}}{U_m}$	C_o, μf	b)
56	0,17	5	
76	0,056	5	
90	0,14	5	c)
56	0,0075	1	
77	0,035	1	
95	0,125	1	c)

a) Frequency of self-oscillation, in kc; b) Remarks; c) Cutoff frequency

In addition to its usefulness as a phase correction circuit, working with a low-capacitance capacitor in the base circuit C_o, the phase correction system can be operated by means of inductance, and, consequently, a large capacitance C_o can be contained in the capacitor. Equation (21), derived above for φ'_{eo}, indicates that in this case (and even at $0 \leq \varphi_{dr} \leq \pi$) the products $x_o x_{eo} < 0$ and the fractional numerator (21) can be converted to zero, if the change $x_o = \omega L_o$ (assuming $\frac{1}{\omega C_o} \to 0$) results in the conditions $r_o r_{eo} - x_o x_{eo} = 0$. The fractional denominator (21) may be equated to zero when $r_{eo}\, x_o = x_{eo}\, r_o + x_{eo}\, r_{eo}$ are larger or less than zero. At large enough a value of x_o the numerator of the fraction (21) will be smaller than zero, and the denominator larger than zero. Thus the fraction will be smaller than zero and the phase angle φ'_{eo} will be less than zero. $\dot{U}_{eo}$ will lag in phase behind $\dot{U}_o$.

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The corresponding vector diagram is shown in Fig.15. From this may be inferred that, in the large zone of the angles φ_{dr}, phase balance may be obtained in the self-oscillator, constructed as an "inverted" circuit.

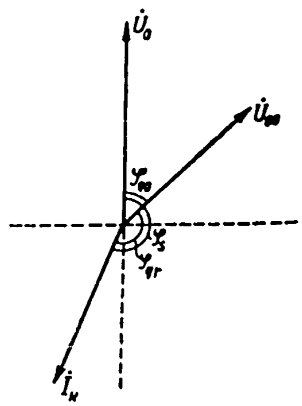

Fig.15

Such a hookup (Fig.16) was tested and showed good results. In the absence of resonance ($r_o r_{eo} - x_o x_{eo} \neq 0$), with inductance in the base circuit L_o, a frequency of self-oscillations of 320 kc (with a PZB triode) was reached. In the presence of "resonance" ($r_o r_{eo} - x_o x_{eo} = 0$), a frequency of 500 kc was obtained. The required information is presented in Table 4.

Table 4

Triode PZB

a)	$K_{oc} = \frac{U_{0m}}{U_m}$	L_o, μhenry	b)
60	0,02	100	c)
82	0,007	100	
105	0,0033	100	
150	0,0053	100	
230	0,012	100	
320	0,03	100	
500	0,062	d)	e)

a) Frequency of self-oscillation, in kc; b) Remarks; c) Lowest value of feedback coefficient; d) "resonance"; e) Limit frequency

At the conclusion of the experiment with the system in Fig.16, the triode PlV was tested with which self-oscillations of a frequency of 1.4 mc were obtained. At this, the inductance, corresponding to the "resonance", was equal to 2 microhenry.

Oscillation Characteristics

In installing a self-oscillating system it is essential to satisfy the requirements not only as to phase balance but also as to amplitude. General directives in this relation can be obtained from the previously discussed oscillatory characteristics $I_{k1} = \varphi(U_{om})$. Since the voltage at the terminals of the collective load (circuit) is proportional only to the harmonic ($U_k = I_{k1} R_e$), the oscillatory characteristics were assumed as $U_k = \varphi(U_{om})$ for R_e, E_k, E_o = const.

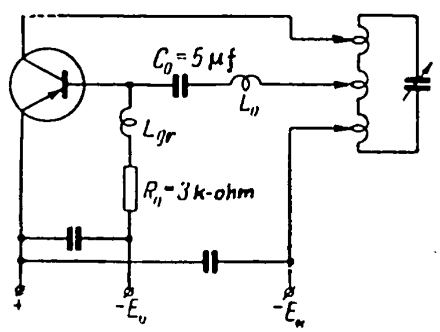

Fig.16

The characteristics for the triode PZB were also taken at a frequency of 23.4°kc, for the different voltage displacements at the base (E_o = 0; +0.2; -0.2 v, cf. Fig.17).

Attention should be paid to the fact that soft self-excitation is possible only when there is a small emission with negative bias in the base. This result could have been foreseen, since the characteristics $I_k = \varphi(E_o)$, at E_k = const start at E_o = 0 and are distributed in the negatively charged area in the base. Greater steepness of the characteristic $S = -\frac{\partial I_k}{\partial E_0}$, necessary for self-excitation, corresponds to a negative value of E_o.

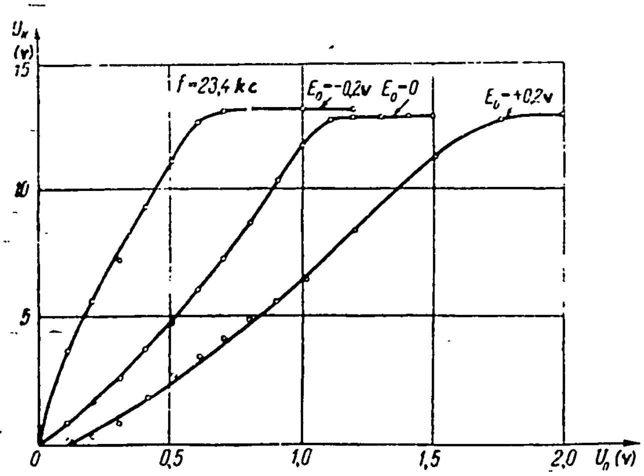

Fig.17

For this reason, the systems of self-oscillation presented above always contain a preliminary negative bias at the base. This is naturally fed from the common source E_k.

Conclusions

1. In self-oscillators with semiconductor junction-type triodes, a disruption

in phase balance may be the cause of oscillation decay at high supercritical frequencies, even leading to an upsetting of the amplitude balance as a result of the power drop of the amplification factor.

2. The use of self-oscillator systems with phase correction in the feedback circuit permits an increase by several times in the cutoff frequencies of self-oscillation. For example, in the self-oscillator with a low-frequency triode PZB, steady oscillations at frequencies up to 500 kc can be obtained, whereas without correction, the cutoff frequency of self-oscillation equals 90 kc.

3. To improve self-excitation conditions in receiving soft self-excitations, it is necessary to feed the base of the triode with some small preliminary negative bias.

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FREQUENCY BAND OF RADIOTELEGRAPH SIGNAL TRANSMISSIONS WITH AMPLITUDE AND FREQUENCY KEYING

by

M.S.Gurevich

The frequency bandwidth is defined which is assigned to radiotelegraph transmissions, with amplitude and frequency keying by rectangular and rounding-off signals.

Introduction

The frequency bandwidth of transmission is important in characterizing the system of radio communication and in indicating the effectiveness of the use of the frequency spectrum. The bandwidth assigned to transmissions is called the frequency band, containing 99% of the transmission power and including any discrete frequency whose power consists of not less than 0.25% of the common transmitted output (Bibl.1 and 2).

The examination presented below is based on the spectra of signal transmissions ("telegraph points"). This method, corresponding to the requirement for "regulation of radio communications" (Bibl.1), easily permits theoretical computations and corresponding measurements. It is known that the spectra of "telegraph points" and the actual communications are closely adjacent. The establishing of closer relations between assigned bandwidth with the transmissions of "telegraph points" and the actual communications demands a special study and is beyond the scope of the present article.

Bandwidths Assigned to Transmissions with Amplitude Keying

It is known that the distribution of energy in amplitude-keyed oscillations is the very same as in the low-frequency keyed signal, excluding the constant multi-

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ple 1/2 which is kept in relation to the keyed oscillations. In the following, the spectra of low-frequency signals will be considered as their equivalent.

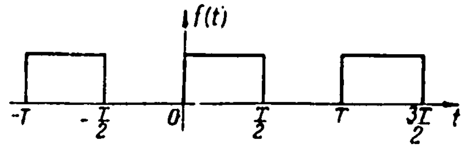

Fig.1

Square-wave signals. When keying by square-wave signals (Fig.1)

$$f(t)=\begin{cases}1 & \text{at} \quad 0<t<T/2 \\ 0 & \text{at} \quad T/2<t<T\end{cases}. \tag{1}$$

Expanding eq.(1) into a Fourier series will yield

$$f(t)=\frac{1}{2}+\frac{2}{\pi}\Big[\sin pt+\frac{1}{3}\sin 3pt+\frac{1}{5}\sin 5pt+\ldots+ \\ +\frac{1}{2n-1}\sin(2n-1)pt+\ldots\Big]. \tag{2}$$

The power corresponding to the component found in the transmission band, calculated by eq.(2), will equal

$$P=\frac{1}{4}+\frac{1}{2}\cdot\frac{4}{\pi^2}\left(1+\frac{1}{3^2}+\ldots+\frac{1}{n_1^2}\right)=\frac{1}{2}\cdot 0{,}99. \tag{3}$$

We find from eq.(3) that $n_1 = 21$.

If B denotes the keying rate, expressed in bauds, then the bandwidth assigned to

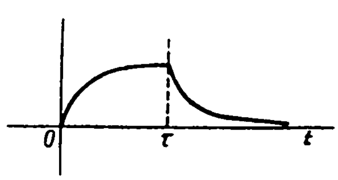

Fig.2

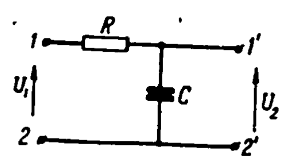

Fig.3

transmission will be equal to

$$\Delta f_n = 21\,B. \tag{4}$$

Rounding-wave signals. Let us examine the signal, depicted in Fig.2 and represented by the functions

$$f(t)=\begin{cases}0 & \text{within the limits } t=-\infty \text{ to } t=0\\ 1-e^{-\frac{t}{\theta}} & \text{"} \quad \text{"} \quad t=0 \quad \text{"} \quad t=\tau\\ \left(1-e^{-\frac{\tau}{\theta}}\right)e^{-\frac{t-\tau}{\theta}} & \text{"} \quad \text{"} \quad t=\tau \quad \text{"} \quad t=+\infty.\end{cases} \tag{5}$$

Signals of such form may be received through the simplest circuit model, consisting of a resistance R and a capacitance C, as shown in Fig.3.

The constant of the circuit time, specified by the rate of signal rounding, equals

$$\theta=RC. \tag{6}$$

Under practical conditions, θ is selected small enough relative to the signal duration that the relation θ/τ will be of the order of 1/5 - 1/40.

Let the periodic voltage of the square-wave form, represented by the series (2), be coupled to the terminals 1, 2 of the circuit (Fig.3), and assume that from these terminals 1' and 2' of the circuit the rounding voltage is taken off. The form of the periodic signals at the circuit output approximates Fig.2; however, for graphic purposes Fig.2 is exaggerated. This signal in actuality subject to much less rounding in the majority of cases.

Let us determine the mean value of power for a period of frequency keying T. If θ/τ is sufficiently small, then

$$P=\frac{1}{T}\int_0^T [f(t)]^2 dt=\frac{1}{2\tau}\left\{\int_0^\tau \left(1-e^{-\frac{t}{\theta}}\right)^2 dt+\int_\tau^\infty \left(1+e^{-\frac{\tau}{\theta}}\right)^2 e^{-2\frac{t-\tau}{\theta}} dt\right\}. \tag{7}$$

After integration, we get

$$P=\frac{1}{2}\left[1-\frac{\theta}{\tau}\left(1-e^{-\frac{\tau}{\theta}}\right)\right]. \tag{8}$$

We next turn to the correlation between output and input voltages of the circuit (Fig.3) which, in a fixed state, can be expressed by the equation

$$\left(\frac{U_2}{U_1}\right)^2=\frac{1}{1+\omega^2\theta^2}. \tag{9}$$

If $\omega = np$, where

$$p=\frac{2\pi}{T}=\frac{\pi}{\tau} \tag{10}$$

the basic angle of keying frequency is formed, so that

$$\left(\frac{U_2}{U_1}\right)^2=\frac{1}{1+n^2p^2\theta^2}. \tag{11}$$

If a square-wave signal (1) is applied to the input of the bend in the circuit, then the following voltage will appear at the circuit output

$$U_2(t)=\frac{1}{2}+\frac{2}{\pi}\Big[\frac{1}{\sqrt{1+1^2p^2\theta^2}}\sin pt+ + \frac{1}{3}\frac{1}{\sqrt{1+3^2p^2\theta^2}}\sin 3pt+\frac{1}{5}\frac{1}{\sqrt{1+5^2p^2\theta^2}}\sin 5pt+\ldots\Big]. \tag{12}$$

The phase ratios of the individual voltage components in the given case are of no interest. The power is determined by the root-mean-square values of this voltage

$$P=\frac{1}{T}\int_0^{\tau}[f(t)]^2dt=\frac{1}{4}+\frac{1}{2}\cdot\frac{4}{\pi^2}\sum_{n=1}^{\infty}\frac{1}{1+n^2p^2\theta^2}\frac{1}{n^2}. \tag{13}$$

If we assume $\frac{\theta}{\tau} = \frac{1}{20}$, then eq.(8) will yield

$$P=\frac{19}{40}. \tag{14}$$

The power in the transmission band, calculated according to eq.(13), is equal to 99% of the common power, determined by eq.(8), i.e.,

$$\frac{1}{4}+\frac{2}{\pi^2}\sum_{n=1}^{N}\frac{1}{N^2}\frac{1}{1+n^2p^2\theta^2}=\frac{19}{40}\cdot 0{,}99=0{,}470. \tag{15}$$

From eq.(15) we find that, to determine the frequency band containing 99% of the power, it is necessary to assume for calculation the first, third, and fifth harmonics, i.e., the bandwidth will be equal to

$$\Delta f = 5B. \tag{16}$$

Producing analogous computations for the other values, we derive the following values for the frequency band assigned to transmission:

$$\left.\begin{array}{lll} \text{at} & \frac{\theta}{\tau} = 1/4.6 & \Delta f = 3B \\ \text{at} & \theta/\tau = 1/40 & \Delta f = 7B \end{array}\right\}. \tag{17}$$

Equations (4), (16), and (17) indicate that the utilization of even the simplest shaping circuit lowers the transmission bandwidth by several times. For practical purposes, more complex versions of shaping circuits are employed. The transmission bandwidth, obtained by their use, can be determined by similar methods in each individual case.

Bandwidth Assigned to Transmissions with Frequency Keying

Square-wave signals. As is known (Bibl.3), for keying without an interruption of phase by rectangular dotting, the signal may be represented by the series

$$\begin{aligned} f(t) = \frac{2}{\pi}\Big\{\frac{m}{m^2}\sin\left(\frac{\pi}{2}m\right)\cos\omega t + \\ + \frac{m}{m^2-1^2}\cos\left(\frac{\pi}{2}m\right)[\cos(\omega-p)t - \cos(\omega+p)t] - \\ - \frac{m}{m^2-2^2}\sin\left(\frac{\pi}{2}m\right)[\cos(\omega-2p)t + \cos(\omega+2p)t] - \\ - \frac{m}{m^2-3^2}\cos\left(\frac{\pi}{2}m\right)[\cos(\omega-3p)t - \cos(\omega+3p)t] + \ldots\Big\}, \end{aligned} \tag{18}$$

where p is the keying frequency;

m = 2D/B is the modulation index;

D is the frequency deviation (half the difference of the operating frequencies).

For all values of m, eq.(18) may be written more compactly in the form

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$$f(t)=\sum_{n=-\infty}^{+\infty}\frac{2}{\pi}\frac{m}{m^2-m^2}\sin(n+m)\frac{\pi}{2}\cos\left[(\omega+np)t+(n+m)\frac{\pi}{2}\right]. \qquad (19)$$

In accordance with eq.(19), the amplitude of the n^{th} harmonic of the spectrum is equal to

$$A_n=A_{-n}=\frac{2m}{\pi}\left|\frac{\sin(n+m)\frac{\pi}{2}}{n^2-m^2}\right|. \qquad (20)$$

Let us determine the mean value of power for a period of frequency keying: If m is a whole number or a half-integer then

$$P=\frac{1}{T}\int_0^T[f(t)]^2dt=\frac{1}{T}\left[\int_0^{T/2}\cos^2 2\pi Dt dt+\int_{T/2}^{T}\cos^2(-2\pi D)t dt=\right.$$
$$=\frac{2}{T}\int_0^{T/2}\cos 2\pi Dt dt=\frac{1}{2}. \qquad (21)$$

On the other hand,

$$P=\frac{1}{T}\int_0^T[f(t)]^2dt=\sum_{n=-\infty}^{+\infty}\left[\frac{2m}{\pi}\frac{\sin(n+m)\frac{\pi}{2}}{n^2-m^2}\right]^2=$$
$$=\frac{4}{\pi^2}\left\{\frac{1}{m^2}\sin^2\frac{m\pi}{2}+2m^2\sum_{n=1}^{+\infty}\left[\frac{\sin(n+m)\frac{\pi}{2}}{n^2-m^2}\right]^2\right\}. \qquad (22)$$

The power computed according to eq.(22) must be equal to 99% of the power determined by eq.(21), i.e.,

$$\frac{4}{\pi^2}\left\{\frac{1}{m^2}\sin^2\frac{m\pi}{2}+2m^2\sum_{n=1}^{n_1}\left[\frac{\sin(n+m)\frac{\pi}{2}}{n^2-m^2}\right]^2=0{,}99\cdot\frac{1}{2}. \qquad (23)\right.$$

From eq.(23) it is possible to determine the n_1^{th} component of the spectrum, at which this equation is satisfied.

The above discussion does not take into consideration the natural rounding of signals in the circuits of the transmitter, which takes place even in cases where no special rounding circuits are used. To determine the bandwidth assigned to transmissions in consecutive level and square frequency-keyed signals, the MKKR recommends (Bibl.2) the following formula for calculating the natural rounding signals in the circuits of the transmitter

$$\Delta f = \begin{cases} 8/3\, D + 4/3\, B & \text{at} \quad 2{,}5 < m \leqslant 8 \\ 2{,}2\, D + 3{,}2\, B & \text{at} \quad 8 \leqslant m \leqslant 20 \end{cases}, \tag{24}$$

where Δf is the bandwidth assigned to transmissions;

D is the frequency deviation (half the difference of the operating frequencies);

B is the rate of keying, expressed in bauds;

$m = 2D/B$ is the frequency modulation index;

Equation (24) is more conveniently represented, in a form slightly different from that used by the MKKR: Expressing the bandwidth through the frequency modulation index in units of keying speed B; we then obtain

$$\Delta f = \begin{cases} 4/3\,(1 + m)\, B & \text{at} \quad 2{,}5 < m \leqslant 8 \\ (3{,}2 + 1{,}1\, m)\, B & \text{at} \quad 8 \leqslant m \leqslant 20 \end{cases}. \tag{25}$$

Rounding signals. We will examine the case of frequency modulation of a single frequency, corresponding to the greatest rounding which can be obtained by means of shaping signals. Such a transmission, with an accuracy up to the constant of amplitude factor, corresponds to the function

$$f(t) = \sin(\omega t + \varphi_0 + m \sin pt), \tag{26}$$

where t is the time, and ω, φ_0, p, and m are any real and positive constants, whereas m is the modulation index.

It is known that the function (26) may be expanded into a converging series of

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sine functions of time, i.e., may be represented as a discrete infinite spectrum

$$f(t) = \sum_{n=-\infty}^{+\infty} J_n(m) \sin[(\omega + np)t + \varphi_0], \tag{27}$$

where n is a whole number and $J_n(m)$ denotes a Bessel function of the first kind, with the series n and the argument m.

Thus the mean power, corresponding to each frequency, is equal to the square of the amplitude $J_n(m)$, correct to the constant factor, independent of n; the common power, corresponding to the series (27) is expressed, correct to within the same constant amplitude factor, by

$$J_0^2(m) + 2\sum_{n=1}^{\infty} J_n^2(m), \tag{28}$$

since

$$J_{-n}(m) = (-1)^n J_n(m).$$

The series (28) converges for any value m, and its sum, independent of m, always (Bibl.4, p.40) has the simple form

$$J_0^2(m) + 2\sum_{n=1}^{\infty} J_n^2(m) = 1. \tag{29}$$

The bandwidth, fixed in conformity with the specifications in "Regulating Radio Communication", will therefore be determined by the smallest whole values of n, positive or equal to zero, which satisfy the following two conditions

$$\sum_{p=n+1}^{\infty} J_p^2(m) \leqslant 0{,}005, \tag{30}$$

$$J_{n+q}^2(m) \leqslant 25 \cdot 10^{-4}$$

or

$$J_{n+q}(m) < 0{,}05, \tag{31}$$

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where q is any whole positive number.

Equation (31) contains an infinite number of conditions, since q is an arbitrary positive number. Therefore we will replace its two conditions, which do not already contain this indeterminate quantity. For this purpose, we will use the known results pertaining to the function variations $J_n(m)$, dependent on n for a constant value of m (Bibl.5).

These results infer that any function $J_n(m)$, for which m is an arbitrary constant and n is a variable, attain their maximum amplitude when $n = n_1$ (after which, as this value oscillates* somewhat, the amplitude and pseudoperiod of these is boosted), this maximum amplitude becomes positive, whereupon $n_1 < m$. For n_1, the correctly approximated equation is

$$n_1 \approx m - 0{,}8086\, m^{\frac{1}{3}}. \tag{32}$$

Furthermore this function, remaining positive, decreases with an increase in n and rapidly converges toward zero.

The maximum amplitude diminishes with a rise in m:

$$J_{n_1}(m) \approx 0{,}6719 \frac{1}{m^{\frac{1}{3}}}. \tag{33}$$

Thus, if $n + 1 \geq n_1$ and $J_{n+1}(m) < 0.05$, then by that much more $J_{n+q}(m) < 0.05$. Consequently, eq.(31) may be substituted by the two inequations

$$n + 1 \geq n_1; \quad J_{n+1}(m) < 0{,}05. \tag{34}$$

Besides, if $J_{n_1}(m) < 0.05$, both conditions (34) will disappear and n may be assumed to be equal to zero, since $|J_n(m)| < 0.05$ for any limits of n.

It follows from eq.(33) that the case where both conditions vanish, takes place when $n \geq 2460$. Naturally, eq.(30) still retains its validity in the face of all this.

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*The number of these oscillations grows with an increase in m.

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Let us discuss first the inequations (34). The Tables by Yanke - Emde (Bibl.6) for $0 \leq m \leq 29$ directly give solutions for whole values of m.

Another problem is that of solving the inequations (30). First, we will express every left-hand term of this equation as a Neumann series (Bibl.4, p.43)

$$J_p^2(m) = \left(\frac{m^2}{4}\right)^p \sum_{q=0}^{\infty} \frac{(2p+2q)!\left(-\frac{m^2}{4}\right)^q}{q!\,(2p+q)!\,[(p+q)!]^2}. \tag{35}$$

In this way we have represented the left-hand side of eq.(30) in the form of a series of positive terms of an absolutely convergent series. Further, the terms containing m may be grouped together in a single step. As the result of such a grouping, we obtain

$$\sum_{p=n+1}^{\infty} J_p^2(m) = \left(\frac{m^2}{4}\right)^{n+1} \sum_{q=0}^{\infty} \frac{(2n+2q+1)!\left(-\frac{m^2}{4}\right)^q}{q!\,(2n+q+1)!\,[(n+q+1)!]^2} =$$

$$= \frac{\left(\frac{m^2}{4}\right)^{n+1}}{[(n+1)!]^2}\left\{1 - \frac{2n+3}{(n+2)^2}\,\frac{m^2}{4} + \frac{(2n+4)\,(2n+5)}{2\,[(n+2)\,(n+3)]^2}\left(\frac{m^2}{4}\right) - \ldots\right\}. \tag{36}$$

This expression represents the diminishing function of its index n + 1 and converges toward zero with a rise in n.

The sign-variable series (36) is convenient for use only from m = 0 to m = 4. At higher values of m, its use becomes difficult, and hence the series converges slowly.

To solve the inequation (30), we proceed in the following way with the relations of the value m:

1) for $0 \leq m \leq 4$ we use the sign-variable series (36);

2) for $5 \leq m \leq 29$ the Yanke - Emde Table can be used.

In conclusion there remains only the simultaneous computation of the inequations (30) and (34). For low values of m, eq.(34) is somewhat more limited than eq.(30). For greater values of m, eq.(30) is determined by n.

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Thus we have determined the bandwidth of frequency keying transmission with dotting and rounding signals. Figure 4 gives a comparison of the results of all three ratings presented, where curve 1 corresponds to the sine signals, curve 2 refers to the square-wave signals, and curve 3 is calculated according to eq.(25).

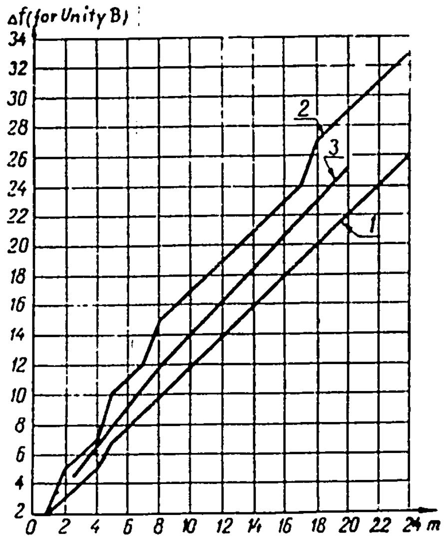

Fig.4

Rounding Signals with Amplitude and Frequency Keying

As indicated by the graph in Fig.4, the quantitative results of the rounding of signals with frequency keying are less substantial than with amplitude keying, where, obviously, the rounding of signals reduces the bandwidth by several times.

However, with amplitude keying in the common transmitter it is not easy to arrive at a convenient measure. Keying is commonly produced by master stages, operating in class C, which renders rounding immediately after the keying stage ineffective. The inclusion of the rounding circuit between the transmitter and antenna encounters considerable difficulty in practice.

It should be stated that there is another fact rendering the use of the rounding circuit with amplitude keying difficult. With considerable signal rounding, a pulse with sharp peaking is obtained. Such a pulse is unsuitable for operating conditions, since with the change in signal level, which occurs in fading, the length of signals received will be altered. This entails predominance of error. If we take this into consideration, then only small rounding signals can be used with amplitude keying, within the limits shown in the first section.

Nonlinear amplification of the transmitter does not build up side frequencies at its output, which are absent at the input voltage. Consequently, there is no need to emit considerable side frequency to obtain flat-top pulses.

Hence frequency keying makes it easier to produce signal rounding than does amplitude keying.

Comparison of the Bands for Both Types of Keying

Using the correlations presented above, the numerical correspondence between the bandwidths assigned to transmissions with amplitude and frequency keying can be established. It is easiest to carry out this comparison for signals of square-wave shape. As indicated by eq.(4), the bandwidth assigned to transmissions with amplitude keying is equal to Δf_a = 21 B. From curve 2 in Fig.4 it is apparent that such a bandwidth with frequency keying transmission corresponds to a modulation index of m = 14.

With this critical value of the modulation index, the bandwidths assigned in both cases are equal. At m < 14, the transmission with frequency keying takes up less band than with amplitude and, conversely, at m < 14, amplitude keying transmission offers better use of the frequency band.

In radio communication, a considerable scattering of the frequency swing D = = 500 cycles takes place. At m = 14,

$$B = \frac{2 \cdot 500}{14} = 71.5 \text{ bauds}$$

Thus, at D = 500 cycles, the frequency keying has the advantage of using the assigned frequency band at speeds exceeding 71.5 bauds.

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CONTACTLESS SWITCHES, USING A TRANSFORMER FERRORESONANT CIRCUIT

by

Ya.G.Koblentz, D.A.Yakovenko

Trigger and counter circuits, forming the basis of the transformer ferroresonant circuit, suggested by the authors, are discussed. These give an improved transmission factor and permit the use of inexpensive ferrite cores. The systems developed can be utilized for ATS (dial office) control devices and in other branches of technology.

Introduction

The use of electronic and contactless facilities opens significant perspectives in automatic telephone station technology, since the new switching elements can be constructed anywhere with high speed, low bulk, and incomparably longer life than electromechanical devices. The search for new switching means was carried out in several parallel directions (Bibl.1), of which the application of semiconductors and ferromagnetic elements are of special interest. The present article develops one of the methods of magnetic switching, namely the construction of ferroresonant devices of automatic control.

The jumps in current or voltage, due to ferroresonance can be utilized in setting up the relay systems, with two steady conditions (Bibl.1). The earlier known system of Isborn-Rutiskhauzer contains as its basis a series circuit LC (ferroresonance of voltages). The system by Dvinker (Bibl.2) published in 1955, is characterized by its not using the basic ferroresonance process in a series circuit consisting of a capacitor and a ferromagnetic coil but, rather, by using nonlinear oscillations, generated in the same circuit by the combined activity at its core of the feeding AC and magnetizing DC.

The investigation of magnetic switching methods, conducted the laboratory of the new ATS systems NIITS (formerly, LONIIS) of the Ministry for Communications of the USSR, have shown the possibility of constructing a series of systems and stations of automatic control as a basis for the transformer ferroresonant circuit (Bibl.1) proposed by the authors. In this, nickel-zinc and manganese-zinc ferrite ring cores were used with a surface diameter of 4 and 7 mm, being fed alternating current with a frequency of about 100 kc. In comparison with the previously known systems, this arrangement significantly raised the transmission factor*.

The Transformer Ferroresonant Circuit

Figure 1a shows the basic element used in constructing the ferroresonant switching systems described below. The primary winding of the saturation transformer is fed by a sinusoidal current. The secondary winding is charged with capacitance, forming a closed ferroresonant circuit L_k C_k. The primary winding can be replaced by one "loop", a wire piercing the core, as shown in Fig.1b. This permits a convenient feeding of a row of cores from a common oscillator, with a minimum of connections between the cores through an oscillator circuit. The transformer ferroresonant circuit shown may be inserted into an equivalent parallel circuit hockup (Fig.1c) and the generated oscillations can be regarded

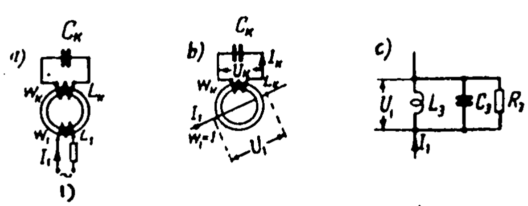

Fig.1

1) Generator

*The transmission factor of the contactless binary element will be called by us (Bibl.1) the ratio of the current or voltage in the operating condition of the system (condition "1") to the corresponding value in the initial state (condition "0")

$$k = \frac{I_{(1)}}{I_{(0)}} \quad \text{or} \quad k = \frac{U_{(1)}}{U_{(0)}}.$$

as ferroresonant currents. The graph analysis and vector diagram are presented in Fig.2, and the ferroresonance loop in Fig.3. The plotting of similar graphs is presented in greater detail in (Bibl.1).

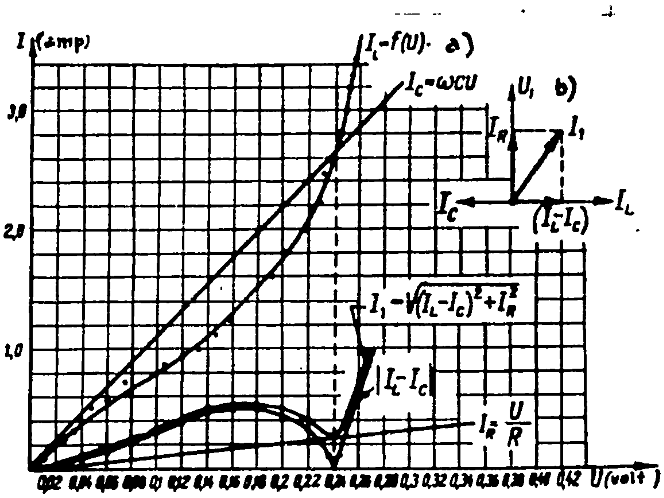

Fig.2

a) $I_L = f(U)$ - Equivalent data in 1st harmonic;

b) Vector diagram

Figure 3 testifies as to the presence of sudden voltage changes (indicated by arrows) with constant amplitude in the feeding alternating current I_1. Points B and F can be regarded as the conditions 0 and 1 of the binary switching element, and the intermediate point D, located in the descending branch of the volt-ampere characteristic, is proportional to the unsteady field which instantaneously runs over the system as soon as the switching unit operates.

In Fig.4 ferroresonance loops, obtained experimentally for the ferrite core at a frequency of 60 kc, are plotted.

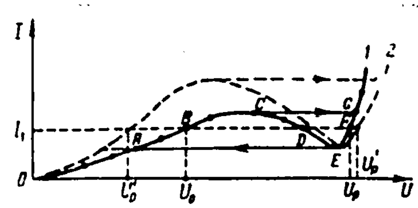

Fig.3

Boosting the Surge Factor

It can be seen in Fig.4 that the voltage surges in the ferroresonant circuit with a ferrite core are relatively small. The transmission factor will have the form

$$k = \frac{U_p}{U_0} = 2 \div 3{,}5$$

dependent on the matching of circuit elements and the AC power. This factor may be increased somewhat by the use of special magnetic materials (molybdenum, Permalloy, superpermalloy, and others) gaged at k = 6 – 10, which is also inadequate in the majority of cases. Investigations reveal that the transmission factor can be significantly raised, if instead of the fundamental frequency higher harmonics made up of ferroresonant oscillations are used.

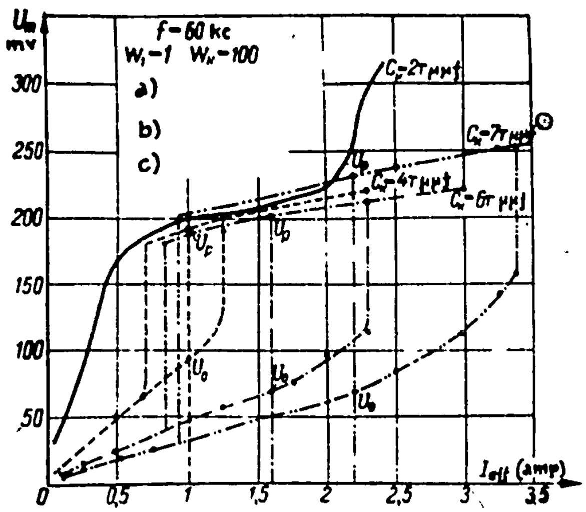

Fig.4

a) Sinusoidal current in primary winding; b) Core d_{mean} = 0.6 cm; c) Ferrite NTS-1000

Let us take a look at the nature of the current and voltage oscillations in the various parts of the system. In Fig.5 oscillograms are depicted: In these, a) is the current I_1 in the primary winding; b) is the voltage in the circuit U_k; c) is the current in the ferroresonant circuit I_k for both steady states of the system, conventionally termed "initial" (left) and "working" (right). It is apparent that, at constant current in the primary winding, the voltage in the circuit varies in sudden starts. The amplitude then changes, for example by doubling itself. This conforms to the data on the experimental ferroresonant loops, presented in Fig.4. The current in the circuit W_kC_k is subject to even more abrupt variations: The amplitude I_k in the working state increases more than 10 times.

If, in the initial state, the oscillations U_k and I_k approach the sinusoidal form, the presence of ferroresonance will cause a considerable distortion of the form of current and voltage due to the appearance of higher (odd) harmonics. The current in the circuit I_k is particularly rich in higher harmonics. For example, according to results of experiments in the working state, the amplitude of the current of the third harmonic exceeds the amplitude of the first harmonic by 3.2 times; that of the fifth harmonic, by 3.3 times; that of the seventh, by 3.0 times, etc.

Even the 19^{th} harmonic still has an appreciable value and contains about 0.6 of the original current.

If the first harmonic is excluded and only the higher ones, beginning with the

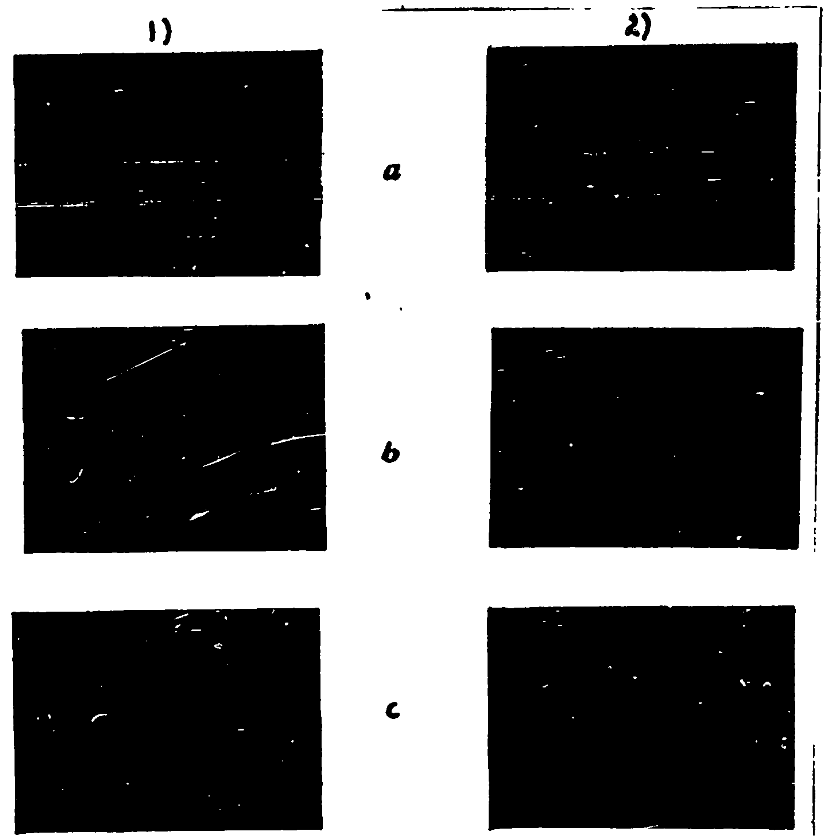

Fig.5

1) "Initial" state; 2) "Working" state

third, are used for switching, then the transmission factor sharply increases and may be reduced to k = 50 - 100.

Let us examine how a boosted transmission factor can be obtained in the systems at the expense of excluding or partially suppressing the first harmonic. If the load is directly switched into the circuit W_kC_k (Fig.6a), a transmission factor of k = 6 - 10 can be obtained. However, the use of such a switching of the load is always limited. By means of an auxiliary transformer (Fig.6b) a much better matching of the load with the circuit can be attained. A considerable increase in the transmission factor is obtained by suppressing the first harmonic by means of the system shown in Fig.6c, where two ferroresonant transformers are used; the first (A) is controlled by the starting winding, while the second (B) has setup in which no nonlinear oscillations can arise. The output windings of both transformers are inserted opposite to each other and consequently to the load Z_n. When both ferroresonant circuits are in the initial state, the current in the load will be minimal since the first harmonics of the voltage in both output coils, as an example, are equal in amplitude and are in antiphase. If the core A were excited by a starting pulse and intense ferroresonant oscillations were generated in it by higher harmonics, then the oscillations would not be compensated by the second core; this will cause a current to flow in the load which is only partially excited by the first harmonic and is mainly excited by higher harmonics at the output winding of the core A. By this means, a high transmission factor may be obtained: k = 50 and more. The use of a second ferroresonant circuit, analogous to the first, makes it possible to get even closer phase angles with compensation. Another method of suppressing the first harmonic is shown in Fig.6d. The circuit of the output

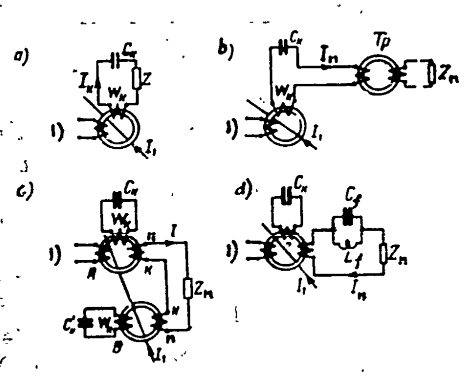

Fig.6

1) Starting winding

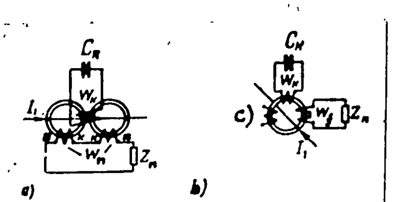

Fig.7

a) Operating principle; b) Conventional representation of the systems;

c) Starting winding

winding of the ferroresonant transformer, placed in series to the load, contains the extremely simple filter C_fL_f set up in the first harmonic.

There remains to be demonstrated another method of increasing the transmission factor (Fig.7). The ferroresonant transformer core consists of two ends. The output winding is composed of two parts and is coiled separately at both ends, so that the number of turns on one of the ends is somewhat larger (10 - 20%) than on the other. Then the coils are put together in such a way that both parts of the output winding are opposite the connected ones, and that all remaining windings are wound on both ends like to a common core. The effect of such a connection is an increase in frequency, sharply raising the asymmetry of the output winding. In the initial state of the system, the first harmonic is considerably weakened. Under working condition (with generation of nonlinear ferroresonant oscillations), the higher harmonics in both output windings are considerably less compensated than the first, and current will flow in the load. By this means, a transmission factor of k = 20 - 30 may be obtained in practice.

The Construction of Several Control Systems

The switching of loads into the transformer ferroresonant circuits described in the preceding portion of the system appears as a simple ferroresonant relay (binary elements). From these elements more complex units can be built.

We will show the construction of the computer-storage circuit. In such a system, when the first pulse enters, the primary element is switched into working condition, with the second pulse, the second element, and so on; this means that the number of cells converted to working condition is equal to the number of pulses received. This principle can be applied to control procedures, for example in the ATS register controller. Figure 8 gives a diagram of one of the variants of a ferroresonant computer storage circuit. It is not difficult to see that this hookup consists of elements similar to the elements in Fig.6c, whereas rectifiers are introduced (selenium or other kinds) in the circuit connections between neighboring cells. In each connecting circuit between two neighboring cells, the rectifiers are inserted facing each other and, in the initial stage, do not conduct alternating current. The control pulses displace the valves in the direction of the flow, thus ensuring conductivity for alternating current between adjacent transformers.

The system is built on a two-cycle principle. Let us assume that the pulse

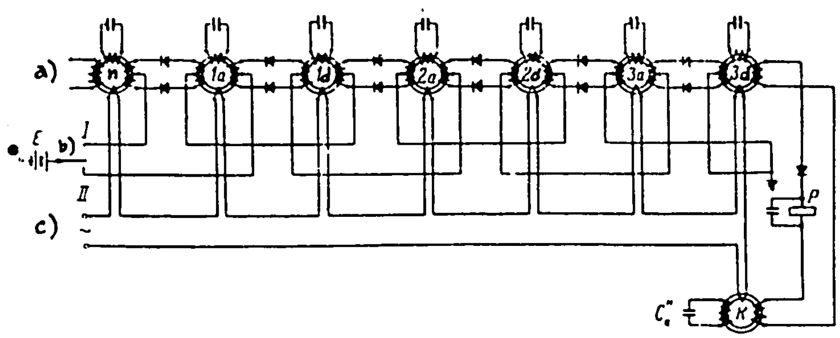

Fig.8

a) Start; b) Key; c) Oscillator

emitter, conventionally called a key and represented in the system as a switch contact, sends out a negative pulse along the lead I. Then the valves are closed in all noncomputer connections. If all ferroresonant circuits are in the initial state, no current will flow in the connecting circuit, since as the windings are connected in opposition and the first harmonics emf are compensated. If this system would first be excited by any circuit, for example, with the initial core H by means of its starting winding, then the higher harmonics, induced in the circuit winding, would not be compensated and would generate current by means of which the adjacent cell would be switched. The process cannot extend any farther, since the next circuit is closed. By switching the "key", the pulse is sent to the wire II and, in the working state, is fed to the next cell.

In this manner, with each pulse the number of switched cells will increase by one unit. The given hookup can work equally from "left to right" or from "right to left", depending on which side gives the initial starting pulse. From any of the transformers, the result may be considered one of the methods presented in Figs.6 and 7. The diagram in Fig.8 shows the switching of a performing device - a relay at the output of the following (seen from the right-hand side) element. For this circuit, an auxiliary ferroresonant transformer K was used, hooked up like the transformer B in Fig.6c. When using ferrite cores with an outside diameter of 7 mm, this arrangement ensures the expected performance of a normal telephone relay.

In constructing the nodes of the apparatus, it is very important to have the so-called trigger system, represented by two elements whose form is always inverse. The trigger must allow suitable switching from one form to the other. For ferroresonance systems, this last requirement is one of the most difficult to satisfy. In contrast to (Bibl.2), we set the task of constructing a trigger ferroresonant system with ferrite cores using fundamental ferroresonant oscillations of greater intensity. The transformer setup allowed this to be done. The schematic of the investigated trigger is shown in Fig.9.

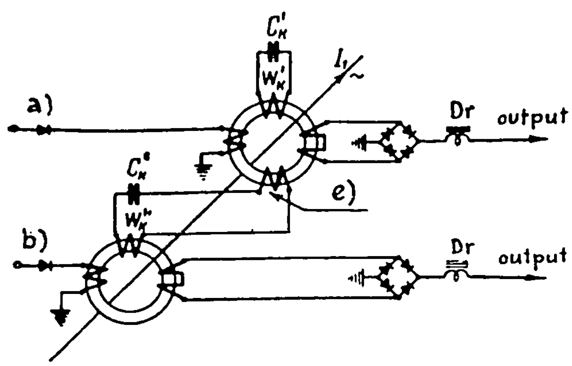

Fig.9

a) Input 1; b) Input 2; c) Connection winding

Two cells were used here, as illustrated in Fig.7, since the circuit with the low cell is connected with the upper one by means of an auxiliary winding in the upper core (connecting coil). The interconnected windings of the different cores are switched relative to the current I_1, common to both, facing each other. This system has the following important advantage: In switching on the feeding current I_1, ferroresonance is generated in one of the transformers (usually in the upper one, be-

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cause of $C_k' < C_k''$). In the outlet winding, at the expense of the higher harmonics, a current is generated which is rectified by the bridge and flows through the output (1) into the load. In sending the primary starting pulse to the input (2) ferroresonance is excited in the lower transformer. The intensive oscillations, generated in the circuit $W_k''C_k''$ and activating the upper core through the connection coils, disrupt the ferroresonance in the upper transformer. In the output winding of the upper transformer, the higher harmonics disappear and the current at the output (1) decays. On the other hand, at the output (2), the transition to the working state of the lower circuit causes a current to flow into the load. If a pulse now is supplied to the upper winding, then the upper circuit changes to a working condition, while the lower returns to the initial state, etc. Chokes serve to separate the direct component entering the load and impede penetration by means of a high-frequency current. The given system may be used as a contactless relay with a holding device and cutoff. This hookup has an amplifying effect: For starting at the input, a smaller current is needed than is being fed at the output. For example, with a core diameter of 7 mm for starting, about 5 ma are enough, while at the output 20 - 30 ma are picked up. In this manner from the output of one trigger (relay), it is possible to start several other analogous systems.

At the base of the ferroresonant relay (Fig.9), the system for dividing the frequency in "half" (trigger with a common counter input) is depicted in Fig.10. In the initial state, the output resistance of the upper circuit has a constant voltage U_3 which is applied to the upper diode D_1, displacing its working point in the cutoff region. When the input pulses arrive, the amplitude of their voltage picks up less cutoff voltage U_3, the diode D_1 is closed, and the counter pulses no longer flow through the input winding of the upper circuit (input 1). At the same time, the blocking voltage U_3 of the lower diode D_1 becomes equal to zero, and the arriving pulse flows through the input winding of the lower circuit, causing it to switch into the condition "1", and the upper circuit into "0". For ordinary functioning,

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it is necessary that the duration of the input pulses be equal to the trigger switching time; therefore, they are conducted across the differentiating circuit R_1C.

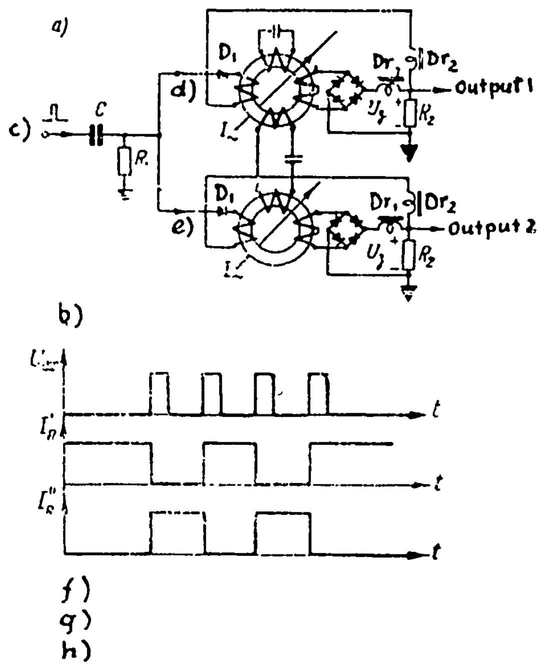

Fig.10

a) Hookup; b) Operating diagram; c) Main input; d) Input 1; e) Input 2; f) U_{in} = Voltage of input pulses; g) I'_R = Current at output I; h) I''_R = Current at output II

When the following pulse enters the lower diode D_1 it will turn out as the blocked voltage U_3, and the entering pulse proceeds through the input winding of the upper circuit, changing it into condition "1" and the lower circuit into "0", and so on. With the arrival of the continuous series of counter pulses at the input of this trigger, the pulses will emerge from the output with half the frequency. The operating diagram of the trigger with a common counter input is shown in Fig.10b. The chokes (Fig.10a) serve to remove the influence of the input windings on the diodes and on the differentiating circuit.

At the base of the trigger with a double input, a distributing counter system was constructed (Fig.11). The counter pulses are conducted parallel to both cells inputs, since the input circuit of each cell consists of the diode D_1 connected in series, the input winding of the lower circuit, the choke Dr_1, and the output resistance R_1 of the preceding cell.

In the initial condition, all upper circuits are in condition "1" and, consequently, all resistances R_1 will have a voltage U_3, blocking the diodes D_1 of the input windings of the lower circuits. On arrival of the counter pulses, whose voltage amplitude is lower than U_3, current no longer flows in the input winding, and

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the system remains in its former state. If one of the upper circuits (for example, cell I) is in the condition "0" and, as a consequence, the lower circuit is in the state "1", then the voltage U_3 in the upper output of the first cell will be equal to

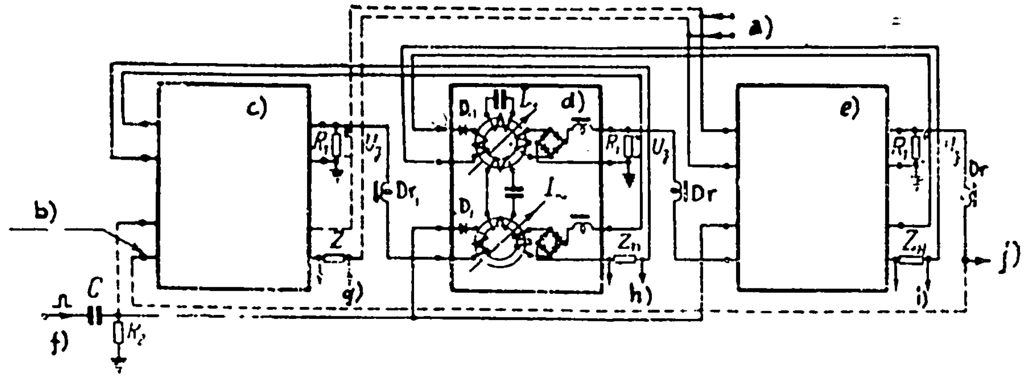

Fig.11

a) From the following cell (Quencher); b) From the previous cell; c) Cell I; d) Cell II; e) Cell III; f) Counter input; g) Output 1; h) Output 2; i) Output 3; j) To the following cell

zero. In this case, the diode D_1 of the second cell is not closed during this time, while all the other diodes D_1 are closed. The first arriving counter pulse, passing through the inlet winding of the lower circuit of the second cell, changes it to con-

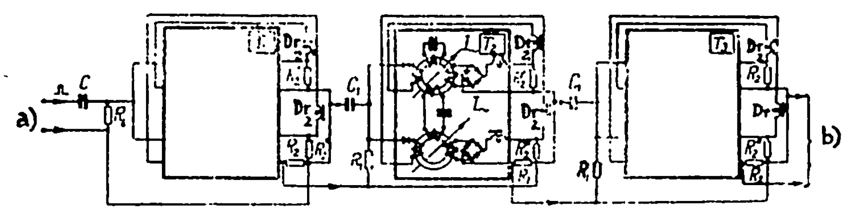

Fig.12

a) Counter input; b) To next trigger

dition "1" and the upper circuit to "0". At the output of the lower circuit of the second cell, a constant voltage appears which passes to the input winding of the upper circuit ("quencher" circuit) of the first cell. As a result, the upper circuit of the first cell switches to condition "1", and the lower circuit to "0". In this

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way, the first counter pulse turns the lower circuit of the first cell to "0", and the lower circuit of the second cell to "1", by this same "displacement" shifting the stored information one step to the right along the counter circuit. The dotted lines in Fig.11 show the connections required for operating this system as a ring counter.

The base of the triggers with a common counter input (Fig.10), connected in stages, carries the binary counter system shown in Fig.12. The voltage coming from

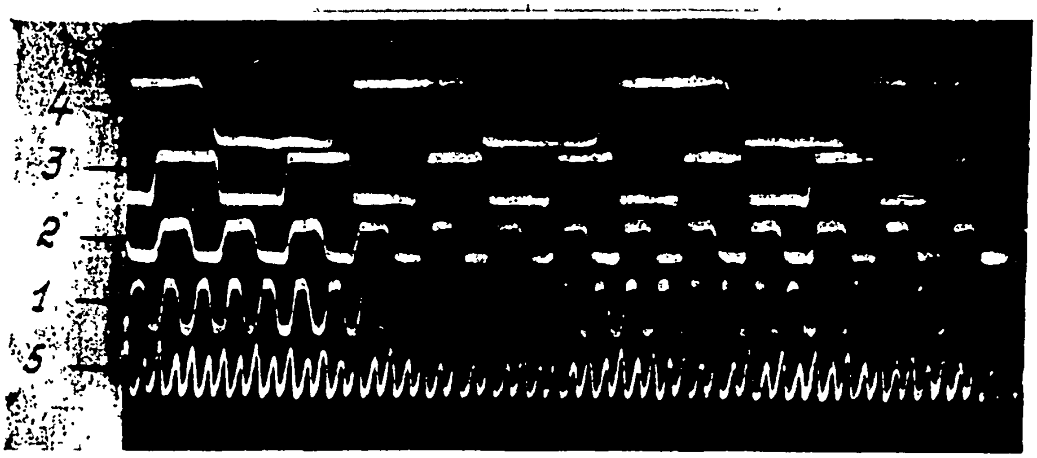

Fig.13

the resistance output R_2 of the first trigger is differentiated by means of the circuit R_1C_1 and is transmitted for starting the second trigger, while the voltage from the resistance R_2 of the second trigger is used for the starting of the third trigger, and so forth.

In supplying an uninterrupted series of pulses to the input of such a system, the frequency of the pulse sequence at the output of the next trigger will be half that at the input, where n is the number of stages. Figure 13 gives an oscillogram, illustrating the splitting "in half" after each trigger cell. Curves, 1, 2, 3, and 4 represent the trigger currents T_1, T_2, T_3, T_4, while in curve 5 the current pulses at the input of the system is given.

Laboratory tests of the described systems indicate their efficiency with a frequency of control pulse sequence of about a thousand per second.

Structural Elements and Units

The above results were obtained with the use of ferrite ring cores, having an outside diameter of 7 or 4 mm, ceramic capacitors, and semiconductor rectifiers. These semifinished parts are extremely inexpensive. In Fig.14, the basic semifinished pieces are shown (from right to left: ferrite cores, ceramic disk capacitors, selenium and copper-oxide valves).

The following basic requirements are made on the ferrite cores: Low power factor, relatively high inductance, stable magnetic properties with a temperature raised up to 75°C. Satisfactory results were obtained by using nickel-zinc ferrite of the NTs-1000 type (Bibl.3) having, at a frequency of 10 kc and a field strength of 2 erg, a dynamic power factor of the order of 0.35 erg and an inductance of $B_m \approx$ $\approx$ 2400 gauss. For the quality factor of a valve for the ferroresonant system, selenium and copper-oxide disks with a diameter of about 7 mm proved suitable. For chokes in the by-passing circuits, ferrite cores were also utilized.

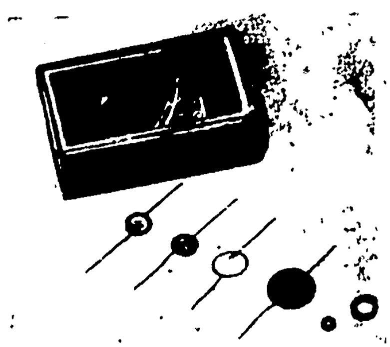

Fig.14

The cells and the small units of the systems were mounted with individual small-size structural members. The conventional shape of one of these structural elements (the tablet), mounted according to the diagram of Fig.10, is shown in Fig.15. Each member is supplied with lugs for connecting other parts of the system. Alternating current passes through the leads (slide wires), running through the cores. Figure 16 shows the ordinary form of the counter-system model, mounted with 10 counter cells. The model consists of 10 members, like those shown in Fig.15, several performing relays inserted into the various outputs of the counter circuit, and a dial by means

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of which information can be obtained.

Conclusions

The merits of the devices considered include: the relatively high operating speed, negligible bulk, high transmission factor, long service life, use of simple and inexpensive semi-finished products.

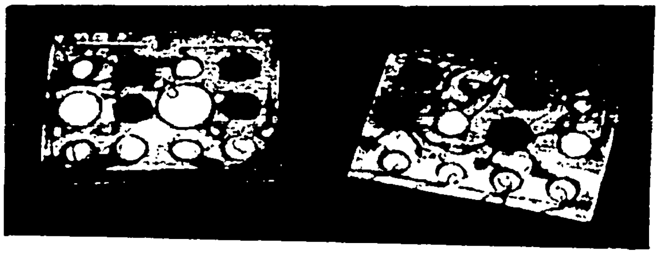

Fig.15

The disadvantages include: the need of feeding alternating current of high frequency and the relatively large energy consumption compared to the energy use in ordinary relay systems of telephoning. The power consumption may be cut down in the future by devel-

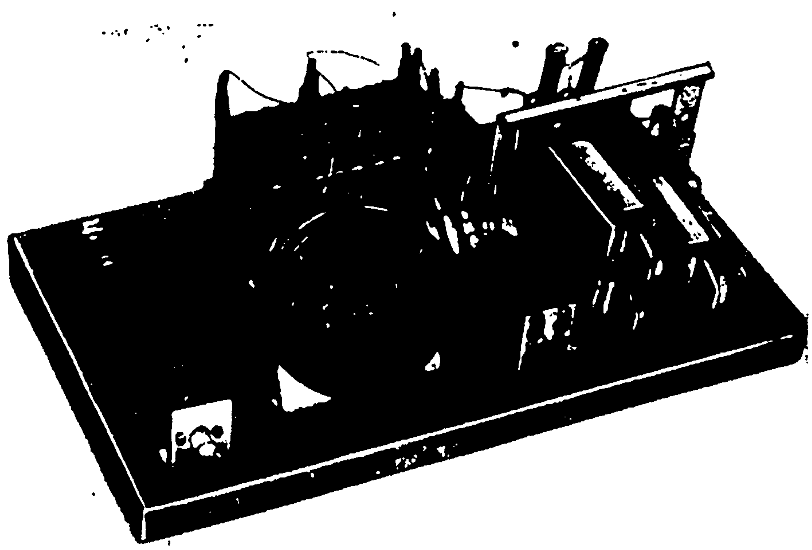

Fig.16

oping ferrites with a lower dynamic power factor.

In comparison with other methods of contactless switching, ferroresonant devices are highly responsive to frequency variations in the feeding current; however, in the systems built at the base of the doubled ferroresonant transformer circuit

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(Figs.9, 10, 11, and 12), an adequate range of frequency oscillations for practical use was attained, as well as feeding voltages of alternating current. For example, the frequency deviation from the norm was $\pm 20\%$; if the frequency is maintained within the limits of $\pm 5\%$, a variation in the feeding voltage of $\pm 15\%$ is permitted.

The quick-action response of ferroresonant systems is lower than that of tube, semiconductor, and pulse magnetic (with a rectangular hysteresis loop) devices and is determined by the frequency of the feeding current. For ATS processes and a series of other branches of automatic devices, the response attained is quite adequate. The elements of the developed systems are run by direct current and, at their outputs, still give a direct current with sufficient power to trigger a common telephone relay. These characteristics make it easy to combine the described systems with conventional ATS (toll office) relay systems and other branches of automatic apparatus which are rendered useful by gradually introducing contactless elements into existing systems.

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METHODS OF PHANTOM TELEPHONE COMMUNICATION

by

N.V.Reshetnikov

Methods of experimental determination of traffic-carrying capacity and loss for completely available and partially available bands are discussed. An appraisal of the examined methods of phantom telephone communication is given.

Introduction

It is generally known that any theorem requires experimental verification under actual conditions or under conditions near to actual. In many branches of science and technology, experimental verification of the theoretical findings presents no great difficulty. However, the theory of telephone communication does not belong into this category. The theory of telephone communication is a science in which any experimental check of the theoretically obtained results meets with numerous great difficulties which are sometimes almost insurmountable.

At first glance it may appear that the validity of several theorems, obtained by means of the theory of probability, may be easily verified at the operating telephone stations. In actuality this is not feasible for the following reasons:

1) At present, our telephone stations are not yet equipped with apparatus for adequate measurement of the magnitude of telephone load and loss. It is true that there are methods of "manually" or semiautomatically measuring telephone loads, but such measurements are crude and demand a large staff of workers. As a result, we do not know the exact values of the load entering the office, nor the value of the load being sent through.

2) The telephone load in the operating offices does not remain constant, but

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changes in quantity from day to day. Therefore, we can never reach the desired load intensity required in conducting tests.

3) Tests conducted at the operating telephone offices should be executed at a high level of maintenance, i.e., with low losses. This is a drawback in conducting experiments with increased loss, i.e., losses of 10 and more.

4) There is no full guarantee that, in the investigated trunk group, all the equipment is correct and that during the time of the investigation no damage has been inflicted on the office apparatus.

These considerations indicate the great difficulties in conducting check tests and measurements in operating offices, but of course do not mean that, in general, testing under service conditions should be completely abandoned. On the contrary, to secure a clear concept of the values of the load and loss, of the nature of the loads and of the work of the office in general, it is essential to introduce automatic measuring devices as soon as possible.

The investigation of a given system of connecting lines, with the purpose of determining the traffic capacity of the system and its loss, is possible by means of the so-called methods of phantom telephone communication.

The widest application has been made of two methods of phantom telephone communication: a) the method with constant holding time and b) the method with exponential distribution of the holding time.

Both methods are examined below.

The Method of Phantom Telephone Communication with Constant Holding Time

The principles of the method was propounded by the British scientist O'Dell in his time. The method was introduced into practice by Berkley (Bibl.1) and developed by G.L.Grigor'yev (Bibl.2).

The nature of this method of artificial load is basically as follows: The testing time, for example, extended to 1 hour, is split into 10,000 equal parts.

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Then, the duration of one of these parts consists of

$$\frac{3600}{10\,000}\ sec. = 0{,}36\ sec.$$

After that, by means of the tables of random numbers (Bibl.3) or by means of the telephone reference book, a four-value combination of random figures is prepared.

We suggest that these four-value combinations determine the ten-thousandth parts of an hour at the beginning of which the call is started. Then, arranging the finished random four-value combinations of numbers in series of their rise in quantity, we obtain a complete chart of the moments of incoming calls. But since this method presupposes a continuous duration of calls, we know the moment of completion of each call. There now remains only the "passing through" of the desired load on the investigated trunk.

To obtain the determined values of artificial load, the holding time should be assigned. This may be assumed, for example, as equal to 200 or 300 ten-thousandths of an hour: 200×0.36 sec = 72 sec or $300 \times 0.36 = 108$ sec. To determine the required number of four-value combinations, the amount of the load should be divided into holding seconds which we must obtain as holding time. For example, to get a load to 4 holding numbers per hour, at $t_{hold} = 72$ sec, it is necessary to prepare $4 \times 3600:72 = 200$ accidental four-value combinations.

To simplify the operation during passage of a load, one hour must be parceled into 100 intervals numbering from 00 to 99, and then each interval into 100 parts. Then, selecting the holding time as 2 or 3 intervals, it is possible to simplify the process of load passage.

This method may be illustrated by an example. Let the moments of incoming calls be determined by the following four-value combinations: 0068; 0093; 0255; 0317; 0499; 0516. Then the moments of completion of the holdings, at $t_{hold} = 72$ sec, will be equal to: 0268; 0293; 0455; 0517; 0699; 0716, respectively.

If the trunks investigated are completely accessible and the lines maintained

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in a strictly determined order, starting from the first, then the changes in the state of the trunk may be illustrated by Table 1.

A detailed examination of this method for investigating systems of step-by-step switching has been done in a previous paper (Bibl.2).

Table 1

Moment of time	Occurrence
0068	Holding of the 1st trunk line
0093	" " 2nd " "
0255	" " 3rd " "
0268	Freeing of the 1st trunk line
0293	" " 2nd " "
0317	Holding of the 1st trunk line
0455	Freeing of the 3rd trunk line
0499	Holding of the 2nd trunk line
0516	Holding of the 3rd trunk line
0517	Freeing of the 1st trunk line
. . . .	. . . .
. . . .	. . . .

<u>Method of Phantom Telephone Communication with Exponential Distribution of Holding Time</u>

The assumption of continuous holding time as the simplest method is made since, with different holding times, the conducting of experiments becomes extremely complicated. However, there does exist a simple enough method of phantom loading, with an exponential distribution of holding time. This method is known under the name of "Monte-Carlo" method (Bibl.4).

In making up the artificial load, it is acceptable not to know the starting and finishing instants of each holding, as was required in the method with continuous holding time. It is sufficient to know simply the sequence of all moments of beginning and finishing the holdings; with this, the holding time can be noncontinuous.

If, at a determined moment of time, the X trunk line is busy, then with exponential distribution of holding time the probability that one of the holdings in the following interval Δt is concluded does not depend on the time of the holding already elapsed. This can be demonstrated as follows: Let f_a be the probability that the holding, having lasted a sec, continues for not less than t sec, so that, at the indicated distribution, we have

$$f_0(t) = e^{-\beta t}.$$

Therefore, it apparently always follows that

$$f_0(a+t) = f_0(a) f_a(t),$$

since, at the demonstrated distribution,

$$e^{-\beta(a+t)} = e^{-\beta t} f_a(t),$$

whence

$$f_a(t) = e^{-\beta t}.$$

This means that, at the indicated distribution of holding time, the law of distribution of the remaining parts of holding does not depend on its "growth", i.e., on the elapsed time. Consequently, all lines will be in equal conditions relative to their being free.

Let the load of several trunk lines be denoted by Y. Then, at an average holding time T, the probability of new calls coming in during the time Δt is determined as

$$\frac{Y}{T}\Delta t.$$

The probability of freeing one of the held lines in the time Δt is $\frac{\Delta t}{T}$. Then the relation of the probability of new incoming calls to the probability of freeing one of the busy lines is equal to

$$\frac{Y}{T}\Delta t : \frac{\Delta t}{T} = Y:1.$$

Since all trunk lines are in equal conditions with respect to their being cleared, the relation of the probability of freeing the first, second, third, ..., X^{th} trunk line and the coming in of new calls will be equal to 1:1:1:...:1:Y. On this basis, we can compile an exact chart of the occurrences in the trunk line, determining the condition changes by sorting. With sorting, the freeing of each of the held lines has equal probability, and the probability of a new call coming in is Y times larger.

To serve as an example, we will bring in a completely accessible trunk with 7 lines on which a load Y = 3 holding numbers per hour is being carried. The change in condition of the trunk is estimated by means of a so-called "determinant" number. Let the yield number 1 correspond to the freeing of line 1, if it was busy; yield number 2 to the freeing of the second line if it was held; yield number 3 to the freeing of the third line, if it was busy; yield numbers "8, 9, or 0" to the arrival of a new call. The determinant numbers may be taken from the Tables of random figures (Bibl.3) or other sources.

If Y + V > 10, then it is necessary to set up two-value combinations of random figures. From the two-value figures, one hundred possible combinations can be set up, i.e., 100 determining numbers.

If Y is a fraction, it will be possible to start as follows: Let Y = 2.5 h.n.; V = 7. As determinant numbers, we use 00 to 18. With this, the numbers

01 and 11 designate the freeing of the 1st line, if it was busy

02 and 12 " " " " " 2nd " " " " "

03 and 13 designate the freeing of the 3rd line, if it was busy

. . . .

while the numbers 00, 08, 09, 10, 18 designate the arrival of new calls. With two-value tables of random numbers, one should proceed accordingly. If the number from the Table is less than or equal to 18, then it is considered a determinant number. If this number from the Table is more than 18, then it must be divided by 18 and the remainder ~~considered as~~ a determinant number.

Let us examine the case where Y = 3 h.n.; V = 7, and the lines of the trunk are held in the sequence of the line number increase, beginning from the first.

Given that, at first, lines 1.2 and 5 were already examined.

The testing may proceed using Table 2.

Table 2

Determinant Figures	Occurrence	Lines held
-	-	1, 2, 5
8	New call	1, 2, 3, 5
7	-	1, 2, 3, 5
6	-	1, 2, 3, 5
8	New call	1, 2, 3, 4, 5
6	-	1, 2, 3, 4, 5
9	New call	1, 2, 3, 4, 5, 6
8	New call	1, 2, 3, 4, 5, 6, 7
0	Lost call	1, 2, 3, 4, 5, 6, 7
5	Freeing of line No.5	1, 2, 3, 4, 6, 7
4	Freeing of line No.4	1, 2, 3, 6, 7
1	Freeing of line No.1	2, 3, 6, 7
3	Freeing of third line	2, 6, 7
0	New call	1, 2, 6, 7
. . .	. . . .	. . . .

In this way the duration of each holding is a random value.

In this example, 400 holdings were carried. The results of the investigations were broken down into four parts with 100 holdings in each. The number of holdings was determined, carried by each line in 4th hundreds, and the number of calls which were not satisfied by the 1st; 1st and 2nd; 1st, 2nd, and 3rd lines, etc.

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The results of these investigations are shown in Tables 3 and 4.

Table 3 gives the distribution of calls on the lines of the trunk, while Table 4 shows the number of calls, not satisfied by the 1st; 1st and 2nd lines, etc.

Table 3

a)	b)			
	1	2	3	4
1	32	17	30	19
2	22	19	23	22
3	19	17	19	24
4	11	12	14	18
5	8	12	7	11
6	5	9	4	5
7	3	5	1	1
c)	0	9	2	0

a) Nos. of lines; b) Number in hundreds; c) Lost calls

As Tables 3 and 4 show, regardless of the relatively small number of tests, the results obtained by the method of phantom loading coincide well with the theoretical data.

Table 4

a)	b)				c)	
	1	2	3	4	d)	e)
1	68	83	70	81	0,755±0,033	0,750
1 and 2	46	64	47	59	0,540±0,038	0,530
1÷3	27	47	28	35	0,343±0,040	0,345
1÷4	16	35	14	17	0,205±0,045	0,205
1÷5	8	23	7	6	0,110±0,035	0,110
1÷6	3	14	3	1	0,053±0,025	0,052
1÷7	0	9	2	0	0,028±0,019	0,022

a) Nos. of the line; b) Number in hundreds; c) Probability of loss;

d) Experimental; e) Theoretical

This same method can be applied to testing the systems of step switching. For example, if the switching system has three loading groups, then the following can be obtained from the determinant numbers:

"1" corresponds to a call in the 1st group;

"2" corresponds to a call in the 2nd group;

"3" corresponds to a call in the 3rd group;

"4" the freeing of the individual line of the first subgroup, if this had been busy;

"5" the freeing of the individual line of the 2nd subgroup, if this had been busy;

"6" the freeing of the individual line of the 3rd subgroup, if this had been busy;

"7" the freeing of one of both lines, if such was held, etc.

After selecting the determinant numbers for the change in trunk condition, the operations in testing remain just as described above for the case where this method was applied for a completely accessible trunk.

Evaluation of the Methods of Phantom Telephone Communication

The described methods for phantom loading may be used for investigating the passage capacities of various switching systems and for determining losses.

However, several essential shortcomings of the method of phantom telephone communication with constant holding time should be noted, namely:

1) The method supposes a constant holding time, which never conforms to actuality. Observation shows that the holding time is distributed according to a quasi-exponential law;

2) The method presupposes that the number of hourly calls is exactly equal to Y/t_{hold}. This too does not conform to actuality. In examining a large number of holding hours, the number of calls in two successive hours may be more than Y/t_{hold} and less than Y/t_{hold};

3) The method is extraordinarily laborious. It requires a constant recording

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of all call conditions.

The use of the method of phantom telephone communication with exponential distribution of the holding time largely frees the testing process of the mentioned defects. However, the complexity of these tests remains perceptible even with this method. Nevertheless, the exponential distribution method requires much less written work than the method with constant holding time.

A complete elimination of the mentioned defects is possible only by using a phantom loading device.

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A POLYTONIC SYSTEM OF NUMBER TRANSMISSION IN LONG-DISTANCE CHANNELS

by

V.N.Zachesov

Comments on the article by C.A.Lovell, J.H.McGuigan, O.J.Murphy; An Experimental Polytonic Signaling System. BSTJ, Vol.34, No.4 (1955)

The extensive development of long-distance automatic telephone connections in the USA raised the logical question of shortening the time of setting up a call between offices. Step by step switches, not satisfying the increased technical demands, due to structural peculiarities, were replaced by the quick-action connecting device of the crossbar type which had the same great switching potentialities. In its turn, the transition to the crossbar connector demanded a revision of the system of number transmission on long-distance channels, since the pulse-number code by which each figure is transmitted by the corresponding number of current pulses is not receivable because of the large time loss in its transmission.

A reduction in time was achieved by using coded number transmissions; the signal system was based on the use of several distinguishing signs or several meanings of one distinguishing feature. The parameter of the signal was designated as the "distinguishing sign" used for differentiating one signal from another in any signaling system.

The frequency method of encoding is in widest use, since it is relatively easy to learn and apply in the decoding of coded messages. The use of the frequency method is based on the ability of band filters to separate frequencies with the purpose of developing the frequencies present in code transmission. The methods of frequency number coding may be diverse. Together with the coding of the various number signs and separate transmission in time, there may be simultaneous coding of

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several signs (groups of signs) or of the whole number.

In the USA the subscriber number is transmitted in the channel by encoding each sign separately, by using the frequency code of "two from five" ($C_5^2 = 10$). This code permits the development of a unit error in the form of a third frequency appearing in combination, or the absence of one of the required two frequencies as a result of the superimposition of interference.

Operation of the quick-action connectors and use of number coding transmissions are made possible by interoffice equipment. The holding time of this equipment and, consequently, its use in maintaining a large number of calls is directly dependent on the time of setting up the call, and on the method of coding the subscriber number (on the time of transmitting the code).

The paper discussed here is devoted to the question of lessening the time in transmitting the number code. The use, in place of the usual frequency transmission of attenuating sinusoidal oscillations of brief duration and the use, at the transmitting and receiving ends, of circuits adjusted to the oscillation at the specified frequency shortens the time of code transmission by 25 - 30%, which permits increased use of the apparatus or, which comes to the same, a reduction in the number of appliances in the office of the same capacity.

In the Bell Telephone Laboratories, a multifrequency system of number coding was worked out and tested under laboratory conditions, which, in contrast to the ordinary multifrequency system, is called polytonic (polytonic signaling system).

The polytonic signaling system is based on the principle of orthogonality used for a specific class of mathematical functions. Numerous time functions $\Phi_1(t)$, $\Phi_2(t)$ ..., $\Phi_n(t)$, are orthogonal in the time interval T, if

$$\int_0^T \Phi_i(t)\Phi_j(t)\,dt = \begin{cases} 0 & \text{at} \quad i \neq j \\ 1 & \text{at} \quad i = j. \end{cases}$$

Sinusoidal functions may serve as an example of this sort of function.

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The property of orthogonality may be used for the separation, at the receiving end, of signals consisting of a combination of sinusoidal functions, without using ordinary band filters for this.

If coding is done by voice-frequency pulses in the quality of code elements, then the separation of signals may be realized by means of generators switched into the receiving end, whose frequencies are identical with the frequencies used in coding (each generator should be connected to its integrator). The received signal, a mixture of several frequencies, "is multiplied" by means of the corresponding circuits in oscillations of the generators "receivers", and the result is integrated into the limits of the period T. As a result either "zero" or "one" is received at the outputs of the various integrators. "Zero" voltage is generated at the integrator output if the received signal contains no frequency to which the generator has been adjusted, while a voltage "one", is produced if such frequency is present.

An analogous effect may be obtained by connecting passive electric circuits instead of generators into the receiving end. These circuits are composed of resistance, inductance, and capacitance and are adjusted each to one of the frequencies used in coding.

The integral

$$V(t)=\int_0^T \theta_i(\lambda)\,\theta_j(T-\lambda)\,d\lambda=\begin{cases}0 & \text{at } i\neq j\\ 1 & \text{at } i=j,\end{cases} \qquad (1)$$

is the orthogonal condition of a multitude of normalizing functions $\theta_1(t)$, $\theta_2(t)$ ..., $\theta_n(t)$ during the time interval T, and describes the circuit response to the effect $\theta_i(t)$ under the condition that the pulse reaction of the circuit is $\theta_i(t)$. The reaction of the passive electric circuit is due to the signal action; accordingly, it is easy to judge whether there are components in the signal to which the circuit is tuned. In this way, the passive electric circuit, composed as above, may fulfill the functions of oscillating, multiplying, and integrating.

This is mathematically proved in the following manner: If, as code elements,

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attenuating sinusoidal oscillations (Fig.1a) are used which are characterized by the expression

$$E_m = E_0 e^{-\frac{at}{T}} \sin m\pi \frac{t}{T},$$

and at the receiving end a circuit is connected whose characteristic is expressed by the formula

$$G_n(t) = K e^{-\frac{at}{T}} \sin n\pi^{\frac{t}{T}},$$

then, integrating the products of these expressions and then solving eq.(1) for t = = T, we obtain:

$$\text{for} \quad m \neq n \quad V(t) = 0,$$

$$\text{for} \quad m = n \quad V(t) = KE_0 (-1)^{n+1} \frac{T}{2} e^{-a}.$$

The voltage in the capacitor of the receiving circuit, excited by the sinusoidal attenuating wave (Fig.1a) to whose frequency the circuit is tuned, reaches its maximum at the instant of time T (Fig.1b). The oscillations in the circuit, excited by the wave to whose frequency the circuit is not tuned, will begin at zero tilt and zero amplitude, reach maximum amplitude, and return to zero tilt and zero amplitude at the instant T (Fig.1c).

The instants T, transferred to the time axes (Fig.1), appear as points at which the oscillations in all circuits, not tuned to the frequency of signal oscillation, are simultaneously equal to zero in both amplitude and tilt. If a signal arriving at the receiver input consists of several quantities of attenuating oscillations of equal frequency, then the oscillations in the circuit receiver take on a more complex form.

The presence or absence of voltage may be established by instantaneous measurement at the capacitors of the circuit, at the moment of time T. The instantaneous measurement of voltages permits determining which of the frequencies are present in the given signal, i.e., it becomes possible to decode.

The above circumstances serve as the basis for designing and manufacturing a model of this rapid-action signaling system.

However, in practice it is always difficult to note the right moment for meas-

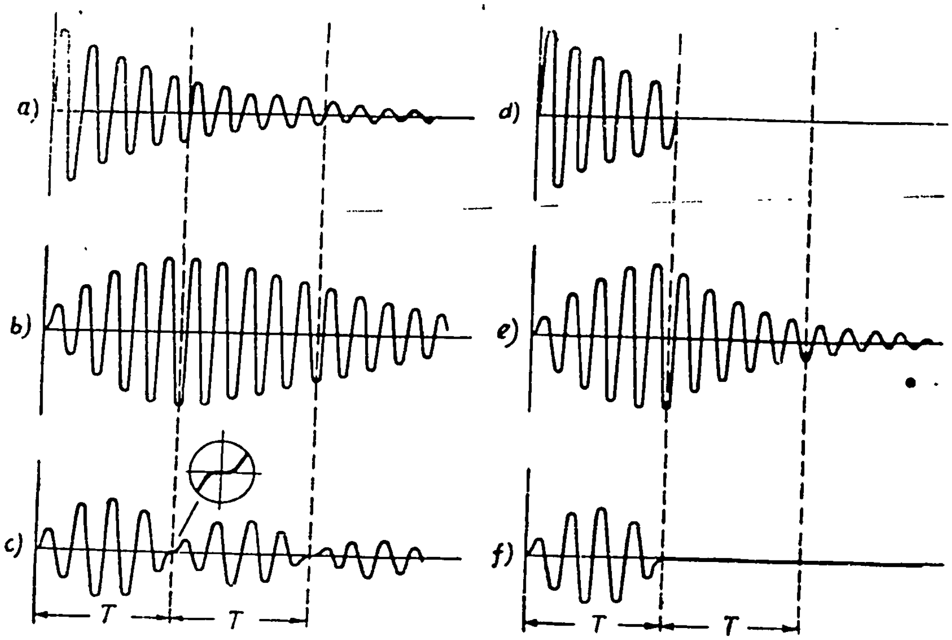

Fig.1

uring the voltage. The difficulty is due to the interruption of signal transmissions, when the transmitting device has an unknown delay time and an unlimited transmission band. It is suggested by the authors that sinusoidal attenuating oscillations of a determined duration (Fig.1d) be used in the experimental arrangement as signals; this facilitates the choice of the moment for measuring the voltage.

From an examination of the oscillations (Fig.1b, c) excited by the frequency, to which one of the circuits is tuned and the other not tuned, the following deduction may be made: The voltage in the capacitor of the first of the circuits is periodical, having throughout the time T a finite value different from zero at the time when it is equal to zero in the capacitor of the other circuit. Moreover, there is no energy stored in the coils of either circuit at the same moment. If at the end of the period T, the signal is "intercepted", then, the oscillations in the tuned circuit will continue (Fig.1e), while they will stop in the untuned circuit thereafter remaining equal to zero (Fig.1f). Thus, the presence or absence of voltage in the capacitors of the circuits may be fixed without rigidity at the moment T, and some while thereafter, during which time the measuring may proceed without being instantaneous. The time reading at the moment of measuring (from 0 to T + Δt) proceeds with relative ease, since the emissions have a sufficiently steep wave front from the beginning determining the starting time. The effect of "intercepting" the signal is obtained by disrupting the attenuating oscillations of the generators from the transmitting circuits.

In the polytonic system of signaling, emissions are received for a duration of 5 msec, with pauses of 5 msec. For coding every number, combinations of two frequencies out of five are used. The code "two out of five" permits formation of the required number of combinations for sending all numbers and is easy to handle. This arrangement obviously is used in the system because of the fact that the further automatic connections, employed in the USA, use the very same code. The number signs are transmitted in time sequence. For coding it is recommended that frequencies of 500, 700, 900, 1100, and 1300 kc be used. When selecting a transmission period of 1 sec, it is possible to transmit 100 decimal numbers.

Figure 2 gives the principal diagram of the test transmitter. The sending of signals proceeds as follows: After the number is fixed in the recorder, consisting of recording devices of various number signs (P_1, P_2, P_3 ...), the multivibrator switches on. Each recorder must have one input for every ten outputs. Each recorder output includes two oscillating circuits, tuned to the frequency corresponding to combinations of one determinant number. The multivibrator emits square pulses (duration, 5 msec) for successive shifting of the distributor from one position to the next. In the next circuit, these pulses are shortened by the differentiating circuit. The electronic distributor forms the circuit from the differentiating circuit to the recorder. The voltage pulse, generated by the differentiating circuit when the square pulses of the multivibrator begin to enter, passing into the resonance

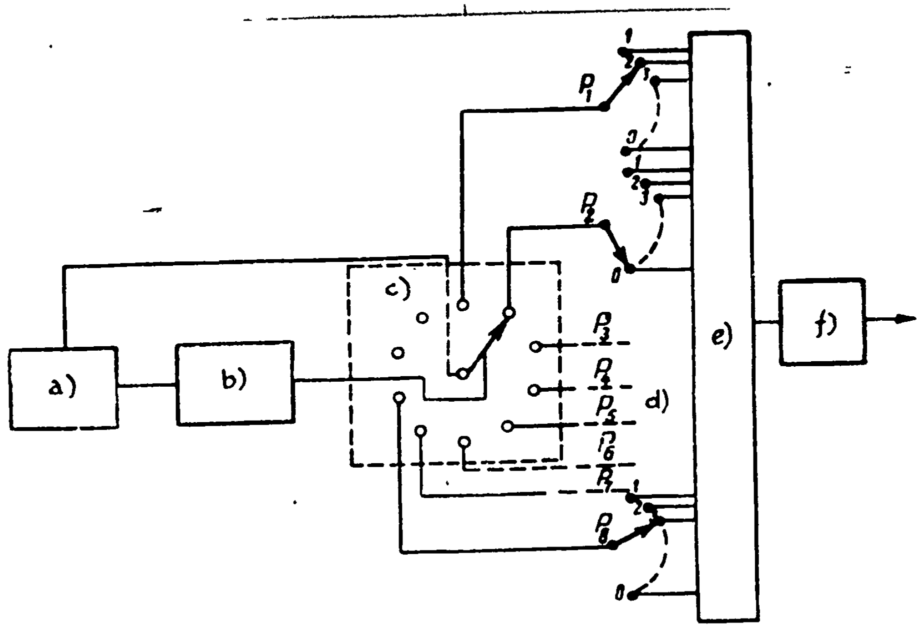

Fig.2

a) Multivibrator; b) Differentiating circuit; c) Distributor; d) Recorder; e) Resonance circuits; f) Output circuit

circuits, excite oscillations there. These oscillations are disrupted by voltage pulses of the differentiating circuit, generated by the latter at the termination of the multivibrator pulsing. Simultaneously, the distributor switches to the following recorder and once again oscillations are excited in the circuits, corresponding to the figure established in the recorder. Oscillations in the circuits are fed to a common resistor, from where they enter the line through the output circuit (BK).

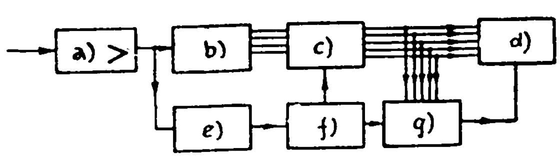

Fig.3

a) Amplifier; b) Receiving circuit; c) Detector; d) Register; e) Starting circuit; f) Time circuit; g) Marker

In Fig.3 a block diagram of the receiver of polytonic signals is shown. The experimental receiver model has several elements, not shown in the diagram and necessary only for carrying out the laboratory experiments.

In the article referred to, where the circuits of the transmitter and receiver are described in sufficient detail, the difficulties in designing the system are discussed. The problem of sending 300 decimal signs per second is mentioned. The shortening of the sending time requires a lengthening of the interval between frequencies used for coding. The transmission of 300 signals per second can be realized in channels with wider pass bands. This rate of number transmission, apparently, can be used between long-distance telephone offices.

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CALCULATION OF THE OUTPUT CAPACITY OF TELEGRAPH SERVICE EQUIPMENT

by

P.V.Prakhov

The basic factors are determined which influence the volume of producing power of telegraph equipment. Existing calculation methods of a number of telegraph sets are critically evaluated, new formulas are presented for calculating a number of devices in telegraph service with manual and automatic telegram reception, depending on the volume and character of the irregularity and variability of the load at the connections, the traffic capacity of telegraph sets, and the length of the service periods in transmitting telegrams.

General Principles

Calculation of the output capacity of the apparatus in telegraph systems is necessary in the following cases: a) in continuous usage (especially in revising the routing systems for telegrams); b) in intraproduction planning of telegraph enterprises and other communications facilities where telegraph is an independent department or service; c) in designing telegraph connections for a specified prospective period. In all these cases, there is a need for verifying the usefulness of existing equipment for the incoming load, at given quality factors for the telegram transmission period, and for determining the requirements in new equipment.

The equipment output capacity in an individual telegraph link is determined by the product of the number S of instruments operating in the link (or of multiples - for multiplex instruments) with the traffic capacity of a single instrument (or multiple) N'_e T-mm/hr, i.e., by the magnitude

$$P = SN'_e. \quad (1)$$

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The calculation of the production capacity of the equipment in the link should consider: the load volume in the link, the character and extent of irregularity and variability of the load, the norms of traffic capacity of the instruments, and the magnitude of the service periods in sending telegrams.

The formula, now used for calculating the number of devices in the link, considers the irregularity of the load according to hours of the day, days of the week, and months of the year (Bibl.1). But this formula does not consider accidental load variations within the limits of an hour or smaller divisions of time and does not give the direct relation of the productive capacity of telegraph equipment to the magnitude of service time in transmitting telegrams; this is its basic shortcoming. Meanwhile, as observations indicate, the load variability exerts a harmful influence on the ordinary magnitude of the equipment output capacity in telegraph links, not to mention irregularity.

Figure 1 gives a concept of the character of irregularity and variability of

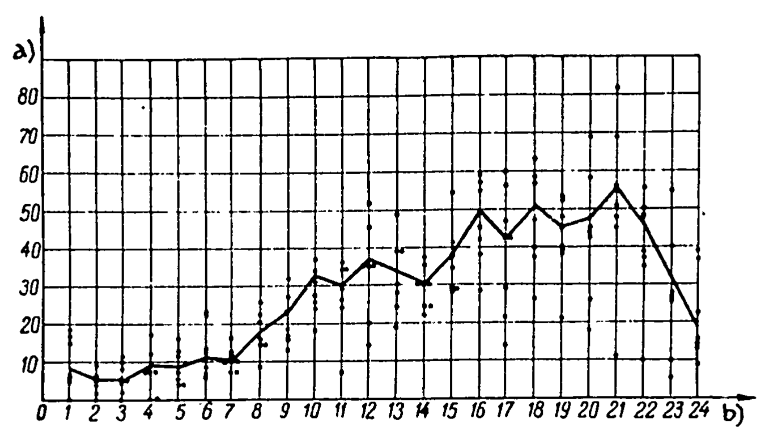

Fig.1

a) Load in telegrams; b) Hour of day

load, according to the individual hours of the day. The solid line in Fig.1 shows the change in the average load for each hour in the course of a day; the dots around the solid line show load variation or scattering.

The productive capacity of the telegraph equipment in the link should guarantee the transmission of telegrams during an earlier fixed service period, regardless of load irregularity and variability. As it is known, on one hand, the service period represents the guaranteed period of telegram transmission whose breakdown as a qualitative norm factor is inadmissible; on the other hand, the service period represents some division of time during the course of which one can expect telegrams to be sent; this allows a leveling out of the load during this period of time. The shorter the service period and the greater the load fluctuations involved, the greater should be the reserves of productive capacity of the equipment, in order to ensure telegram transmission without breakdown in the service period. In this way, the service period of telegram transmission represents one of the important factors determining the magnitude of output capacity of the link equipment.

Ye.N.Bukhman (Bibl.2) raised the question for the first time of calculating the productive capacity of equipment in telegraph links from its dependence on the magnitude of the service period. The calculation method suggested by him derives from a combined examination of hourly load during the course of days and assumes that there is only one apparatus in the circuit. The method is characterized by extreme complexity of calculation and, by its nature, excludes the use of nomographs, since it is derived from the individual graph of hourly load in the link. The calculation method recommended by Ye.N.Bukhman offers only some theoretical interest.

The method developed by P.P.Fayngluz and M.A.Vlasov (Bibl.3) of calculating the switchboard positions in telegraph systems as a function of the service period is due to an obvious misunderstanding and is erroneous for telegraph link conditions.

The method of computing the output capacity of equipment in telegraph circuits is determined by various means, depending on the method of telegram pickup (manual, automatic - with or without tearing of the tape) and depending on whether the circuit is coupled to a capacitor. Below, the method of calculating the productive capacity of telegraph equipment will be considered for manual and automatic recep-

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tion with tape tearing, for the case when the circuit is not connected to a capacitor.

Method of Calculating the Output Capacity of Telegraph Equipment with a Continuous Inflow of Telegrams

The method of calculating the output capacity of equipment in telegraph links does not only vary with the amount of service time but also as a function of this, either supplying or not supplying an uninterrupted flow of telegrams from the active telegram reception points straight to the transmission apparatus. In the practice of using telegraph equipment, an even flow of telegrams within the telegraph network is guaranteed by the use of tape conveyers and by the distribution of switchboard positions in the processing of telegrams. However, frequently tape conveyers are used only for moving the telegrams from the receiving station to the central sorting section; from this central sorting to the switchboard position of the sets, the telegrams are distributed by hand. Meanwhile, as the test transmission from Moscow TsT has shown, it is absolutely possible to supply a steady flow of telegrams from the central sorting to the auxiliary sorting points. From these sorting points to the

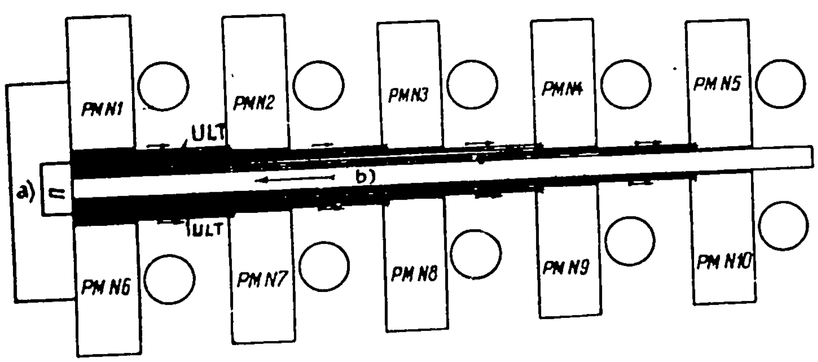

Fig.2

a) Sorting-machine operator's position; b) Belt conveyer

operator's position at the sending apparatus, a steady flow can possibly be organized if an allocation of work stations is applied, as shown in Fig.2.

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(Here PM N2 is the work station of the telegraph operator No.2; ULT is the narrow tape conveyer; P is the telegram receiver).

With an uninterrupted flow of telegrams from the work stations of telegram reception to the sending apparatus, the method of calculating the output capacity of the equipment is as follows:

At a given telegraph link, telegrams from all other links and telegram reception switchboards (banks, telephone) arrive. With an uninterrupted flow of telegrams, at any instant of time, one telegram may come through the link or two, three, and more telegrams may arrive simultaneously. If the transmittal of each telegram immediately after its arrival at the link is a prerequisite, as many sending sets as the maximum number of telegrams that can enter during the sending time of one telegram are required. It is obvious that a large number of sets would be needed in this case, a requirement difficult to meet.

Therefore, the number of sets must be limited so that delay in transmitting the telegrams is unavoidable. When the number of telegrams entering the link at the same time is larger than the number of sending sets, a determined number of telegrams (equal to the number of free sets) will be passed on, and the remaining telegrams will await transmission. It is obvious that the maximum permissible waiting time must not be larger than the station (intrinsic) control period, established for transmitting telegrams. Under these circumstances, the optimum capacity of the equipment in the link must be defined, so that the necessary quality of processed telegrams can be supplied during the period and so that no excess reserve above the actual required minimum need be maintained. The solution of this problem, in the final analysis, consists in a determination of the probability of some allowable waiting time, during which a breakdown in the given service period for telegram transmission is excluded. Hence, the permissible waiting time of telegrams τ and the station service period for sending telegrams t_{ks} are correlated by the formula

$$\tau = t_{ks} - \overline{t}. \qquad (2)$$

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Here, $\bar{t}$ is the instantaneous sending time of the telegram in the equipment.

Equation (2) results from the conditions of telegraph links where the time of instantaneous transmission of the telegram on the equipment is included in the service period.

Derivation of the formula for the waiting probability, beyond the above period P (< τ), involves a large amount of work. The most general form of this equation, as derived for telephone link conditions, was given by Ye.N.Bukhman (Bibl.2). Our investigations showed that this formula (under specific conditions) may also be used for telegraph links.

The equation has the form

$$P(>\tau) = P(>0)\,e^{-(S-y)\frac{\tau}{\theta}}. \qquad (3)$$

Applicable to telegraph link conditions, the values introduced into eq.(3) will be denoted as: S, the wanted number of sending devices in the circuit; y, the load in the given circuit while sending (in hours held); τ, the permissible waiting time for telegrams on the link before sending; θ, the mean waiting time for completion of a started transmission of a telegram; while

$$P(>0) = \frac{\frac{y^S}{S!}\frac{S}{S-y}}{\sum\limits_{j=0}^{j=S-1}\frac{y^j}{j!} - \frac{y^S}{S!}\frac{S}{S-y}} \qquad (4)$$

is the probability of holding all equipment in the given circuit at the instant the telegram arrives; in equal measure, P (> 0) represents the probability of expectation for a telegram irrespective of any period. The load y in the circuit is defined in holding hours according to the formula

$$y = Q_{my} k_{busy} N_m H_d \bar{t} = C_{dis}\bar{t} = \frac{C_{dis}}{N'_e}. \qquad (5)$$

Here Q_{my} is the mean daily (for a year) load in the link while sending; k_{busy} is the coefficient of busy hour concentration; H_m is the coefficient of seasonal irregularity of load in the circuit; H_d is the coefficient of daily irregularity of the load; $N_e^!$ is the standard used for the traffic capacity of the apparatus (telegrams per hour); $\bar{t}$ is the mean sending time for a single telegram (in parts of an hour). If, within one hour, $N_e^!$ telegrams are transmitted on the apparatus, then one telegram is sent on the average of $\bar{t} = \frac{1}{N_e^!}$ hours. In determining $C_{dis} = Q_{my}k_{busy}H_mH_d$, the mean daily telegram load in the link Q_{my} is present only in sending. The number of sets in the circuit, determined by this, must be doubled in the final computation (considering the exchange process in the reception). The mean expectancy time for the completion of telegram transmission is determined by the formula

$$\theta = \frac{\bar{t}(1+V_m^2)}{2}. \qquad (6)$$

Here V_m is the coefficient of variation, $\bar{t}$, i.e., the mean time for transmitting a single telegram. The time variation factor for the sending of a telegram depends on the variation in telegram length according to the number of words and the variation in the number of signs in the words of the telegram. Under these conditions, the common time variation factor for the transmission of a telegram is contained in the formula

$$V_m^2 = V_{dm}^2 + V_{dc}^2. \qquad (7)$$

Here V_m is the time variation factor in sending the telegram;

V_{dm} is the length variation factor of the telegram, depending on the number of words;

V_{dc} is the length variation factor of a word, depending on the number of letters.

The length variation factor of a word, according to the number of signs V_{dc}, has a stable value and is equal to 0.4. The length variation factor for the tele-

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gram, according to the word number V_{dm}, fluctuates in various links (chiefly as a function of the structure of the telegraph exchange, according to the telegram categories) within the limits of 0.3 - 0.7 and may be taken in practice to be equal to 0.5. Then the quantity θ may be assumed as equal to $0.7\bar{t}$.

The time τ of the maximum permissible expectancy of telegram transmission in the link is determined by eq.(2). With this, the station service period in the transmission of the telegram t_{ks} should be determined as a mean suspended value in considering the dimensions of the service periods and the structure of the telegraph exchange according to the various categories of telegrams.

To determine the norm N_e' of the traffic capacity used, the following formula may ordinarily be recommended

$$N_e' = N_m'\eta = \frac{n60}{am}\eta \, \frac{\text{telegrams}}{\text{hr}}. \tag{8}$$

Here N_m' is the technical norm of traffic capacity of the apparatus; n is the number of revolutions per minute of the distributor brushes of the apparatus; a is the mean number of signs in the words of the telegram; m is the average telegram length in words; η is the utilization coefficient of the technical norm ($\eta < 1$) of the apparatus traffic capacity.

Despite the apparent simplicity, the calculation of N_e' by means of eq.(8) is very complex in practice, since it is difficult to determine the coefficient η. Therefore, back in 1951 the author published a new formula (Bibl.4) which considerably facilitated determination of N_e', as a function of the various operating conditions in the telegraph link. In part, the equation for determining N_e' in start-stop systems has the form

$$N_e' = \frac{K\,[60 - (t_{pz} + t_{kn})]}{am + Ka} \, \frac{\text{telegrams}}{\text{hr}}. \tag{9}$$

Here K is the working speed of the telegraph operators (clicks of the tapper per minute); a is the mean number of signs in a word in a telegram (assumed to be 8.5 for ST-35 apparatus and 9.0 for manual sets); m is the mean word length of a telegram, including the words used in the head (as an estimate, one can assume this as equal to 25); α is the time of auxiliary work ($\frac{\text{min}}{\text{telegram}}$) spent on one telegram; $t_{pz} + t_{np}$ is the time (in minutes) for preparing and completing the work and for normal breaks deducted from the work hour (assuming this to be 2 - 3 min); the time α depends on the operating conditions in the apparatus and varies from 0.12 to 0.8, depending on whether work arrives at the apparatus operated by an assistant operator, or whether it arrives by automatic visual control or by addition of a control tape to the text of the telegram to be sent.

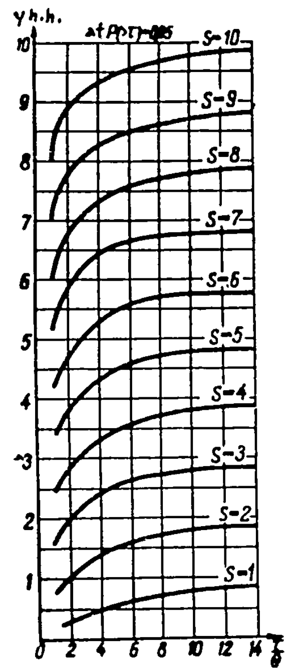

Fig.3

If the norm used for the traffic capacity of the set is determined in words per hour, then the coefficient m will appear in the numerator of eq.(9), denoting the mean telegram length according to words.

For automatic telegram reception, with both broken and transferred tapes and with automatic direct transmission of telegrams, even if some operations of an auxiliary nature are still done by hand (inserting tape into the transmitter, tapping out acknowledgement, inquiries, information, etc.), the norm used for the apparatus traffic capacity may be determined by the following formula, analogous to eq.(9),

$$N'_e = \frac{n\,[60 - (t_{pz} + t_{np})]}{am + n\alpha} \ \frac{\text{telegrams}}{\text{hr}} \qquad (10)$$

Here the quantity K, introduced in eq.(9), is substituted by n which denotes the rpm of the distributor brushes of the apparatus. The remaining denotations are same as in eq.(9).

According to the chronometric findings in the Leningrad TsT, the time α of sup-

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plementary work, taken from the reckoning in one telegram with automatic transmission, consists of 0.2 min. In automatic transmission, it is necessary to send the signs at the end of a telegram and to separate the individual telegrams, which increases the length of the telegram by about one word. In this use of the norm for teletype apparatus with automatic pickup, about 74 telegrams are averaged per hour. This is about 32% higher than the officially assumed norm used for the traffic capacity of start-stop systems at 56 telegrams per hour (1400 words per hour) with manual operation.

A direct calculation of the number of devices in the link from eq.(3) is always difficult, so that nomograms should be used. One of these charts is shown in Fig.3. In this nomogram, the abscissa gives the ratio $\frac{\tau}{\theta}$ and the ordinate, the load y in holding hours. The individual curves of the nomogram show the necessary number of sets with a different ratio $\frac{\tau}{\theta}$ and a load y.

Field of Application of Equation (3) for Telegraph Links

If both sides of eq.(3) are expressed in logarithms, we obtain the following expression after a slight transformation,

$$S = y + \frac{\theta}{\tau} \ln \frac{P(>0)}{P(>\tau)} = \frac{Q_{my} k_{busy} H_d H_m}{N'_e} + \frac{\theta}{(t_{ks} - \bar{t})} \ln \frac{P(>0)}{P(>\tau)}. \qquad (11)$$

The expression obtained for S has an implicit form: The value P (> 0) is a function of S and y. However, it clearly shows that the number of sets S is numerically equal to y only at a very large t_{ks}, when using eq.(3). In fact, the hourly change in load in the course of a day shows that, at a rise in t_{ks}, the number of sets S may be smaller than y in busy hours at the expense of a possible leveling of the load.

It is of interest to define the value at which the service period eq.(3) no longer satisfies the primary conditions under which it was derived and to determine whether it must be substituted by another formula.

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The most important premise in the derivation of eq.(3) is the condition that the distribution of the number of simultaneously incoming telegrams in the circuit is a Poisson distribution. In the Poisson distribution, the mean value of $\bar{x}$, the mean-square deviation σ, and the coefficient of variation $V = \frac{\sigma}{\bar{x}}$ are interconnected by the following correlations:

$$\sigma = \sqrt{\bar{x}} \text{ and } V = \frac{1}{\sqrt{\bar{x}}}.$$

The verification of these conditions in circuits with a continuous inflow of telegrams shows that the Poisson distribution of inflowing telegrams in the circuit takes place during time intervals not exceeding 10 - 12 min. At greater intervals the Poisson distribution no longer takes place, and the distribution of the simultaneous inflow of telegrams in the link acquires a normal character. In circuits in which there is no continuous inflow of telegrams, the Poisson distribution of inflowing telegrams in the circuit is generally absent. Thus, eq.(3) may be used for calculating the equipment output capacity only in the circuits of those telegraph networks where an uninterrupted inflow of telegrams into the circuit is supplied (for example, when using belt conveyers), in view of the fact that the allowed expectancy time τ should not be greater than 10 - 12 min. The use of eq.(3) for τ > 10 - 12 min may require the installation of excess equipment as follows from eq.(11).

Calculation Methods for Equipment Output Capacity in Telegraph Links with Discontinuous Inflow of Telegrams

In this Chapter, we will discuss the graph of hourly telegram inflow into the link (Fig.4) in which the mean number of telegrams, arriving in the course of one hour, is plotted. Let us assume that, at a sufficiently large permissible expectation time τ of sending telegrams, the productive capacity $P = SN_e'$ of the equipment takes up less load C of telegrams at the busy hour. Then, during peak periods some accumulation of arriving (but not of transmitted) telegrams will occur. These ac-

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cumulating telegrams must be sent in the time τ when the load drops and becomes less than the output capacity of the equipment, established in the circuit.

If the hourly load (in telegrams) is equal to C, and the total traffic capacity S of the sets is SN'_e, where N'_e denotes the traffic capacity of one set (or multiple) in telegrams per hour, then the accumulation R, during the course of one hour, will consist of

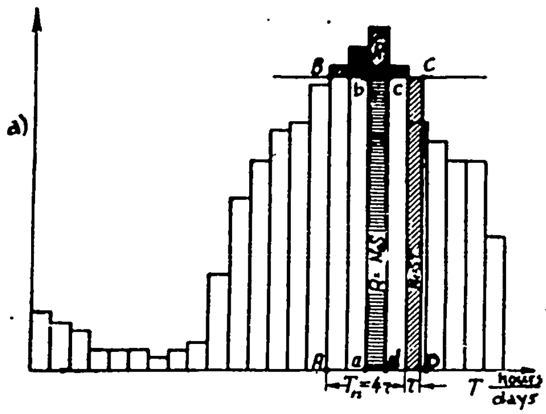

Fig.4

1) C = load in t_{max}; 2) T $\frac{\text{hrs.}}{\text{days}}$;

$$C = SN'_e = R.$$

If the load exceeds the traffic capacity of the equipment in the course of T_n hours, then the total accumulation for this time is equal to

$$\Sigma R = \Sigma C - T_e SN'_e.$$

To prevent a breakdown of the service period, the following equation must be satisfied:

$$\Sigma C - T_n SN'_e = SN'_e \tau,$$

whence

$$S = \frac{\Sigma C}{N'_e (T_n + \tau)}. \qquad (12)$$

In eq.(12), the load C taken for each hour, is the average load. The accidental fluctuations in load (both for each hour and within the limits of one hour) have been disregarded in this formula. As observations have shown, these accidental load accumulations from the intermediate value have a normal distribution. For considering the accidental load variations and the corresponding increase in productive capacity of the equipment in the load fluctuation, it is necessary to increase the mean load so that it can be covered by the computation

$$C_1 = C(1 + x_0 V_\tau).$$

Here x_0 is the coefficient, determining the probability P_{x_0}, so that the factual load exceeds the calculated value. Since the calculation assumes that the important accumulation from the mean load is only on the large side, the probability P_{x_0} is determined from the expression

$$P_{x_0} = \frac{1 - \Phi(x_0)}{2},$$

where $\Phi(x_0)$, in turn, is determined in accordance with the integral of probabilities

$$\Phi(x_0) = \frac{2}{\sqrt{2\pi}} \int_0^{x_0} e^{-\frac{x^2}{2}} dx.$$

Further, $V\tau$ is the coefficient of load variation in the time interval of τ.

The need, when calculating the load, to proceed from a consideration of the coefficient $V\tau$ is explained as follows: It must be assumed that the hours when the load exceeds the productive capacity of the equipment, consist of separate divisions each of which is equal to τ in duration. In each space of τ there is an accumulation of telegrams awaiting transmission. The telegrams piling up in the first space τ are sent during the first portion of the second time gap, etc. This process of load shifting in the intervals of time τ leads to the accumulation of such a load at the end of the period that it must be processed when the equipment output capacity $P = SN_e'$ for the time τ is present.

In Fig.5 this process of load displacement during time intervals is shown for the example of one hour. In a like manner, this can be stretched out for any number of hours.

Under load displacement condition for the time interval τ it is important to consider the load fluctuations for periods of a duration τ. In the light of this,

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eq.(12) takes on the form

$$S = \frac{\Sigma C\,(1 + x_0 V_\tau)}{N'_e\,(T_n + \tau)}. \qquad (13)$$

In eq.(13) the value $\Sigma C - C_1 + C_2 + \dots + C_{m_n}$ represents the total load for T_n hours, i.e., for the period when the load rises above the output potential of the circuit equipment.

A direct calculation of the number of sets S, according to eq.(13) is difficult, since there still remains the indeterminate value T_n. To convert eq.(13) into an easily usable form, we transform it. We figure the hourly load for the period of the peak load from the daily load Q and the factors of concentration of the corresponding hours. Then,

$$S = \frac{(C_1 + C_2 + \dots + C_{m_n})(1 + x_0 V_\tau)}{N'_c\,(T_n + \tau)} = \frac{Q\,(1 + x_0 V_\tau)\left(\frac{C_1}{Q} + \frac{C_2}{Q} + \dots + \frac{C_{m_n}}{Q}\right)}{N'_e\,(T_n + \tau)}.$$

Since the values $\frac{C_1}{Q} = k_1$, $\frac{C_2}{Q} = k_2$, and so on represent the coefficients of concentration of the corresponding hours of the day, we have

$$S = \frac{Q\,(1 + x_0 V_\tau)(k_1 + k_2 + \dots + k_{m_n})}{N'_e\,(T_n + \tau)} \qquad (14')$$

or

$$S = \frac{Q\,(1 + x_0 V_\tau)}{N'_e}\,k. \qquad (14)$$

Here

$$k = \frac{(k_1 + k_2 + \dots + k_{m_n})}{(T_n + \tau)} \qquad (15)$$

represents the mean hour coefficient of load concentration for the period of time $(T_n + \tau)$ hours.

As indicated in Fig.4, the total load for T_n hours ($\Sigma C = C_1 + C_2 + \dots + C_{m_n}$) is changed by the equidimensional load, corresponding to the area ABCD in Fig.4 which is equal to $P(T_n + \tau) = SN_e^{\prime}(T_n + \tau)$. For this one hour $P = SN_e^{\prime}$ (corresponding to the area abcd in Fig.4) and equaling Qk without consideration of load fluctuation and $Qk(1 + x_oV\tau)$ with consideration of load fluctuation.

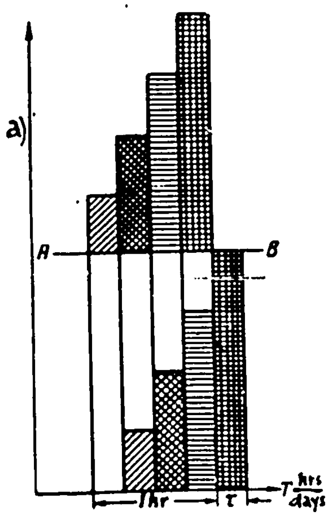

Fig.5

a) C = load in telegrams;

b) $T \frac{\text{hrs}}{\text{days}}$; c) Hour

Thus with a specified number of sets in the link and using eq.(14) as basis, it is necessary to consider all hours (including the busy hour and its adjacent hours) whose load concentration factors rise above the mean concentration coefficient k for $(T_n + \tau)$ hours. This problem is solved by the method of successive approximation. To solve the question it is necessary to have a graph of the hour load in the circuit, with computed concentration coefficients, at least for the hours of the peak load period.

In figuring out the number of sets according to eq.(14) and in determining k, the following should be remembered: In this formula τ has an "hour" irregularity and, as before, is connected with the service period of eq.(2). The station service period is determined as a mean suspended value from the service periods of various categories of telegrams.

The daily load Q in eq.(14) is determined from the correlation

$$Q = Q_{my} H_m, \qquad (16)$$

where Q_{my} is the mean daily load in the circuit for a year; H_m is the coefficient of seasonal (monthly) load fluctuation.

In eq.(16) H_d, the coefficient of load fluctuation for the days of the week, is absent since the load fluctuations for the days of the week are considered as the coefficient of load variation $V\tau$. The coefficient of load variation $V\tau$ is determined on the basis of observation of load variations in the circuit for various spaces of time τ. It depends on the magnitude of the space τ and on the value of the load in the link. There will be practically no errors in calculating if one assumes the coefficient of variation $V\tau$ to be dependent only on τ. In Fig.6, a graph is given for $V\tau$ as a function of the time of observation τ, plotted from observation data in a number of connections. Using this graph it is possible to perform a calculation of the needed number of telegraph sets in the circuit, for various service periods.

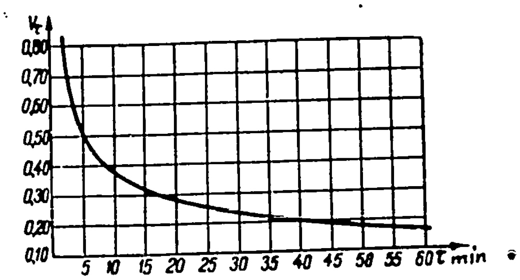

Fig.6

The coefficient x_o determines the probability of the actual load exceeding the calculated value. Its values for various probabilities are listed in Table 1.

Table 1

Values x_o	Probability P_{x_o}
1.28	0.10
1.64	0.05
1.98	0.03
2.33	0.01

In calculating, the values $x_o = 1.64$ can be used, which matches the probability $P_{x_o} = 0.05$. This means, practically, that five telegrams out of 100 during the peak load period of the most heavily loaded day of the month with the greatest load may be expected to have a service period exceeding the given one. Since the load for the other periods of these days and for the other days of the year are assumed smaller than calculated, the rest of the time will not show a lag in telegrams in the link, compared to the established service period, because of the shortcoming of the equipment output capacity.

At a small τ (not rising above 10 min), eq.(13) and eq.(14) which conforms to it can be replaced by the following simpler formula for figuring out the number of sets S on the line

$$S = \frac{C(1 + x_0 V_\tau)}{N_e'(1 + \tau)} . \tag{17}$$

Here C is the load of telegrams in the circuit at the busy hour, determined in turn by the expression $C = Q_{my} k_{busy} H_m$. The remaining designations are the same as in eq.(13). If τ exceeds 10 - 12 min, then for both continuous and discontinuous inflowing telegrams in the circuit it is necessary to use a more general form of eq.(14) for calculating S.

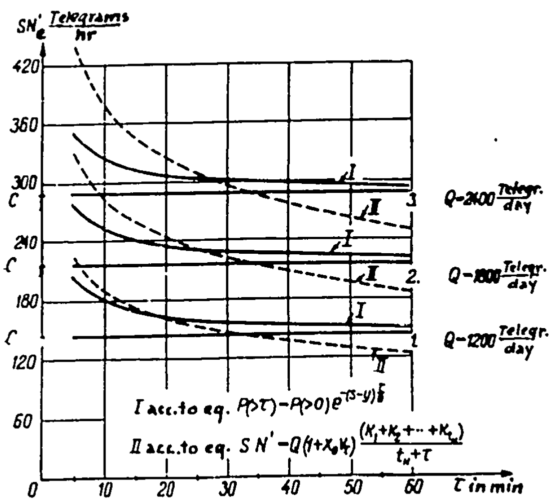

Fig.7

In Fig.7 the dependence of the output capacity $P = SN_e'$ of the equipment in the circuit on τ is plotted for a different value of the mean daily exchange Q of telegrams in twenty-four hours. As can be seen from the graph, eq.(3) yields an increased value $P = SN_e'$ when τ is higher than 20 - 25 min. Calculation according to eq.(14) indicates that, at $\tau > 30$ min, the output potential of the equipment in the link may be less than the load in the busy hour. When $\tau < 20 - 25$ min, the required value P determined by eq.(14) is larger than that obtained from eq.(3). This is quite understandable. Under conditions of discontinuous inflow of telegrams along the link and a small τ, the accumulation of telegrams produced in batches of 5 to 10 telegrams demands a boost in the reserve output capacity of the equipment.

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Article received by the Editors, 10 March 1956.

A RADIO-RELAY SYSTEM FOR TELEVISION TRANSMISSION

To interconnect television transmitters in Cologne, Frankfurt, and Neustadt, ultrashort-wave relay stations were used. The fundamental principles of constructing this relay equipment are discussed below with a short description of its features.

Design Principles of the Apparatus

The operating frequency was selected in the 1700 - 2300 mc band, which corresponds to the frequency distribution, approved by the International Conference on Radio Communication in Atlantic City.

The transmission band of each channel was set to 30 mc and the frequency separation between channels to 60 mc. This allowed 10 channels or five pairs of channels to be made in the indicated band range. The operating frequency range was dictated by the presence of tubes of "SVCh" of the 2C39 type.

In choosing the channel bandwidth, the video-frequency band extended from zero to 6 mc, and broadband frequency modulation was used to receive the required signal-to-noise ratio in the transmission of the accompanying sound signals.

For sending in only one direction, two high-frequency channels were required so that the frequencies of the relay station receivers and transmitters could be selected with discrimination; in this way, feedback through the antennas could be eliminated. If several routes intersect or run side by side, auxiliary channels are required. The slight frequency separations between channels permit transmission of a large number of channels within the given range.

In systems with many channels there is a possibility of mutual interference, due to the presence of heterodynes and insufficient image frequency suppression in the receiver. This interference may be eliminated by selecting the heterodyne fre-

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quency in such a manner that the image channel coincides with the space between the operating channels. On the basis of these requisites, the minimum separation between channels should be twice as large as the channel transmission band, if the heterodyne frequency is distributed along the fringe of the channel. In our case, the frequency separation in frequency between channels should be equal to 60 mc.

The intermediate frequency level is selected as high as possible above low ($5 \times 15 = 75$ mc), considering the available broad-band amplification tubes. In this case, the heterodyne frequency of the high-frequency channel coincides with the fringe of the adjacent channel. In this manner, at relay stations there is always the danger that an interference signal, emitted by the heterodyne transmitter reaches into the receiver of the adjacent channel.

This danger is completely eliminated by the proper choice of heterodyne frequencies.

In two channels, the heterodyne frequency of the channel with the highest frequency is taken as 75 mc above the operating frequency, and the heterodyne frequency for the lower-frequency channel, as 75 mc lower. Consequently the separation between heterodyne frequencies is assumed equal to two intermediary frequencies plus the frequency separation between channels (210 mc).

The required signal-to-noise ratio for the video signal, after 20 transmitted sections, is 35 db (ratio of maximal values) or 40 db (ratio of effective values). Using well-known correlations between the signal-to-noise ratio, the ratio of signal to potential, to sensitivity of the receiver, to amplification of the antenna, to weakening of the signal in the section, and to the number of relay stations, the following values may be obtained which determine the construction of the apparatus: transmitter potential 5 v, noise factor of the receiver 25 (14 db), frequency drift corresponding to the total amplitude of the modulating signal 12 mc, band frequency 6 mc, and antenna amplification 33 db.

The additional requirements relate to distortion in transmitting. The linearity

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of the modulator characteristics, of the demodulators, and of the video amplifiers should be such that the changes in the critical characteristics do not exceed 10% of the mean value in all video-frequency ranges.

The irregularity of group delay time, which leads to transfer distortions (from black to white) in the image must not be above 0.1 sec after 20 relays. These time lag distortions are usually introduced by intermediate and high-frequency circuits, but also by leads if there is any reflection from antennas and from devices connecting the apparatus with the feeders. In the described system, the irregularity of the group delay time characteristics in the circuits with intermediate and high frequency does not exceed 0.005 μ sec, and the reflection factor of the input terminals of the antenna is lower than 2.5%.

Structural Arrangement of the Apparatus

The equipment of the terminal station is mounted in three bays. In the upper portions of the bays, the switching panels are distributed with indicators of the signal level and of feeding voltages. The antenna switch is placed under the upper part of the housing of the receiving bay. This is remote-controlled and used for changing transmission direction. The indicator lamps show the direction in which the transmission is beamed. The base of each of the three bays is equipped with a fan.

In the right bay a UHF transmitter, a powerful IF amplifier, and a suitable power supply are mounted. The left bay contains: in the upper part, a UHF receiver; in the lower part, an IF amplifier, a video amplifier, and two power supply units. The central bay contains: an image control device for video and intermediate frequency, a modulator, and a power unit.

The relay station equipment differs from the one described by absence of the modulator and power unit. The sidewalls of the bay are closed with removable partitions. The individual structural elements are mounted on vertical panels, perpen-

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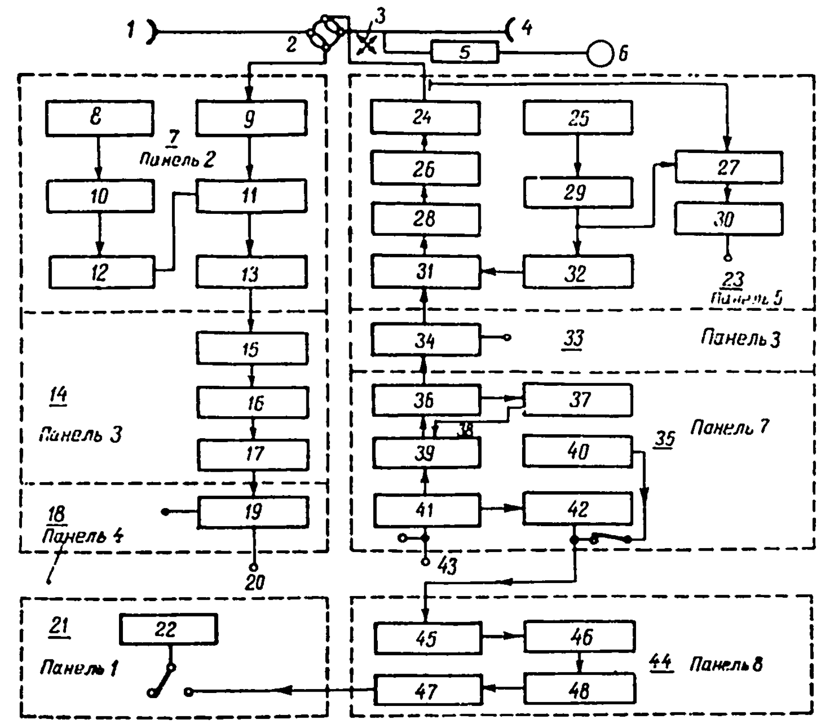

Fig.1 - Block Diagram of the Terminal Station

1. Receiving antenna.
2. Antenna switch.
3. Directional coupler.
4. Transmission antenna.
5. Detector.
6. Indicator of standing-wave factor.
7. Panel 2.
8. Quartz oscillator.
9. Filter
10. Frequency multiplier.
11. Mixer.
12. Filter.
13. Preamplifier.
14. Panel 3.
15. IF amplifier.
16. Limiter.
17. Discriminator.
18. Panel 4.
19. Video amplifier.
20. Video signal output.
21. Panel 1.
22. Cathode-ray oscillograph.
23. Panel 5.
24. Filter.
25. Quartz oscillator.
26. Amplifier.
27. Detector.
28. Side filter.
29. Multiplier.
30. Frequency multiplier.
31. Modulator.
32. Amplifier.
33. Panel 6.
34. Amplifier.
35. Panel 7.
36. Amplifier.
37. Reference oscillator of black level frequency.
38. Erasure signal.
39. Modulator.
40. Calibration oscillator.
41. Amplifier.
42. Filter.
43. Video signal input.
44. Panel 8.
45. Amplifier.
46. Limiter.
47. Amplifier.
48. Discriminator.

dicularly distributed on the front walls of the bay. When the sidewalls are removed, access is gained to all circuits and tubes.

The Electrical System of the Apparatus

Figure 1 gives a block diagram of the terminal station. In this diagram, the elements of the basic receiving channel 1, distributed in panels 2, 3, and 4 are illustrated in a vertical row from the right, and the elements of the auxiliary channel from the left. Analogously, the elements of the basic transmitter channel, set up in panels, 5, 6, and 7, are shown in a vertical row from the left, and the auxiliary devices are seen from the right. The diagram also shows the series in which the different stages are set up in the individual panels.

The entire video signal with blacking-out and synchronized pulses (total amplitude 1.5 v in a load of 75 ohms) is sent to the transmitter.

After amplification, this signal if fed to the modulator, at whose output a signal of intermediate frequency of 75 ± 6 mc is formed. The frequency of 72.36 mc corresponds to the black level. It is equal to the frequency of the reference oscillator with quartz stabilization. In the case of frequency separation, an erasure voltage is generated which controls the frequency of the modulator and produces a suitable correction.

The modulator panel also contains a calibration oscillator, which can change over in frequency from 68 to 83 mc for testing the system. The synchronizing level corresponds to a frequency of 69 mc, the maximum white level corresponds to 81 mc, and the black level (as was the case above) to a frequency of 72.36 mc.

The intermediate-frequency signal received at the output of the modulator is amplified by the powerful amplifier (panel 6) from 0.5 v up to 20 v and is fed to the modulator of the UHF transmitter (panel 5). Here it is mixed with the heterodyne signal, whose frequency is quartz-stabilized, and the required sideband is separated by a filter. The sideband signal is supplied to the antenna across an amplifier, an ultrahigh-frequency filter and the antenna switch. The antenna filter is connected to a meter for controlling the level of the transmitted signal. The feeder also contains a direction coupler, permitting control of the standing-wave coefficient for balancing the antenna. The quality of the image in the emitted signal is confirmed by the control device.

The signal received through the antenna switch and the ultrahigh-frequency input filter is fed to the mixer. The heterodyne signal is generated by the oscillator with quartz stabilization and series multiplication in several stages of the multiplier. The intermediate-frequency signal received at the output of the mixer after amplification in the preamplifier (UPCh) of the "cascade" type is fed to the basic IF amplifier (panel 3). This amplifier has automatically controlled amplification which compensates for changes in signal within limits of 40 db. Behind the limiter, the signal enters the discriminator, at whose output the video signal is received. This is conducted to the video amplifier (panel 4), where it is again amplified. If necessary, the black level is regulated here and the signal flows to the coaxial jack.

The operation of the apparatus may be tested with intermediate frequency by means of the test receiver, mounted in panel 8, with the oscillograph (in panel 1). The receiver modifies the FM signal of intermediate frequency into an AM video signal, while the oscillograph serves for visual monitoring of the televised image.

The panel underneath the tube contains the switch which connects the cathode-ray tube with the control device. This control device has frequency scanning, equal to the frequency of the scan line and of the vertical sweep. It permits control of level of the televised signal.

The apparatus of the relay station are almost identical with the apparatus of the terminal station, except for panel 7. The modulator is set in the panel in the terminal station, while the intermediary station contains a device for obtaining the heterodyne signal of the receiver from the heterodyne signal of the transmitter by

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means of a frequency shift at 210 mc. The intermediate frequency signal from the output of the IF amplifier is fed directly to the powerful amplifier.

Parabolic antennas with reflectors, having a diameter of 3 m, are used in this system. The matching of the antenna with the feeder line of the transmitter is such that the coefficient of the traveling wave is more than 0.95.

The apparatus described permits all necessary measurements to be made.

(Electrical Communications, Vol.32, No.4, Dec. 1955).

FOREIGN PATENTS

<u>From the Editor</u>. The editor would like to inform his readers that since concise information is available on some foreign patents in the field of multichannel high-frequency communication, photocopies of patents (descriptions in the original language, diagrams) may be obtained personally or c.o.d. at the All-Union Patent-Technical Library of the Committee on Inventions and Discoveries with the Council of Ministers of the USSR.

Address of the library: Moscow-Center, Proezd Serova, 4, Entrance 7a. Telephone for information: B8-64-52.

Photocopies of the patents may also be procured through the order office of the All-Union Institute of Scientific and Technical Information, Academy of Science, USSR and Gostekhniki.

Address of the Institute: Moscow, D-219, Baltiyskiy pos., d. 42b. Telephone: D7-00-10, Ext.51.

Patent FRG (Federal German Republic), Class 21a^{4}, 50_{02}, No.935 675, 24.11.55. Neiswinter. Two-Way Single-Sideband Communication System. [Neiswinter (Western Electric Company Inc.) Zweiweg-Einseitenband-Uebertragungs-system].

A multiple-channel two-way system with a single side frequency is presented which is characterized by its carrier currents, transmitted in every sending direction without interruption, differing in frequency by several hundred cycles. The sidebands used in both transmission directions overlap to a considerable degree, due to the fact that, in the given frequency band, a greater number of channels can be installed than is usually the case. Together with this, the control of receivers is obtained only by means of the carrier current transmitted from the opposite station. The possibility of an independent carrier current acting on the receiver is excluded. Double-lead terminals are connected to the four-wire part of the channel by a spe-

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cial decoupling device.

Patent FRG (Federal German Republic), Class $21a^2$, 39_{10}, No.928 110, 23.05.55. Haessler. The Line Spectrum for the Four-Wire HF Telephoning System [Haessler (Allgemeine Elektrizitaets-Gesellschaft) Vierdraht-Traegerfrequenzuebertragungssystem].

The frequency spectrum for multichannel systems of HF telephony is discussed, which is adapted to the organization of communications with both two-wire and four-wire systems. It is possible to form groups by this with a different number of channels, making the transition easier from the two-wire system to the four-wire and vice versa. For the eight-channel system operation on a two-wire circuit, the spectra assigned as the base are 6 - 30 and 36 - 60 kc, while for the sixteen channel they are the spectra 12 - 60 and 72 - 120 kc.

Patent FRG, Class $21a^2$, 39_{10}, No.932 202, 07.02.55. Graf, Bath. Multichannel System of HF Telephony, whose Line Spectrum is Formed from the Bands of Individual Subgroups. [Graf, Bath (Siemens & Halske A.G.) Mehrfachtraegerfrequenz-Nachrichten-uebertragungssystem].

The object of the invention is a multichannel HF system for compressing overhead lines, characterized by its band frequency, transmitted in every beamed direction, being formed from individual three-channel bands, arranged symmetrically relative to a frequency of 90 kc.

The first three-channel group occupies a band of		72 - 84 kc and 96 - 104 kc
second	"	" 60 - 72 kc and 104 - 120 kc
third	"	" 48 - 60 kc and 120 - 132 kc
fourth	"	" 36 - 48 kc and 132 - 144 kc

Each of the three-channel groups may work independently, using common linear amplification and common oscillation equipment at the terminal stations.

Patent FRG, Class $21a^2$, 39_{10}, No. 923 201, 07.02.55. Haessler, Christiansen. Reception Device for Multichannel HF Telephone System. [Haessler, Christiansen

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(Standart Elektrizitaets-Gesellschaft A.G.) Empfangsanordnung fuer Mehrkanal-Traegerfrequenz-Systeme].

The object of this invention is a reception device for multichannel systems of HF telephony at short distances from the transmitter of both side frequencies, distinguished by its multiple use of electron tubes. One and the same tube is used for demodulation, for amplification of LF current, call signals, and for pulses in dialing. In addition, the tube is used as a carrier-current oscillator for the modulator.

Patent FRG, Class $21a^4$, 54, No. 932 686, 05.09.55. Hugenholtz. Receiver for Single Sideband Frequencies with a Carrier Frequency Trim. [Hugenholtz (N.V.Philips' Gloeilampenfabrieken, Holland) Empfaenger zum Empfang von Einseitenbandsignalen].

The schematic for the reception of a single sideband frequency is presented, characterized by it constantly transmitting two auxiliary frequencies for synchronizing the carrier currents of the receiving and sending stations with the transmitting station. These frequencies are harmonics, different in number, of the fundamental frequency (rigidly connected to the carrier frequency of the transmitting station). At the receiving station the resultant frequency received by comparison with the auxiliary frequencies, is controlled by a motor, serving to align the carrier frequency of the station.

Patent FRG, Class $21a^2$, 39_{10}, No.936 518, 15.12.55. Werther. Method of Forming Frequency Line Spectra in Multichannel HF Telephone Systems. [Werther (Siemens & Halske A.G.) Anordnung fuer Traegerfrequenz-Fernsprechuebertragungsanlagen].

The method of forming a line spectrum of frequencies for multichannel HF telephone systems is presented, characterized by low-frequency channels combined into subgroups in three of the channels. By means of carrier currents with a frequency of 4.8 and 12 kc, each subgroup is transferred into a band of 4 - 16 kc from which, by means of a second modulation step it is carried over into the needed part of the spectrum. To form the band 36 - 84 kc, carrier frequencies of 52, 64, 56, and 68 kc are recommended. An inversion of the frequencies (if need be) can be realized by subsequent stages by converting the frequencies.

Patent FRG, Class $21a^4$, 39, No.931 603, 11.08.55. Pungs, Meinshausen. Method of Magnetic Demodulation. [Pungs, Meinshausen. Verfahren zur magnetischen Demodulation].

The method of demodulation amplitude modulated signals is treated, characterized by using, as a nonlinear element, a coil with a ferromagnetic core which is saturated with very low field intensity. The operating point is selected at the place of magnetized inflection. Magnetization about the inflection point is caused by direct current.

Patent FRG, Class $21a^2$, 36_{10}, No.931 354, 08.08.55. Bruehl. Device for Noise Measurement of Nonlinear Harmonic Distortion and other Interfering Noises in Multiple Channel Systems of HF Telephony. [Bruehl (Telefunken G.m.b.H). Anordnung zur Messung von Klirrgeraeuschen und sonstigen Stoergeraeuschen bei einem Vielkanalsystem].

The load method of the group tract of a multiple-channel HF telephone system is presented by means of a limited number of noise oscillators or by means of a single noise oscillator and elements of the receiver route of a multiple-channel system end station. The route elements through which the measured channel passes are not loaded so that the interference voltage may be measured in this channel.

USA Patent, Class 250 - 36, No.2 710 920, 14.06.55. Appert, Erickson. Multifrequency Generator for Multichannel HF Telephone Systems [Appert, Erickson (Lenkurt Electric Co.,Inc.) Multichannel Frequency Generator].

This invention is a generator of carrier currents for multichannel HF telephone systems. A crystal oscillator serves as the source of fundamental frequency at whose output a broad-band amplifier is inserted with a large number of feedback circuits. Each of these feedback circuits contains a band filter, passing one of the carrier frequencies needed by the system.

Canadian Patent, No.512 163, 26.04.55. Cooper. System of HF Communication,

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Equipped with a Device for Determining the Point of Damage. [Cooper. Systèmes de Communication].

An HF communication system is discussed which consists of two end stations and a series of attended and unattended stations, characterized by the end station being equipped with pulsating instruments for determining the point where damage has occurred.

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NEW BOOKS

M.Strett. Semiconductor Devices. Principles of Operation, Properties and Use. Translated from the German 1954 edition by V.F.Rakhmanov and V.P.Shchelkin, Ed. by Prof. G.I.Atabekov. Gosenergoidat, 1956. 208 pp, 10,000 copies, price, bound, 7 r.

Semiconductor Devices and their Use. Collection of Articles Edited by Ya.A.Fedotov. Issue No.1 "Sovetskoe Radio" 1956. 624 pp, price, bound, 16 r 25 k.

In this collection, 36 articles are compiled written in 1955 and expounding the physical principles of the function of semiconductor devices. Discussions are given on the basic characteristics of domestic low and high-power semiconductor triodes, powerful triodes and tetrodes, and fluctuation noises; the temperature function of triode parameters and means of temperature compensation as well as other questions in semiconductor technology are treated.

S.D.Gvozdover. Theory of Ultrahigh Frequency Electronic Devices. Gostekhizdat, 1956. 528 pp, 25,000 copies, price, bound, 10 r 90 k.

Textbook for higher education institutions.

S.V.Savchenko. The Photoelectric Effect and its Technical Applications. Gostekhizdat, 1956, 116 pp, 30,000 copies, price 1 r 95 k.

The basic laws of the photoelectric effect are discussed in a popular form, describing the different aspects of photocells and their practical use in industry.

M.A.Lebedinskiy. Electric Vacuum Materials. Gosenergoizdat, 1956, 336 pp, 15,000 copies, price, bound, 8 r.

Textbook for the technical schools of the Ministry of Radiotechnical Industry.

M.A.Rozenblat. Magnetic Amplifiers. "Sovetskoe Radio", 1956, 824 pp, price, bound, 23 r.

The problems of the theory of calculating the basic types of magnetic amplifiers in practical use are considered in this book. Particular attention is paid to the development of the basic principles of constructing magnetic amplifiers and the

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physical nature of processes involved. The fundamental means and methods of improving their characteristics are carefully examined, as well as the reduction of inertia, the lowering of sensitivity thresholds, and the raising of the stability of magnetic amplifiers.

B.M.Betin. Radio Transmission Devices. Theory and Computation. Gosenergoizdat, 1956, 352 pp, 15,000 copies, price, bound, 7 r 70 k.

Manual for technical schools of the Ministry of Radiotechnical Industry.

N.N.Il'ina. Characteristics of Wiring Circuits of Modern Radio Broadcast Transmitters. Svyaz'izdat, 1956, 48 pp, 15,000 copies, price 1 r 50 k.

Brochure of the series "Lectures on Communications Technology".

V.F.Barkan and V.K.Zhdanov. Radio Receiving Systems. Oborongiz, 1956, 496 pp, 35,000 copies, price, bound, 12 r 25 k.

Textbook for aviation technical schools.

V.K.Zvorykin and D.A.Morton. Television. Electronic Problems in Color and Monochrome Image Transmission. Translated from English. Edited by Prof.S.K.Katayev. Foreign Literature Publishing House, 1956, 780 pp, price, bound, 46 r 10 k.

The physical basis of black-and-white and color television is carefully examined in this book, with a description of the most important circuits and systems of television equipment. In addition, a great amount of information in the form of equations is introduced here, based on practical tests in the laboratories and factories of American radio corporations in the processing, production, and use of television sets.

M.I.Krivosheev. Measurements in Television Equipment. Svyaz'izdat, 1956, 62 pp, 11,000 copies, price, 2 r 10 k.

Brochure in the series "Lectures on Communications Technology".

N.N.Garnobskiy. Theoretical Basis of Electric Wire Connections. Part I. General Theory of Passive Linear Circuits with Lumped Constants. Svyaz'izdat, 1956, 692 pp, 10,000 copies, price, bound, 21 r 50 k.

Intercity Communication Lines. A Collection of Translated Articles. Edited by V.N.Kuleshov. Svyaz'izdat, 1956, 132 pp, 6,000 copies, price, 5 r 75 k.

This collection contains nine translated articles on coaxial cables laid across the Atlantic Ocean and on the territory of Italy, on cables in rural areas, on new cable-laying methods in sewers and trenches, on methods of mounting high-frequency symmetrical cables and on their application to city and long-distance cable lines.

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