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SOME GEOMETRICAL PROPERTIES OF THE OPTIMUM CODE

by

N.K.Ignat'yev

The problem is discussed of selecting the best configuration of signal space for distributing therein the dots of the optimum-code signal (i.e., a code ensuring the maximum entropy of the signal at all other conditions remaining equal). Instances are investigated when the dots of a signal become distributed within an n-dimensional sphere, on the surface of that sphere, and within an n-dimensional cube.

It can be concluded from the proof of Shannon's theorem (Bibl.1) concerning maximum channel capacity that the ideal code ensuring such capacity is geometrically characterized by the distribution of signal dots on the surface of an n-dimensional sphere with a radius of $\sqrt{nP}$ (where P is the signal power), and statistically characterized by the normal distribution of the signal and the uniform probability of all code combinations.

But these characteristics pertain only to the ideal code where the number of elementary sendings n in every code combination tends toward infinity and the total amplitude of the signal is uncurtailed by anything.

As for the analogous geometrical and statistical characteristics of the optimum real code (hereafter referred to as optimum code), these remain unclarified. The present paper makes an attempt at defining these characteristics.

Hereafter the optimum code will be construed as referring to a uniform n-valued code which, at all other conditions (such as equality of signal power or signal scope and equality of noiseproof feature) being equal, has the maximum entropy.

We will judge the equality of the noiseproof feature of codes by the volume of space corresponding to every signal dot, and the entropy of codes by the volume of the entire space of signal. At such an approach toward the problem posed, there is no need for exploring the geometry of the distribution of signal dots within the given volume of space*. Thus, we need merely be concerned with the best configuration of that volume. At such an approach the results will be very generalized but will still apply only for codes with a high entropy, in which the manner of the mutual distribution of signal dots within a given volume of space (at the proportional occupation of volume) exerts virtually no influence on signal power.

It is assumed that a priori selected probabilities of appearance cannot be imparted to code combinations and that, most generally speaking, these probabilities should be regarded as uniform.

Volume, Power and Entropy of Signals

Let us establish certain initial correlations:

Let $x_1, x_2, \ldots, x_n$ be the coordinates of a signal dot in the n-dimensional space which also express the successive voltages of the elementary sendings of a given signal.

To establish some simplest relationships between the geometrical and energy characteristics of the signal let us assume that it is emitted at a resistance of 1 ohm and that it consists of elementary pulses with a duration of 1 sec.

Thereupon the value

$$r^2 = x_1^2 + x_2^2 + \ldots + x_n^2$$

will express the square of the distance between a signal dot from the origin of coordinates, and at the same time it also expresses signal energy, while the value

*It is known that the optimum distribution of signal dots inside a given volume of space is one at which the signal dots coincide with the centers of the most densely aligned n-dimensional spheres.

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$$P_k = \frac{r^2}{n}$$

expresses the power of the given signal.

The mean signal power for all M code combinations is (if they all have a uniform probability):

$$P = \frac{1}{M} \sum_{k=1}^{M} P_k.$$

If the volume of signal space is filled uniformly with signal dots, and if a sufficiently great number of signal dots is distributed along every direction of signal space (i.e., if the entropy of the signal is sufficiently high), the summation of individual dots can be replaced by integration of the volume of signal space, V:

$$P = \frac{1}{V} \int_V \frac{r^2}{n} \, dv.$$

For simplicity, let us assume that every signal dot is corresponded by a space volume equal to the unit. Then the volume of signal space V will express the number of signal dots, i.e., V.= M. Consequently, the signal entropy is

$$H = \frac{1}{n} \log_2 V.$$

In instances when signal dots become distributed on the surface of a certain figure and not within its volume, volume V in eqs.(1) and (2) should be substituted by surface S.

We shall establish the relationships between signal power and enthropy on the basis of expressions (1) and (2).

Normal Signal Distribution

Let us examine the basic properties of a signal with normal distribution, whose probability density is expressed as follows:

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$$p_N(x_1, x_2, \ldots, x_n) = \frac{1}{(2\pi P)^{\frac{n}{2}}} e^{-\frac{1}{2P}\sum_{i=1}^{n} x_i^2} \tag{3}$$

The entropy of the signal with such a distribution reaches, as is known, its maximum possible value equal to

$$H_N = \log_2 \sqrt{2\pi e P}. \tag{4}$$

On considering distribution (3) as a function of a dot in an n-dimensional space it can be easily concluded that p_N is the function of distance r between the signal dot and the beginning of the coordinates, i.e.,

$$p_N(x_1, x_2, \ldots, x_n) = \frac{1}{(2\pi P)^{\frac{n}{2}}} e^{-\frac{1}{2P}r^2} \tag{5}$$

Thus, the probability of appearance of various signal dots depends only on the value of the radius of r on which they are located.

From the geometrical viewpoint, the distribution $p_N(x_1, x_2, \ldots, x_n)$ can be regarded as the volume density of the signal. Therefore, upon multiplying it by the distribution density of the space volume along the radius of r, we obtain the radial density of the signal p(r). Considering that the volume of an n-dimensional sphere with radius r equals:

$$v = \frac{\pi^{\frac{n}{2}} r^n}{\Gamma\left(\frac{n}{2}+1\right)}, \tag{6}$$

the density of distribution of space volume along the radius is

$$\frac{dv}{dr} = \frac{\pi^{\frac{n}{2}} n r^{n-1}}{\Gamma\left(\frac{n}{2}+1\right)}. \tag{7}$$

On composing the product of eqs.(5) and (7) we obtain

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$$p(r) = \frac{n}{\Gamma\left(\frac{n}{2}+1\right)\sqrt{2P}} \left(\frac{r}{\sqrt{2P}}\right)^{n-1} e^{-\left(\frac{r}{\sqrt{2P}}\right)^2} \qquad (8)$$

This distribution, known under the name of χ^2-distribution, reaches its maximum at the following value of radius:

$$r = \sqrt{(n-1)P}$$

and, in proportion with the increase in n, it approximates a distribution of the following form:

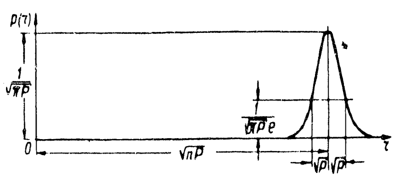

Fig.1

$$p(r) \approx \frac{1}{\sqrt{\pi P}} e^{-\left(\frac{r-\sqrt{nP}}{\sqrt{P}}\right)^2}, \qquad (9)$$

which is illustrated in Fig.1.

From the above formula it can be concluded that at $n \to \infty$ all signal dots will lie on a single radius equal to $\sqrt{nP}$. Here they will all have a uniform probability.

These characteristics of normal distribution, which appears to be optimum from the viewpoint of utilization of the signal power, serve as a foundation for the belief that all dots of a signal have to be distributed on the surface of an n-dimensional sphere in order to obtain the optimum code (i.e., that the "surface-spherical" signal distribution is to be used).

Nonetheless, from the viewpoint of optimal utilization of signal power at a finite n, it is preferable to fill the entire volume of the n-dimensional sphere with signal dots (i.e., to employ the "volume-spherical" signal-distribution).

Let us investigate the basic characteristics of these distributions.

Spherical Signal Distributions

Let us begin by examining an instance of volume-spherical distribution which can be expressed as follows:

$$p_V = (x_1, x_2, \ldots, x_n) = \begin{cases} \frac{1}{V} \text{ at } \sum_{i=1}^{n} x_i^2 \leqslant R^2 \\ 0 \text{ at } \sum_{i=1}^{n} x_i^2 > R^2, \end{cases} \tag{10}$$

where R is radius of the sphere confining the space volume of a signal.

In this case, the total volume of the signal space will be:

$$V = \frac{\pi^{\frac{n}{2}} R^n}{\Gamma\left(\frac{n}{2}+1\right)}. \tag{11}$$

The volume of the space limited by the current radius r will be

$$v = \frac{\pi^{\frac{n}{2}} r^n}{\Gamma\left(\frac{n}{2}+1\right)}$$

while its differential will be

$$dv = \frac{\pi^{\frac{n}{2}} n r^{n-1}}{\Gamma\left(\frac{n}{2}+1\right)} dr.$$

Consequently, in accordance with eq.(1), it can be written that

$$P = \frac{1}{R^n} \int_0^R r^{n-1} dr,$$

whence we obtain

$$P = \frac{R^2}{n+2}$$

or

$$R = \sqrt{(n+2)P}. \tag{12}$$

On substituting eq.(12) into eq.(11), we obtain

$$V = \frac{[\pi (n+2) P]^{\frac{n}{2}}}{\Gamma\left(\frac{n}{2}+1\right)}. \tag{13}$$

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On then substituting eq.(13) into eq.(2) we obtain the entropy of the signal

$$H_V = \log_2 \frac{\sqrt{n(n+2)P}}{\Gamma\left(\frac{n}{2}+1\right)^{\frac{1}{n}}}. \tag{14}$$

We proceed with our examination of the instance of surface-spherical distribution, which can be thus expressed:

$$p_S(x_1, x_2, \ldots, x_n) = \begin{cases} 0 & \text{at } \sum_{i=1}^{n} x_i^2 < R^2 \\ \frac{1}{S} & \text{at } \sum_{i=1}^{n} x_i^2 = R^2 \\ 0 & \text{at } \sum_{i=1}^{n} x_i^2 > R^2, \end{cases} \tag{15}$$

where

$$S = \frac{\pi^{\frac{n}{2}} n R^{n-1}}{\Gamma\left(\frac{n}{2}+1\right)} \tag{16}$$

and which expresses the surface of the sphere of radius R on which the signal dots are distributed.

Inasmuch as in the given case r = R, in accordance with eq.(1)

$$R = \sqrt{nP}$$

and consequently

$$S = \frac{\pi^{\frac{n}{2}} n (nP)^{\frac{n-1}{2}}}{\Gamma\left(\frac{n}{2}+1\right)}, \tag{17}$$

whence, in accordance with eq.(2), we obtain the entropy of the signal.

$$H_S = \log_2 \frac{\sqrt{\pi n P^{\frac{n-1}{n}}}}{\Gamma\left(\frac{n}{2}+1\right)^{\frac{1}{n}}}. \tag{18}$$

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Let us now conduct a comparison of the entropy values H_V and H_S in accordance with eqs.(14) and (18).

First of all, considering that

$$\lim_{n\to\infty}\Gamma\left(\frac{n}{2}+1\right)^{\frac{1}{n}}=\sqrt{\frac{n}{2e}},$$

we therefore obtain at $n \to \infty$

$$H_V = H_S = H_N.$$

Thus both the above-examined distributions become equivalent to each other at their limit, and then they also become equivalent to normal distribution. This result can be attributed to the circumstance that at an increase in the number of dimensions the entire volume of the n-dimensional sphere becomes concentrated at its surface.

For comparison of the entropy values H_V and H_S at finite values of n, let us express them reciprocally by eliminating P. Proceeding from eqs.(14) and (18) we obtain

$$H_S=\frac{n-1}{n}H_V+\frac{1}{n}\log_2\frac{n\sqrt{\pi}\left(\frac{n}{n+2}\right)^{\frac{n-1}{2}}}{\Gamma\left(\frac{n}{2}+1\right)^{\frac{1}{n}}}.$$

At high values of entropy the second term of this formula can be neglected, and it can be considered that

$$H_S \approx \frac{n-1}{n}H_V. \tag{19}$$

From the above formula it can be concluded that the transition from volume-spherical to surface-spherical distribution leads to the loss of $\frac{1}{n}$ part of the signal entropy, which is equivalent to the loss of one of the n dimensions of the signal space.

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Let us investigate the geometrical sense of the loss of one spatial dimension at the use of a surface-spherical signal distribution.

The position of a signal dot on the surface of an n-dimensional sphere of a known radius R can be completely determined by the presence of n - 1 coordinates of that dot, since the n^{th} coordinate can be found from the equation of the sphere

$$x_1^2 + x_2^2 + \ldots + x_n^2 = R^2.$$

However, if the radius of the sphere is not known (i.e., if the scale of the coordinates of the dot lying on the sphere's surface is not known), then it is necessary to know all n of the coordinates of that dot in order to determine its position completely Thus, knowledge of the n^{th} coordinate, which is redundant from the viewpoint of transmission of information at a known signal scale, becomes a prerequisite when the signal scale is not known.

Accordingly, when a surface-spherical signal distribution is employed, one of the dimensions of space is "expended" on compensating the loss of the signal scale.

The possibility of effecting reception of a coded signal at any (so long as sufficiently slow) fluctuations in its level (i.e., despite the loss of its scale) is of a tremendous practical significance. The core of the matter is that long-distance transmission of signals invariably causes the level of these signals to be subject to chaotic fluctuations, owing to the changes in conditions of propagation and also owing to changes in the amplification factor of the transmission channel. These fluctuations are, as a rule, so slow compared with the duration of code combinations that the amplitude ratios within every code combination can be regarded as unchanged. In the case of surface-spherical distribution such fluctuations in the signal level, which geometrically correspond only to changes in the length of the radius-vector of a signal dot, do not incur the appearance of any error.

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Cubic Distribution of Signal

In instances when the limiting factor in the increase in channel capacity is not signal power but signal scope, the distribution of signal dots is bound to take place within an n-dimensional cube.

If the signal scope is denoted by L, then the volume of the signal space

$$V = L^n. \qquad (20)$$

On further proceeding from eq.(1) the formula for mean signal power can be written as follows:

$$P = \frac{1}{nL^n} \int_{-\frac{L}{2}}^{\frac{L}{2}} \int_{-\frac{L}{2}}^{\frac{L}{2}} \cdots \int_{-\frac{L}{2}}^{\frac{L}{2}} (x_1^2 + x_2^2 + \ldots + x_n^2)\, dx_1 dx_2 \ldots dx_n.$$

Whence, after integration, we obtain $P = \frac{L^2}{12}$

or

$$L = \sqrt{12P}. \qquad (21)$$

Consequently, in accordance with eq.(20)

$$V = (12P)^{\frac{n}{2}} \qquad (22)$$

and the signal entropy becomes

$$H_Q = \log_2 \sqrt{12P}. \qquad (23)$$

On comparing eqs.(14) and (23), we obtain

$$H_V - H_Q = \log_2 \frac{\sqrt{\frac{\pi (n+2)}{12}}}{\Gamma\left(\frac{n}{2}+1\right)^{\frac{1}{n}}}. \qquad (24)$$

From the above formula it can be concluded that the difference in the entropies of volume-spherical and cubic distributions at an increase in n (from 1 to ∞) in-

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creases from zero to $\log_2 \sqrt{\frac{\pi e}{6}}$ (which altogether amounts to about $\frac{1}{4}$ of a binary digit).

A code with cubic (or more exactly, with volume-cubic) signal distribution displays the same shortcoming as the code with volume-spherical distribution: it requires strict maintenance of the constancy of signal level.

To eliminate this shortcoming it is possible - just as in the case of spherical distribution - to eliminate the signal dots from the interior of the volume and to pass over to "surface-cubic" distribution, which will likewise lead to the loss of approximately $\frac{1}{n}$ part of the entropy of the signal.

Conclusions

The above analysis demonstrated that the projection of the geometric properties of the ideal code onto a real code is not the best solution of the problem posed.

At a limited number of sendings in a code combination, the maximum entropy is ensured by volume-spherical distribution.

At a limited signal scope, the maximum entropy is ensured by volume-cubic distribution. Such an entropy proves to be slightly smaller than the entropy of volume-spherical distribution, but the realization of the optimum code can then become much simpler. For instance, a pulse code with a volume-cubic signal distribution is materialized in the form of the pulse expression of the number of code combination in the calculating system, based on b (where b is number of code levels).

Volume distributions can also be employed at a variable signal scale. For this purpose, scale pulses have to be introduced into the signal. The slower the possible fluctuations in signal scale, the rarer can be the distribution of these pulses, and the smaller will be the reduction in entropy of the signal owing to these pulses.

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FLUCTUATION OVERSHOOTS AND THEIR CORRELATION

by

V.I.Tikhonov

This article cites the results of an experimental investigation of the distribution of the overshoots of normal and Rayleigh fluctuations, in terms of length, and it evaluates their correlation.

Electronic relays and lock-on circuits are widely used in various coding systems. For an analysis of the effect of random noise on an electronic relay with a single stable equilibrium position the following relay model may be used in a number of cases: it operates whenever the noise voltage $\xi(t)$ exceeds the relay operating level a during a time of $\Delta t > \tau_0$, where τ_0 is the time of relay operation. Here it is possible to compute with sufficient ease the mean number of spurious operations

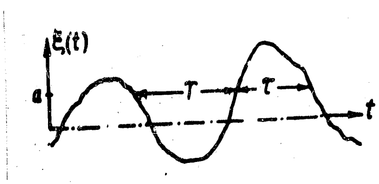

Fig.1

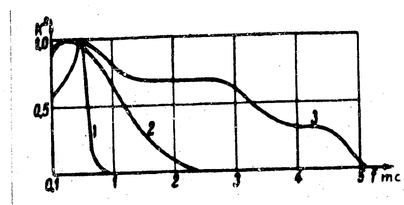

Fig.2

of the relay and to determine also the other statistical characteristics, if the values of noise on the relay operating level, spaced by the duration of the relay pulse, are not correlated (Bibl.1). Such a postulate holds true for the pulse noises subject to Poisson's law. However, it may be doubtful whether this postulate is also valid for fluctuation noise.

It will be demonstrated below that fluctuation overshoots can be considered as ... uncorrelated even at a relay operating level of $a = \sigma$, where σ^2 is the scattering of fluctuations of $\xi(t)$. Here an overshoot is construed as referring to an event where the fluctuation voltage becomes $\xi(t) \geq a$ during a time τ termed the pulse duration (Fig.1). The interval T between two successive overshoots is analogously defined.

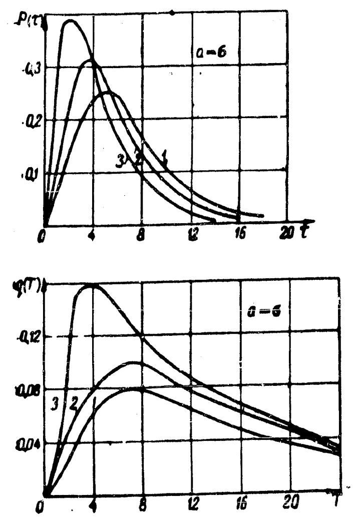

Fig.3

A special experiment has been devised to solve the above problem. The normal fluctuation noise at the outputs of three amplifiers was photographed from the screen of an OK-17 oscillograph. The relationship between the square of the amplification factor and the frequency for each amplifier is illustrated in Fig.2.

The numerous oscillograms taken were processed and used in plotting the functions of the probability density $p(\tau)$ and $q(T)$ for overshoot durations τ and intervals T between them on the level $a = \sigma$ (Fig.3).

We will demonstrate that a nominal division of τ and T corresponds to a time of 0.17 μsec for the first amplifier, 0.12 μsec for the second amplifier, and 0.087 sec for the third amplifier.

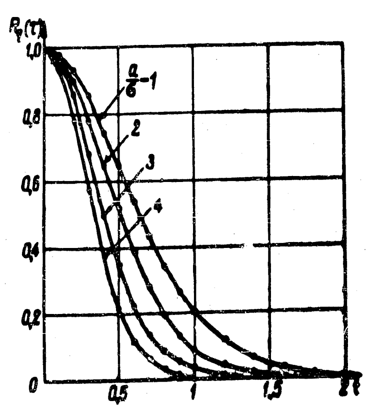

Fig.4

Of the characteristics of the three amplifiers only the frequency characteristic of the second amplifier is sufficiently well approximated by the following function:

$$K^2(f) = \exp\left(-\frac{1}{\alpha}\pi^2 f^2\right), \quad \alpha = 20.5 \ \mu sec^{-2}. \quad (1)$$

Correspondingly, the correlation factor for normal fluctuations at the output of the second amplifier will equal

$$) = \exp(-t^2), \quad t = \alpha^{\frac{1}{2}}\tau. \qquad (2)$$

Let us now examine the fluctuations $\eta(t)$ obtainable from normal fluctuations of $\xi(t)$ by the following transform:

$$\eta(t)=\begin{cases}\xi(t) & \text{at } \xi(t)>a\\ 0 & \text{at } \xi(t)<a.\end{cases} \tag{3}$$

On using the expansion of the two-dimensional normal function of probability density into a series, according to the correlation factor (Bibl.2,3), it can be demonstrated that the correlation function for $\eta(t)$ is determined by the following formula:

$$K_{\eta}(\tau)=\sigma^2\left\{\Phi^2\left(-\frac{a}{\sigma}\right) R_{\xi}(\tau)+\sum_{n=2}^{\infty}\frac{1}{n!}\left[\Phi^{(n-1)}\left(\frac{a}{\sigma}\right)\right]^2 R_{\xi}^{n}(\tau)\right\} \tag{4}$$

where σ^2 is the scattering of fluctuations of $\xi(t)$ and $\Phi(x)$ is the probability integral.

We obtain therefrom the equation of the correlation factor:

$$R_{\eta}(\tau)=\sigma^2\sigma_{\eta}^{-2}\left\{\Phi^2\left(-\frac{a}{\xi}\right) R_{\xi}(\tau)+\sum_{n=2}^{\infty}\frac{1}{n!}\left[\Phi^{(n-1)}\left(\frac{a}{\sigma}\right)\right]^2 R_{\xi}^{n}(\tau)\right\}, \tag{5}$$

where σ^2 is the scattering of "limited" fluctuations of $\eta(t)$, equal to

$$\sigma_{\eta}^2=\sigma^2\left\{\Phi^2\left(-\frac{a}{\sigma}\right)+\sum_{n=2}^{\infty}\frac{1}{n!}\left[\Phi^{(n-1)}\left(\frac{a}{\sigma}\right)\right]^2\right\}. \tag{6}$$

Tables of the derivatives of the probability integral (Bibl.3) were used in computing the values of the correlation factor $R_{\eta}(\tau)$, taking into consideration the terms of the series (4) up to n = 7, at four values of the operating level $\frac{a}{\sigma}$ = = 1, 2, 3, 4, when the correlation factor for $\xi(t)$ has the form of eq.(2). The results of the computations are depicted in Fig.4.

The graph in Fig.4, for $a = \sigma$, shows that $R_{\eta}(\tau) = 0.01$ at $t = \alpha^{\frac{1}{2}} t \approx 2$, i.e., at $\tau = 2\alpha^{-\frac{1}{2}} \approx 0.44\ \mu$sec. If we pass over to the nominal divisions used in Fig.3, we find that, if two adjacent overshoots on the $a = \sigma$ level are separated by an interval of $T > 4$, then such two overshoots can be considered as being virtually noncorrelated, in view of the fact that the correlation factor for them equals $R_\eta(\tau > 4) < 10^{-2}$. However, the relative number of intervals with $T < 4$ is low and does not exceed 12%.

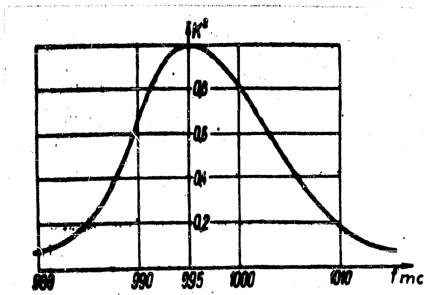

Fig.5

It follows that the overshoots of normal fluctuations with a correlation factor having the form of eq.(2) on the $a \geq \sigma$ level can be considered approximately noncorrelated.

An analogous experiment was conducted for the purpose of an approximate qualitative appraisal of the overshoots of Rayleigh scattering obtained at the outputs of the detectors in standard radio receivers. The relationship between frequency and the square of the amplification factor of a channel of an IF amplifier radio receiver is depicted in Fig.5. The detector was represented by connecting a 6Kh6 diode in series and a RC circuit in parallel ($R = 100$ kilohms, $C = 200$ $\mu\mu f$).

The results of the processing of 300 oscillograms, giving a picture of the distribution of the overshoots of Rayleigh fluctuations and of the intervals between them are given in Table 1. In the given case, a nominal division for τ and T corresponds to a time of 10.2 μsec. The bottom of each column gives the total number N of overshoots (intervals), the mean value τ of the duration of overshoot (interval), and the scattering.

Next, out of the 300 oscillograms only those were selected which had at least one overshoot with a duration of $\tau \approx 3$ on the $a = \sigma$ level (the total number of such overshoots was 333). These oscillograms were then processed to find the distribution of the T interval following solely the overshoot of fixed duration, $\tau = 3$. In other words, the nominal function of probability density $q(T/\tau) = q(T/3)$ was being

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Table 1

Distribution of Overshoots of Fluctuation Envelope by Duration τ

and by the Intervals T between Overshoots

τ	$a=0$		$a=\sigma$		$a=1,5\sigma$		$a=-\sigma$		T	$a=0$		$a=\sigma$		$a=1,5\sigma$		$a=-\sigma$	
	n_i	$\frac{n_i}{N}$ %	n_i	$\frac{n_i}{N}$ %	n_i	$\frac{n_i}{N}$ %	n_i	$\frac{n_i}{N}$ %		n_i	$\frac{n_i}{N}$ %	n_i	$\frac{n_i}{N}$ %	n_i	$\frac{n_i}{N}$ %	n_i	$\frac{n_i}{N}$ %
1	143	6,4	406	25,7	190	27,2	647	39,5	1	144	7,0	19	1,3	4	0,75	60	3,46
2	233	10,4	369	23,4	233	33,4	460	28,1	2	308	14,9	61	4,3	6	1,13	85	4,90
3	347	15,6	333	21,1	170	24,4	220	13,4	3	356	17,3	89	6,3	12	2,26	105	6,06
4	311	13,9	181	11,5	68	9,7	112	6,8	4	266	12,9	105	7,4	13	2,45	130	7,50
5	452	20,3	133	8,4	19	2,7	112	6,8	5	262	12,7	121	8,6	21	3,96	164	9,46
6	197	8,8	77	4,9	8	1,1	37	2,3	6	165	8,0	85	6,0	20	3,77	132	7,62
7	171	7,7	36	2,3	6	0,9	25	1,5	7	145	7,0	107	7,6	23	4,34	136	7,85
8	113	5,1	16	1,0	1	0,1	13	0,8	8	108	5,2	74	5,3	27	5,09	126	7,27
9	63	2,8	15	1,0	2	0,3	6	0,4	9	70	3,4	64	4,6	17	3,21	90	5,19
10	73	3,3	9	0,6	1	0,1	5	0,3	10	83	4,0	86	6,1	18	3,40	130	7,50
11	30	1,3	3	0,2	—	—	—	—	11	50	2,4	60	4,2	23	4,34	63	3,64
12	37	1,7	1	0,1	—	—	—	—	12	36	1,8	69	4,9	24	4,53	88	5,08
13	17	0,8	—	—	—	—	—	—	13	20	1,0	59	4,2	14	2,64	49	2,83
14	15	0,7	—	—	—	—	—	—	14	15	0,7	48	3,4	22	4,15	42	2,42
15	10	0,4	—	—	—	—	—	—	15	22	1,1	43	3,0	20	3,77	45	2,60
16	7	0,3	—	—	—	—	—	—	16	9	0,4	36	2,6	21	3,96	42	2,42
17	4	0,2	—	—	—	—	—	—	17	3	0,2	46	3,3	20	3,77	36	2,08
18	5	0,2	—	—	—	—	—	—	18	—	—	25	1,8	16	3,02	32	1,85
19	2	0,1	—	—	—	—	—	—	19	—	—	16	1,1	15	2,83	26	1,50
	$N = 2231$		1579		698		1637			$N = 2062$		1410		530		1733	
	$\bar{\tau} = 5,11$		2,94		2,38		2,35			$\bar{T} = 5,1$		11,1		17,8		9,3	
	$\sigma_\tau^2 = 7,46$		3,56		1,69		2,70			$\sigma_T^2 = 10,4$		62,7		123,5		40,1	

Table 2

T	1	2	3	4	5	6	7	8	9	10	
$\frac{n_i}{N}$ %	1,3	4,3	6,3	7,4	8,6	6,0	7,6	5,3	4,6	6,1	$N = 1410$
$\frac{m_i}{M}$ %	2,3	5,0	6,5	6,4	9,2	7,3	8,4	6,9	5,0	6,5	$M = 261$

(cont'd)

T	11	12	13	14	15	16	17	18	19	
$\frac{n_i}{N}$ %	4,2	4,9	4,2	3,4	3,0	2,6	3,3	1,8	1,1	$N = 1410$
$\frac{m_i}{M}$ %	5,7	5,4	5,0	3,8	3,4	2,7	4,6	2,3	1,5	$M = 261$

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experimentally determined.

For comparison, Table 2 gives two laws for the distribution of intervals between overshoots $q(T) = \frac{n_i}{N}$ and $q(T/3) = \frac{m_i}{M}$ at $a = \sigma$.

A survey of the above tabulated data indicates that $q(T)$ approximately coincides in character with $q(T/3)$. This proves that the overshoots of the Rayleigh fluctuations on the $a \geqslant \sigma$ level can be also regarded as approximately noncorrelated.

Let us note that the probability densities $p(\tau)$ and $q(T)$ for the durations of fluctuation overshoots and of the intervals between them are satisfactorily approximated by the so-called gamma distribution (Bibl.4):

$$W_{\beta,\gamma}(\zeta) = \begin{cases} \frac{1}{\Gamma(\beta+1)\gamma^{\beta+1}} \zeta^{\beta} \exp\left(-\frac{1}{\gamma}\zeta\right) & \text{at} \quad \zeta \geqslant 0, \\ 0 & \text{at} \quad \zeta < 0, \end{cases} \qquad (7)$$

whose parameters β and γ are expressed by the mean value m and by the scattering σ^2 of the parameter ζ according to the formulas:

$$\gamma(\beta+1) = m, \quad \gamma^2(\beta+1) = \sigma^2. \qquad (8)$$

For example, with respect to the data in Table 1, pertaining to $a = \sigma$, we have

$$p(\tau) = W_{\beta_1,\gamma_1}(\tau), \quad \beta_1 = 1{,}43, \quad \gamma_1 = 1{,}2;$$

$$q(T) = W_{\beta_2,\gamma_2}(T), \quad \beta_2 = 1{,}0; \quad \gamma_2 = 5{,}65.$$

In conclusion, we will demonstrate that a practically important inference can be made (Bibl.5) from the fact of the noncorrelation of fluctuation overshoots at $a \geqslant \sigma$. Let the mean number of overshoots during unit time on the level a be equal to m. Then, the probability P(m) that m overshoots will occur during a time interval t will be determined by Poisson's law:

$$P(m) = \frac{(nt)^m}{m!} e^{-nt}. \qquad (9)$$

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REACTIVE TRIGGER CIRCUITS: THEIR OPERATIONAL ANALYSIS AND COMPUTING METHODS

by

A.S.Vladimirov

The article gives the results of a theoretical analysis of a reactive trigger circuit with plate coupling. This analysis is used as the basis for recommendations as to the selection of parameters ensuring the most stable mode of performance of the circuit. Further, methods of engineering computations are presented, together with an example of computing a circuit by these methods.

Introduction

A reactive trigger circuit* permits under known conditions the formation of sharply varying voltages at known conditions.

However, the reactive trigger circuit has only one steady state. An external signal causing the circuit to depart from the steady state also causes a sharp change in its mode of operation. After removal of the control voltage, the circuit abandons its new state and returns to its original (steady) state, on its own and not owing to another external pulse. The duration of this return depends on the circuit parameters. Thus, when the circuit is actuated by an external control signal, it may form output pulses of a definite duration.

Reactive trigger circuits and its modifications have become widely used in various radio-engineering devices. They are utilized in many radar equipment units,

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*This circuit is also known as: reactive trigger, retarded multivibrator, single-cycle relaxator.

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multiplex telephone and television systems, electronic computers, and various special control-and-measuring pulse devices.

Reactive trigger circuits, like the entire class of electronic relaxation circuits, are systems with a sharply expressed nonlinearity, which causes known difficulties in the development of mathematically rigorous general methods of analysis of such circuits.

In many of the published works concerned with the analysis of reactive trigger circuits the authors, intending to simplify the problem, compute the basic parameters on proceeding from the assumption of the presence of some given operating condition in the circuit, although that mode is not theoretically justified a priori and its assumed existence tends to impose specific conditions on the relations among the parameters: the authors examine the processes theoretically without considering the nonlinearity of tube characteristics. This leads to definite inaccuracies in the conclusions and in the computation results.

A general mathematical theory of the operation of this class of circuits, taking the nonlinearity of tube characteristics into consideration, has been evolved by Academician Andronov and his school. The rigorous theoretical analysis of a rheostatic trigger circuit and of a number of relaxation systems, conducted by Academician A.A.Andronov and Professor S.E.Khaykin (Bibl.1) has made it possible to determine the character of equilibrium states in the circuits and to examine the peculiarities of behavior of the systems when in the phase plane. These investigations served to formulate a number of important qualitative conclusions. Subsequent studies revealed that these conclusions can be successfully used as the foundation for formulating basic qualitative relationships. Thus, on the basis of this regorous method and upon considering the conditions encountered in practice, the theory of the symmetrical rheostatic trigger circuit has been developed to the stage where basic qualitative relationship and methods of the related engineering computations were obtained (Bibl.2, 3).

The methods of analysis indicated in the given paper are applied to a reactive trigger circuit, and the methods of the engineering computation of that circuit are proposed on the basis of the conducted theoretical investigation. It is to be noted in advance that these data are far from treating exhaustively the problem of the analysis of such circuits. In particular, the present paper deals with the theoretically most simply analyzable and rather widely applied plate-coupling version of the circuit: the transient processes in the circuit are not investigated (spurious parameters are disregarded). However, the conducted analysis and the recommended computing methods are satisfactorily corroborated in practice in computing of circuits forming pulses of a duration noticeably longer than the transit time in the circuit (up to 3 - 5 μ sec) and have proved useful in solving a number of practical tasks.

There is some basis for assuming that the methods proposed here can be applied also to other versions of the reactive trigger circuit.

Theoretical Analysis of Circuit Performance in the Absence of Control Signal Action

The reactive trigger circuit analyzed here is depicted in Fig.1.

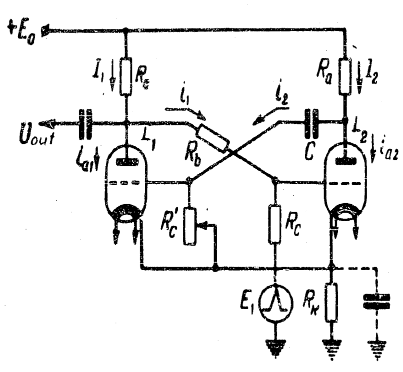

Fig.1

Let us begin our investigations of the processes in the circuit with the case of the absence of an external signal E_1.

Let us compose differential equations of the basic links of this circuit.

To simplify the analysis we will disregard the plate reaction and the grid currents of the tubes in the circuit.

Assuming that the resistance R_k is under the influence of a direct voltage E_k equal to the grid bias voltage of the tube L_2, the following equations can be written:

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$$R_a I_2 + \frac{1}{C}\int i_2 dt + i_2 R_c' + E_k = E_a; \qquad (1)$$

$$R_a I_1 + R_b i_1 + R_c i_1 = E_a. \qquad (2)$$

Considering that

$$I_2 = i_2 + i_{a2} \text{ and } I_1 = i_1 + i_{a1},$$

where

$$i_{a1} = \varphi_1 (i_2 R_c'), \text{ while } i_{a2} = \varphi_2 (i_1 R_c),$$

after the necessary transformations we obtain

$$(R_a + R_c') \frac{di_2}{dt} + R_a R_c \varphi_2' (i_1 R_c) \frac{di_1}{dt} + \frac{i_2}{C} = 0, \qquad (3)$$

$$R_\Sigma \frac{di_1}{dt} + R_a R_c' \varphi_1' (i_2 R_c') \frac{di_2}{dt} = 0. \qquad (4)$$

Here

$$R_\Sigma = R_a + R_b + R_c.$$

On determining $\frac{di_1}{dt}$ from eq.(4) and substituting into eq.(3), we will have

$$\frac{di_1}{dt} = \frac{-i_2}{C\left[(R_a + R_c') - \frac{R_a^2 R_c R_c'}{R_\Sigma} \varphi_1' (i_2 R_c') \varphi_2' (i_1 R_c)\right]}. \qquad (5)$$

Assuming that the tube characteristics have the form of curves with a steepness that decreases on both sides of the zero grid voltage, which sufficiently well coincides with the real tube characteristics at operation of the circuit in question (Bibl.3, 4), let us find the equilibrium states in the system and determine their stability. For this purpose we will use eq.(5).

As can be concluded from eq.(5), an equilibrium state in the circuit $(\frac{di_2}{dt} = 0)$ will occur at $i_2 = 0$.

The equilibrium state will be stable if the current i_2 and $\frac{di_2}{dt}$ will have different signs at a change in i_2 near the equilibrium state (Bibl.1).

The circuit in question will operate as a circuit for the formation of pulses only in the presence of a stable equilibrium. In the opposite case, the circuit becomes a self-oscillatory one.

The requirement for equivalence of $\frac{di_2}{dt}$ and i_2 is met when

$$R_a + R_c' > \frac{R_a^2 R_c R_c'}{R_B} \varphi_1'(i_2 R_c') \varphi_2'(i_1 R_c). \qquad (6)$$

The obtained formula determines the required mode of the circuit tubes.

The equilibrium state $i_2 = 0$ corresponds to the absence of a voltage drop in the resistor R_c', i.e., to a charged capacitor C. The plate current of the tube L_1 is determined by the zero grid bias. This makes it possible to find the transconductance of the plate current of the tube L_1 at the quiescent point.

The quiescent point on the characteristic of tube L_2 should be so selected as to satisfy eq.(6), i.e.,

$$S_2 < \frac{(R_a + R_c') R_B}{R_a^2 R_c R_c' S_1}, \qquad (7)$$

where S_1 and S_2 are the transconductance of the tubes L_1 and L_2, respectively, at the quiescent points corresponding to a stable equilibrium state.

Equation (7), which ensures the absence of a self-oscillatory process in the circuit, is not a reliable guide to the selection of the statistical mode of the circuit. If the circuit is to have a mode of operation ensuring an avalanche-like shift of currents (trigger mode), it is necessary that an unstable state be created in the circuit while it is acted upon by an external exciting pulse. Accordingly, at the selected circuit elements, the effect of the signal should ensure such variations in the transconductances of the tubes L_1 and L_2 as to cause i_2 and $\frac{di_2}{dt}$ to become equivalent.

As can be concluded from eq.(5), the following formula should be satisfied in the critical case:

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$$\frac{R_a^2 R_c R_c' \varphi_1'(i_{2k} R_c') \varphi_2'(i_{1k} R_c)}{(R_a + R_{\Sigma bc})(R_a + R_c')} = 1,$$

To ensure a reliable mode of formation (avalanche-type process), the parameters of the circuit should be selected on the condition of unconditional fulfillment of the following equality:

$$\frac{R_a^2 R_c R_c' S_1^* S_2^*}{(R_a + R_{\Sigma bc})(R_a + R_c')} = K_y, \qquad (8)$$

where $K_y > 1$.

Here S_1^* and S_2^* are the transconductances of the linear segments of tube characteristics in the circuit

$$R_{\Sigma bc} = R_b + R_c.$$

By analogy with the rheostatic trigger circuit, let us term the coefficient K_y the stability factor of the trigger circuit. Let us note that here the value of K_y, as in a rheostatic trigger circuit, characterizes the extent to which the total gain of the circuit has to exceed unity (Bibl.3). However, while the coefficient K_y in a rheostatic trigger circuit determines the size of the area of existence of stable states of equilibrium, i.e., the stability of the circuit when acted upon by spurious signals and the instabilities of its parameters, in a reactive trigger circuit the stability depends on the selected statistical mode, whose boundary of existence is given by satisfying eq.(7). Further, the selection of the value of the coefficient K_y in the circuit in question greatly influences the process of its triggering by determining the degree of stability of the fixation of the leading edges of the formed pulses. As will be shown below, the value of the stability factor of the trigger circuit K_y should be selected within the limits of $K_y = 5 - 15$.

Control of Circuit by External Pulsed Signals

Let us examine the processes occurring in the circuit when acted upon by ex-

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ternal control signals. In the circuit shown in Fig.1, the controlling emf E_1 is connected in series with the resistance R_c in the circuit of the control grid of the tube L_2.

On introducing the following designations:

$$U_{g1} = i_2 R'_c, \tag{9}$$

$$U_{g2} = i_1 R_c + E_1 - E_k, \tag{10}$$

where U_{g1} and U_{g2} are the voltages active on the control grids of the tubes L_1 and L_2, while E_1 is the amplitude of the control pulse, and considering that, according to eqs.(9) and (10),

$$i_2 = \frac{U_{g1}}{R'_c},$$

$$i_1 = \frac{U_{g2} + E_k - E_1}{R_c},$$

we obtain the following initial relations:

$$R_a \frac{U_{g1}}{R'_c} + R_a \varphi_2(U_{g2}) + \frac{1}{C}\int \frac{U_{g1}}{R'_c}\,dt + R'_c \frac{U_{g1}}{R'_c} + E_k = E_a, \tag{11}$$

$$R_a \frac{(U_{g2} + E_k - E_1)}{R_c} + R_a \varphi_1(U_{g1}) + R_b \frac{(U_{g2} + E_k - E_1)}{R_c} + + R_c \frac{(U_{g2} + E_k - E_1)}{R_c} + E_1 = E_a. \tag{12}$$

After simple algebraic transformations we have:

$$(R_a + R'_c)\,U_{g1} + R_a R'_c \varphi_2(U_{g2}) + \frac{1}{C}\int U_{g1}\,dt + R'_c E_k = E_a R'_c, \tag{13}$$

$$R_\Sigma U_{g2} + R_a R_c \varphi_1(U_{g1}) + R_\Sigma E_k - (R_a + R_b)\,E_1 = E_a R_c. \tag{14}$$

From these equations, we obtain

$$\frac{dU_{g1}}{dt} = \frac{-U_{g1}}{C\left[(R_a + R'_c) - \dfrac{R_a^2 R_c R'_c}{R_\Sigma}\,\varphi'_1(U_{g1})\,\varphi'_2(U_{g2})\right]}. \tag{15}$$

Equation (15) is analogous to eq.(5) which was obtained for the statistical mode; this means the circuit will be in an equilibrium state at $U_{g1} = 0$, and in the event of a proper selection of parameters [see eq.(7)] this state will be stable. When the control pulse acts on the circuit, the grid bias voltages, and consequently the transconductances of the tubes L_1 and L_2 at the quiescent points, change in value. At a specific, critical value of a positive external pulse acting, for example, on the grid of the tube L_2, the denominator of eq.(15) becomes equal to zero; at a slightly higher value of the pulse ($E_1 > E_{1cr}$) the sign of the denominator will be altered and, on the condition that eq.(8) is satisfied, the equilibrium state of the circuit will prove unstable - a sharp change in the operating conditions of the circuit will occur: tube L_1 which previously admitted plate current will be blocked, and tube L_2 will open.

Thus, for an analysis of the states of the circuit, it is necessary to examine the behavior of the denominator of eq.(15). It can be demonstrated that the grid voltage of the tube L_1 is in a definite relationship to the grid voltage of the tube L_2 and that the denominator of eq.(15) is a function of a single variable. Therefore, to facilitate the analysis of eq.(15) let us establish a relationship between changes in U_{g1} and changes in U_{g2}.

On making use of eq.(13) and assuming that the capacitor voltage

$$U_c = \frac{1}{C} \int \frac{U_{g1}}{R'_c} dt$$

remains unchanged during the action of the momentary external pulse, we obtain

$$U_{g1} = -\frac{R_a R'_c \varphi_2(U_{g2})}{R_a + R'_c}. \qquad (16)$$

Then, eq.(15) will assume the form

$$\frac{dU_{g1}}{dt} = \frac{-U_{g1}}{C\left\{(R_a + R'_c) - \frac{R_a^2 R_c R'_c}{R_{\text{п}}} \varphi'_1 \left[-\frac{R_a R'_c}{R_a + R'_c} \varphi_2(U_{g2})\right] \varphi'_2(U_{g2})\right\}}. \qquad (17)$$

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Let us determine the critical value of the exciting pulse E_{1cr} at which the circuit will not yet operate.

The external emf $E_1 = E_{1cr}$, is correlated by equating the denominator of eq.(17) to zero, i.e., the following formula should hold:

$$\frac{R_a R_c'}{R_a + R_c'} \varphi_2' (U_{g2k1}) = \frac{1}{\frac{R_a R_c}{R_\Sigma} \varphi_1' \left[- \frac{R_a R_c'}{R_a + R_c'} \varphi_2 (U_{g2k1}) \right]}. \qquad (18)$$

Here $U_{g2} = U_{g2kl}$ is the grid voltage of the tube L_2 corresponding to the critical value of the external pulse*.

Inasmuch as voltage U_{g2kl} determines the value of E_{1cr}, it is necessary to derive eq.(18) in relation to U_{g2kl} in order to determine the value of the excitation pulse of the circuit.

At the given tube characteristics, this equation is conveniently derived by graphical means. During the plotting of the graphs of the functions

$$\frac{R_a R_c'}{R_a + R_c'} \varphi_2' (U_{g2}) = \psi_1 (U_{g2})$$

and

$$\frac{1}{\frac{R_a R_c}{R_\Sigma} \varphi_1' \left[- \frac{R_a R_c'}{R_a + R_c'} \varphi_2 (U_{g2}) \right]} = \psi_2 (U_{g2})$$

it is necessary to take account of eq.(7), i.e., to keep in mind that, in a statistical state, the tube L_1 is open and the tube L_2 is closed.

Figure 2 depicts the graphical derivation of eq.(18) for a circuit with tubes of the 6N15P and 6N8 type, with circuit parameters corresponding to the following stability factors:

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*Here the grid voltage of the tube L_1 will be correspondingly determined by the formula

$$U_{g1} = U_{g1k1}.$$

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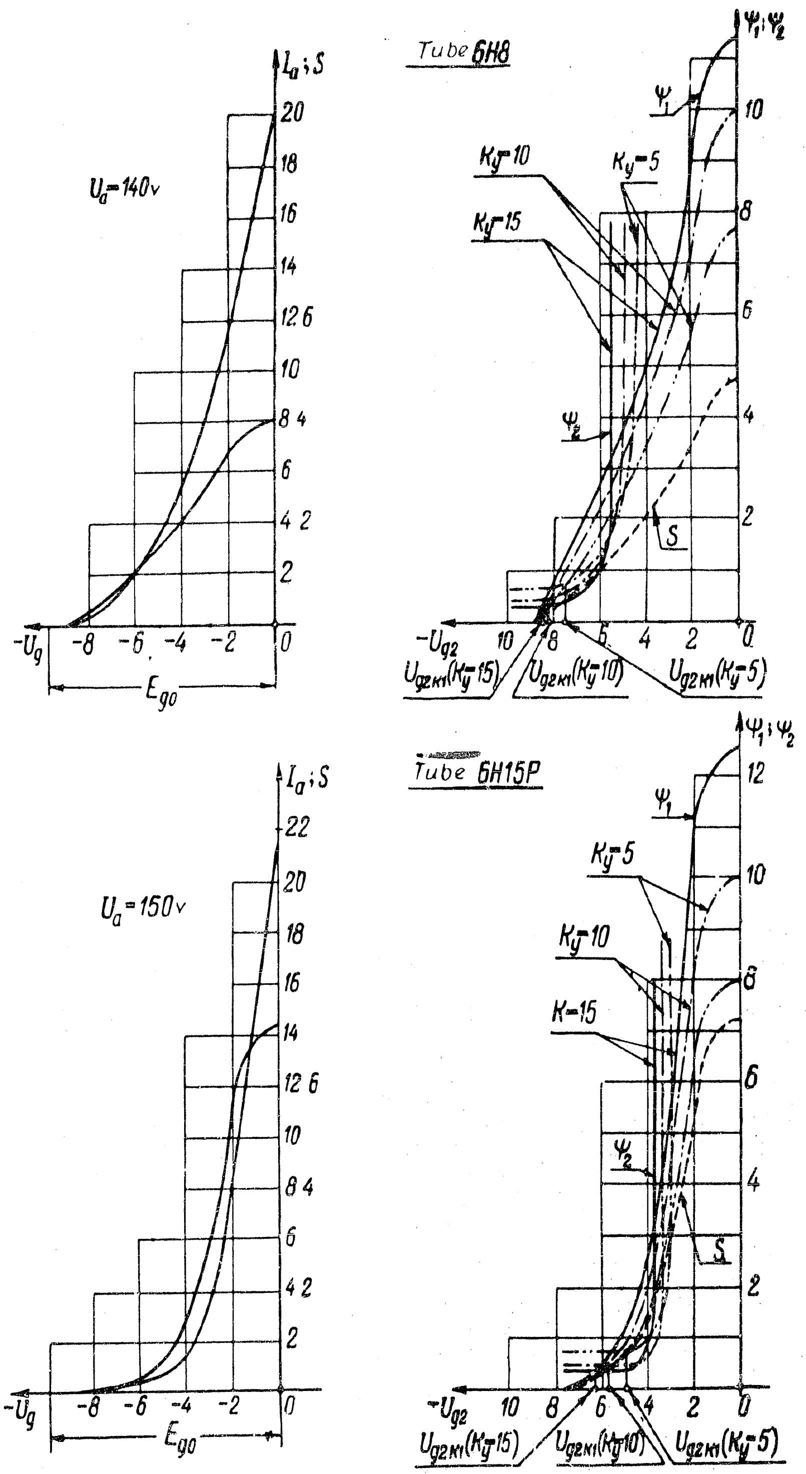
Tube 6H8
I_a; S
U_a=140v
$-U_g$
E_{go}
Ψ_1; Ψ_2
Ψ_1
Ψ_2
K_y=10
K_y=5
K_y=15
S
$-U_{g2}$
$U_{g2к1}(K_y=15)$
$U_{g2к1}(K_y=10)$
$U_{g2к1}(K_y=5)$
Tube 6H15P
U_a=150v
K=15

Fig.2

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$$K_y = 5, \quad K_y = 10, \quad K_y = 15.$$

The plotting of the curves $\psi_1(U_{g2})$ and $\psi_2(U_{g2})$ was conducted with a preliminary use of the graphs of the grid and transconductance characteristics for various values of grid voltage. The intersection points of the curves $\psi_1(U_{g2})$ and $\psi_2(U_{g2})$, determining $U_{g2} = U_{g2k1}$, are the derivatives of eq.(18).

To determine E_{1cr} we will use eq.(14) which, in the given case, should be written as follows;

$$U_{g2k1} = \frac{E_a R_c}{R_\Sigma} - E_k - \frac{R_a R_c}{R_\Sigma}\varphi_1(U_{g1k1}) + \frac{R_a + R_b}{R_\Sigma}E_{1\,cr}.$$

On designating

$$\frac{E_a R_c}{R_\Sigma} - E_k - \frac{R_a R_c}{R_\Sigma}\varphi_1(U_{g1k1}) = E_{g2k1}, \tag{19}$$

where E_{g2k1} is the resulting control-grid bias corresponding to the operating mode of the circuit, we obtain

$$U_{g2k1} = E_{g2k1} + \frac{R_a + R_b}{R_\Sigma}E_{1\,cr},$$

Consequently,

$$E_{1\,cr} = \frac{(U_{g2k1} - E_{g2k1})R_\Sigma}{(R_a + R_b)}. \tag{20}$$

It is of interest to analyze the character of the curves determining the derivation of eq.(18). As indicated by the curves ψ_1 and ψ_2, their points of intersection vary in position at various parameters of the circuit. When the circuit is based on parameters determined by low values of the stability factor K_y, the intersection of the curves takes place at high transconductance of the tube L_2 (at low negative values of U_{g2}), i.e., the triggering of the system occurs at relatively high control signals. At low values of K_y, the intersection point is displaced into the area with lower values of transconductance – the circuit is controlled by smaller exciting pulses. Thus, the level of the threshold of operation of the circuit depends both on the form of the characteristics of the tubes used and on the circuit parameters selected.

An examination of Fig.3, which shows the area of derivation of eq.(18) on a rel-

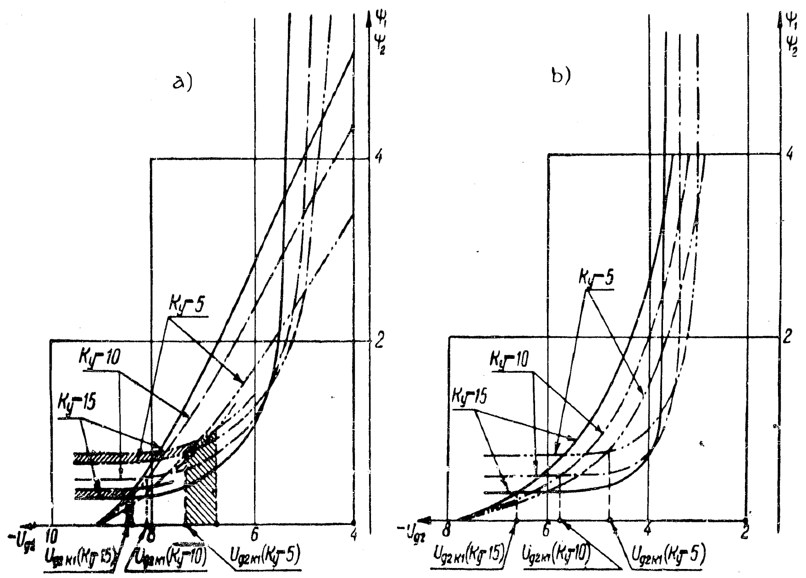

Fig.3

a) Tube 6N8; b) Tube 6N15P

atively larger scale, indicates that the instability of the circuit parameters (the cross-hatched area in the drawing refers to the range of variations in the operational threshold at a 15% change in resistance R_a) at low K_y values leads to a substantial (compared with high K_y values) change in the operational threshold. The same conclusion can be drawn upon investigating the influence of the shift in the tube characteristics and of the change in transconductance, due to aging and replacement of the tubes and also due to the background noise of the alternating current of the power supply. Therefore, considering the finiteness of the steepness of the leading edge of the starting pulse and, especially, of the discharge curve of the capacitor C, it can be concluded that the mode of operation at a low K_y is not

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suitable from the viewpoint of ensuring the stability of the leading edge of the formed pulse and the reliability of operation of the circuit. It is to be noted, however, that at extremely high K_y the circuit begins to operate in the extreme non-linear segment of the response curve, which is not sufficiently stable.

A comparison of the curves in Fig.3, plotted for tubes of the 6N8 and 6N15P types, indicates that the presence of an extensive sloping segment ("tail") in the response curve of the tube 6N15P also leads to an extension of the operational threshold of the circuit, and causes the circuit to be noticeably affected by changes in parameters and by replacement of tubes and, finally, causes the fixation of pulse leading edges to be less explicit.

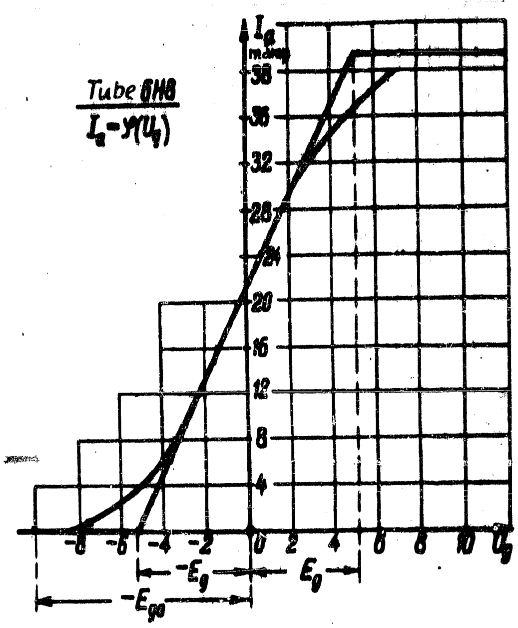

Fig.4

Therefore, in practical computations of the circuit, it is necessary to select a value of K_y ranging from 5 to 15 and to avoid using tubes with tailed grid characteristics.

In order to compute the circuit elements it is also necessary to define the selection of the parameters of R_c' and C, determining the duration of the pulses shaped by the circuit. As indicated by experiments, an accuracy fully sufficient in practice is obtained when using the formulas furnished by N.V.Semakov (Bibl.5). At a duration of the shaped pulse τ_{pulse} occasioned by the operating conditions, and at a selected capacitance C, the resistance R_c' is determined by the following formula:

$$R_c' = \frac{\tau_{pulse}}{2{,}3\, C \lg \frac{U_{out}}{U_{glk}}}. \tag{21}$$

Here U_{g1k} is the grid voltage of the tube L_1 at the instant of flip-flop of the circuit during the discharge of the capacitor C;

U_{out} is the amplitude of the pulse formed by the circuit.

The above analysis of the trigger circuit makes it possible to compute its parameters and to determine the pulse value ensuring its operation; however, for this purpose it is necessary to conduct certain plotting operations. This involves known inconveniences when the above formulas are used in engineering practice and thus necessitates a search for reasonably simplified computing methods that would not violate the rigorousness of the conclusion made. One of the ways to such a simplification, which has been fairly widely adopted (Bibl.3, 6) is the substitution of a real grid characteristic by a rectified characteristic consisting of three rectilinear segments. Figure 4 depicts a real and a rectified characteristic of a 6N8 type tube.

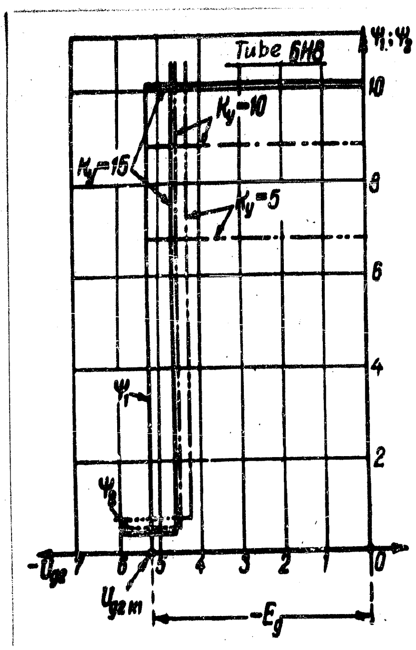

Fig.5

A derivation of eq.(18) for rectified tube-grid characteristics is depicted in Fig.5. In the given case, we have

$$\dot U_{g2k1} = |E_g|,$$

so that eq.(19) assumes the following form:

$$\frac{E_a R_c}{R_b} - \frac{R_a R_c}{R_b}\varphi_1(0) - E_k = E_{g0},$$

where E_{go} is the resultant initial control-grid bias of the tube L_2.

Therefore,

$$E_{1\,cr} = \frac{(E_g - E_{g0})\,R_b}{R_a + R_b}. \tag{22}$$

in this formula the values of E_g and E_{go} should be substituted together with the corresponding signs.

Equation (21) for computing the resistance R_c' assumes the following form:

$$R_c' = \frac{\tau_{pulse}}{2{,}3C \lg \frac{U_{out}}{E_g}}. \qquad (23)$$

Methods of Computing a Reactive Trigger Circuit

In practical applications of reactive trigger circuits it is usually necessary to so select the circuit tubes and parameters that the circuit will ensure the shaping of pulses of definite duration and of the required amplitude. Moreover, it is necessary to know the value of the external control pulse effecting a reliable operation of the circuit.

The following data are usually given: the necessary value of the output pulse voltage U_{out}; the duration of the shaped pulses τ_{pulse}; and the voltage of the plate power supply E_a.

1. Selection of Tubes and of Their Operating Conditions

The selection of the tube type to be used in the circuit is largely determined by a given amplitude of the pulse formed. In order to obtain output pulses of large amplitude under given conditions of tube operation, and to insure stability of leading edges of the shaped pulses (with abruptly discontinuing characteristics), it is necessary to select tubes with relatively high values of plate current at zero grid bias. Considerations of compactness dictate the use of dual tubes. The tubes most suitable for pulse-shaping networks generating pulses of short duration are preferably pentodes with a high transconductance and low capacitances.

Let us determine the mode of operation of the network tubes.

As can be seen from Fig.1, the value of the plate voltage for an open tube L_1 will be determined by the equality

$$U_a = E_a - \Delta E_a - E_k - U_{out} \qquad (24)$$

At values of U_{out} and E_a known from the given conditions of the problem, it is merely necessary to study the ΔE_a, which is the voltage drop at the plate load from the bleeder current R_b, R_c, and the voltage acting on the cathode resistance E_k.

To eliminate the excess voltage drop at the plate load from the bleeder current, ΔE_a is selected within the range of

$$\Delta E_a = (0,05 \div 0,15)\, E_a.$$

The voltage drop at the cathode resistance usually equals

$$E_k = 30 \div 60 \text{ v}.$$

In accordance with the determined U_a, the grid characteristic of the tube is rectified, and the value of the cutoff voltage E_g is determined.

As can be concluded from eq.(7), at zero grid bias of the tube L_1, the tube L_2 should be blocked by the bias taken from the resistance R_k. For this purpose, the initial grid bias voltage E_{go} of the tube L_2 should be 15 - 20% higher than the voltage corresponding to the blocking of the tube.

2. Computing the Circuit Parameters

a) The computation of circuit parameters is conveniently begun by selecting the parameters determining the duration of the shaped pulses. As pointed out in the literature (Bibl.7), for eliminating any deviation of the shape of these pulses from the rectangular, the capacitance C should be of the low order of 50 - 100 μμf, while the resistance R_c' should be, as in ordinary amplifying circuits, no greater than 1 - 1.5 megohms. In the case of the shaping of pulses of considerable duration, the capacitance C should be taken at 100 and more micromicrofarads.

The resistance R_c' is computed by eq.(23).

Considering the operating conditions of the tubes of a reactive trigger circuit

$$R_a = \frac{U_{out}}{I_{a0}}, \qquad (25)$$

where I_{ao} is the plate current of the tube at zero grid bias and at the plate voltage determined from eq.(24).

Further we determine the values of the bleeders R_b and R_c. As it can be concluded from Fig.1,

$$R_{\Sigma bc} = \frac{R_a (E_a - \Delta E_a)}{\Delta E_a}. \qquad (26)$$

On the basis of eq.(8) we compute the minimum allowable value of the resistance R_c, ensuring the trigger mode at a given K_y:

$$R_{c\,min} = \frac{R_\Sigma (R_a + R_c') K_y}{R_a^2 R_c' S^2}. \qquad (27)$$

Then,

$$R_b = R_{\Sigma bc} - R_c.$$

Here S is the transconductance of the rectified tube characteristic.

c) Next we compute the cathode resistor R_k.

The voltage E_k taken from this resistor and blocking the tube L_2 when in the static state, equals

$$E_k = E_{g0} - E_{g+}.$$

The positive polarity voltage E_{g+}, supplied to the grid of the tube L_2 by the divider, is determined by the explicit formula

$$E_{g+} = \frac{E_a - \Delta E_a - U_{out}}{R_{\Sigma bc}} R_c. \qquad (28)$$

On determining E_k, we find that

$$R_k = \frac{E_k}{I_{a0}}.$$

In concluding, we compute the value of the control pulse.

The critical value of the signal is computed from eq.(22). For reliable operation of the circuit, the obtained value of the control signal should be increased by 15 - 20%.

Example of Computation

Compute a pulse-shaping network ensuring the shaping of output pulses with a duration of τ_{pulse} = 10 μsec and an amplitude of U_{out} = 70 v. The voltage of the plate supply is E_a = 250 v.

1. According to the conditions of the problem the network should shape pulses of low duration; therefore, we select a tube of the 6N1P type with small interelectrode capacitances.

Assuming that $\Delta E_a = 0.1E_a$ = 25 v and E_k = 40 v, we obtain

$$U_a = 115\ v.$$

On rectifying the grid characteristic of the tube, we have

$$E_g = -4\ v,\ I_{a0} = 5\ mamp,\ S = 3{,}2\ \frac{mamp}{v}.$$

2. Given a capacitance of the shaping network: C = 100 μμf. Then, according to eq.(23) we obtain $R_c = 35 \times 10^3$ ohms.

3. On the basis of eq.(25) we find

$$R_a = 4.7 \times 10^3 \text{ ohms. Let } R_a = 5.1 \times 10^3 \text{ ohms.}$$

4. Let us determine the values of the bleeders R_b and R_c. On using eq.(26), we obtain

$$R_{\Sigma bc} = 45 \cdot 10^3\ ohm.$$

Taking K_y = 15, we obtain $R_{c\ min} = 3.5 \times 10^3$ ohms.

Let us take $R_c = 5.1 \times 10^3$ ohms. We compute the resistance R_b and select R_b = $= 39 \times 10^3$ ohms.

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5. We compute the value of the cathode resistor R_k. From eq.(38) we find the preliminary value

$$E_{g+} = 17 \text{ v}.$$

Proceeding from the grid characteristic of the tube we select $E_{go} = 15$ v.

Then, $E_k = -32$ v and $R_k = 2.1 \times 10^3$ ohms.

6. Let us determine the value of the pulse which has to be fed to the input of the circuit in order to control the circuit.

According to eq.(22), we have

$$E_{1\,cr} = 12 \text{ v}.$$

For reliable operation of the circuit we will assume that

$$E_{excit.} = 15 \text{ v}.$$

Experimental investigations of circuits with parameters chosen in accordance with the above computations showed satisfactory agreement of experimental and calculated values.

In concluding, I wish to express my gratitude to S.V.Novakovskiy for the advice he gave while reviewing the manuscript.

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PROPAGATION OF METER AND DECIMETER WAVES OVER ROUGH SURFACE OF THE EARTH

by

K.D.Dymovich

The propagation of radio waves over uneven surfaces of the earth (in particular over a surface that is sinusoidal in one plane) is examined from the viewpoint of laws governing geometrical optics; a transcendental equation was derived whose graphical solution determines the coordinates of the reflection points. Further, the article furnishes a formula for determining the difference in the travel of direct and reflected rays. The limits of applicability of the proposed computing method are appraised. A comparison of the results of this method with the more rigorous L.M.Brekhovskikh diffraction method and with experimental data reveals satisfactory agreement.

The existing method for computing the field strengths of ultrashort radio waves, which presupposes an ideally smooth earth's surface, cannot always satisfy the designers of communication lines.

Most radio communication lines pass over terrain whose surface can in no way be considered ideally smooth. A comparison of the measured values of field strength with the analogous values computed for an ideally smooth surface indicates that these values diverge considerably, with the divergence sometimes reaching 500 to 1000% and more. Therefore, the possibility of computing the field strength by taking the terrain features into consideration is an absolutely necessary requirement.

An approximate method for computing field strengths in line-of-sight zones is presented below. This method makes it possible to take account of the terrain features and is useful for the range of meter and decimeter waves but does not apply to the centimeter-wave range because there, as is known, the reflection from the earth has a scattering character.

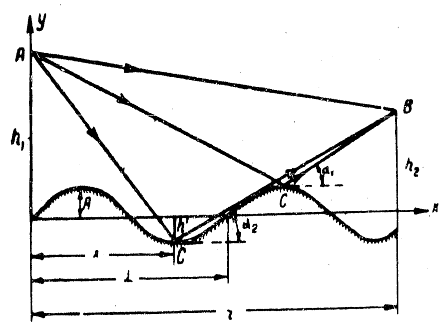

Fig.1

Let us assume that the profile of the earth's surface can be given by the following equation (see Fig.1):

$$y = f(x), \tag{1}$$

where x is distance along the path of wave propagation, as read from the origin of the coordinates. The transmitting antenna h_1 is located at the origin of the coordinates, while the receiving antenna h_2 is located at a distance r from the origin of the coordinates along the x-axis. To be determined is the field strength at the locus of the receiving antenna.

The coordinates of the reflection points can be located in the following manner: On determining the equations for the rays AC and CB as equations of straight lines intersecting a given point in a given direction

$$f(x) - h_1 = k_1 x \quad \text{for ray} \quad AC$$

and

$$f(x) - h_2 = k_2 (x - r) \quad \text{for ray} \quad CB,$$

where

$$k_1 = \operatorname{tg} \gamma_1$$
$$k_2 = \operatorname{tg} \gamma_2,$$

and considering that the angle of incidence at the reflection point equals the angle of reflection, i.e. (see Fig.2),

$$180 - \gamma_1 + \beta = \gamma_2 - \beta,$$

or

$$\operatorname{tg}[180 - (\gamma_1 - \beta)] = \operatorname{tg}(\gamma_2 - \beta), \tag{2}$$

it is possible to obtain an equation whose derivation will be furnished by the coordinates of the reflection point.

The angle β is the angle between the tangent and the x-axis at the reflection point

$$\operatorname{tg} \beta = f'(x). \tag{3}$$

The angles γ_1 and γ_2 are determined by the following formulas:

$$\operatorname{tg} \gamma_1 = \frac{f(x) - h_1}{x}, \tag{4}$$

and

$$\operatorname{tg} \gamma_2 = \frac{f(x) - h_2}{x - r}. \tag{5}$$

On substituting eqs.(3), (4), and (5) in eq.(2) and carrying out some simplifications [considering that $x \gg f(x) - h$], the following necessary equation can be

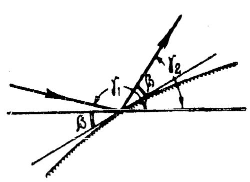

Fig.2

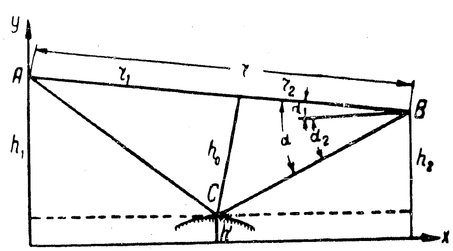

Fig.3

obtained:

$$(r - 2x) f(x) + (x - r) 2x f'(x) = r h_1 - (h_1 + h_2) x. \tag{6}$$

This transcendental equation is easily derived by graphical means.

The difference in travel between a direct ray and a ray reflected from the earth can be determined in the following manner (see Fig.3):

$$\Delta r = \sqrt{r_1^2 + h_0^2} + \sqrt{r_2^2 + h_0^2} - r \approx r_1 \left[1 + \frac{1}{2}\left(\frac{h_0}{r_1}\right)^2\right] + \\ + r_2 \left[1 + \frac{1}{2}\left(\frac{h_0}{r_2}\right)^2\right] - r = \frac{h_0^2 r}{2 r_1 r_2},$$

or, using the designations in Fig.1,

$$\Delta r \approx \frac{h_0^2 r}{2x(r - x)}. \tag{7}$$

Simple geometrical relationships (see Fig.3) can be used for readily obtaining

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where

$$h' = f(x).$$

It is known [cf. for example (Bibl.1, 2)] that, at reflection from the surface of a sphere, the formula determining the field strength of a ray reflected from the earth is complemented by the so-called "divergence" factor G, which determines the scatter of electromagnetic energy compared with reflection from a plane surface. In this case, the G factor for a spherical surface is determined by the following formula (cf. Bibl.2)

$$G^2 = \frac{1}{1 + \frac{2x(r - x)}{\rho r}}, \tag{9}$$

where ρ is the radius of curvature of the surface at the point of reflection.

It is obvious that an irregular convex surface will also scatter electromagnetic energy, so that the "divergence" factor should be included in the computations. This factor, at a sufficiently sloping surface, can be determined by eq.(9), and in this case ρ will have the following value:

$$\rho \approx \frac{1}{\frac{d^2 f(x)}{dx^2}}. \tag{10}$$

Such a representation serves also to determine the G factor for concavities, but then this will be a factor of "convergence" and not of "divergence".

It is known that the value of the field strength of a direct ray is determined as follows:

$$E = \frac{245\sqrt{PD}}{r_{km}} \cos \omega t \left(\frac{mv}{m}\right),$$

where P is the power radiated, in kw;

D is the directivity factor of the transmitting antenna compared with an isotropic radiator.

For the i^{th} reflected ray we have

$$E_i = \frac{245\sqrt{PD}}{r+\Delta r} R_i G_i \cos\left(\omega t - \theta_i - \frac{2\pi}{\lambda}\Delta r\right),$$

where θ_i is the angle of phase loss at the reflection of the i^{th} ray from the surface;

R_i is the reflection factor of the i^{th} ray;

λ is length of the radio wave

At horizontal polarization of the radio wave, the resultant field strength at the reception point will be determined by the following formula:

$$E = \frac{245\sqrt{PD}}{r}\left[\cos\omega t + \sum_{i=1}^{N} R_i G_i \cos\left(\omega t - \theta_i - \frac{2\pi}{\lambda}\cdot\Delta r_i\right)\right],$$

where N is number of reflected rays.

On determining the modulus of this equation and discarding the phase as of no interest, it is not difficult to obtain the following formula for determining the resultant field strength at the reception point

$$E = \frac{173\sqrt{PD}}{r}\sqrt{1 + 2\sum_{1}^{N} R_i G_i \cos\left(\theta_i + \frac{2\pi}{\lambda}\Delta r_i\right) + \left[\sum_{1}^{N} R_i G_i \cos\left(\theta_i + \frac{2\pi}{\lambda}\Delta r_i\right)\right]^2 + \left[\sum R_i G_i \sin\left(\theta_i + \frac{2\pi}{\lambda}\Delta r_i\right)\right]^2}. \quad (11)$$

For a single reflected ray, this formula becomes the ordinary reflection formula

$$E = \frac{173}{r}\sqrt{PD}\sqrt{1 + (RG)^2 + 2RG\cos\left(\theta + \frac{2\pi}{\lambda}\Delta r\right)};$$

For two reflected rays, the expression changes into another known (cf. Bibl.2) formula:

$$E = \frac{173}{r}\sqrt{PD}\sqrt{1 + (R_1G_1)^2 + (R_2G_2)^2 + 2R_1G_1\cos\left(\theta_1 + \frac{2\pi}{\lambda}\Delta r_1\right) + 2R_2G_2\cos\left(\theta_2 + \frac{2\pi}{\lambda}\Delta r_2\right) + 2R_1G_1R_2G_2\cos\left[(\theta_1 - \theta_2) + \frac{2\pi}{\lambda}(\Delta r_1 - \Delta r_2)\right]}.$$

All above statements apply to any rough sloping surface which can be approximated by the function f(x). In particular, this applies, for example, to the case of

$$f(x) = A \sin \omega x,$$

where A is amplitude of the irregularity;

$\omega = \frac{2\pi}{L}$;

L is the spatial wavelength of the local relief.

In this case, eqs.(6), (8) and (10) will correspondingly assume the following form:

$$(r - 2x)\sin \omega x + (x - r)\, 2 \cdot \omega x \cos \omega x = \frac{h_1 r - (h_1 + h_2)\, x}{A}, \qquad (6')$$

$$h_0 = (h_1 - h_2)\frac{r - x}{r} + h_2 - h', \qquad (8')$$

where $h' = A \sin \omega x$;

$$\rho = \frac{\omega^2}{(2\pi)^2}\, \frac{1}{A}\, \frac{1}{\sin \omega x}. \qquad (10')$$

The calculated formulas (7), (9) and (11) will remain unchanged.

With respect to the nature of this method, the following remark can be added:

It may happen that a reflected ray will be shielded by the crest of an irregularity and will thus, naturally, fail to create a field at the reception point. Therefore, it is always necessary to examine any reflected ray as to its "shadow", especially the rays reflected from concavities. The criterion for determining the "shadow" for a normal sinusoidal surface can be the following inequality (see Fig.1):

$$\operatorname{tg} \alpha_2 < \operatorname{tg} \alpha_1,$$

from which it obviously follows that

$$\frac{h_2 - h'}{r - x} < \frac{h_2 - A}{r - \frac{4n+1}{4} L},$$

where $n = \frac{x}{L} + \Delta l$;

Δl is a value below unity, complementing the fraction up to the first prime number.

The formulation and solution of this problem involved a number of assumptions which have to be justified.

First, it was assumed that the character of the local relief in a direction transverse to the path is of negligible importance. Generally speaking, this assertion is not correct. It is correct only when the change in local relief in that direction has a smooth character. Actually, as is known (see Bibl.1) the area important to reflection is constituted by ellipses extended along the path (along the x-axis); the dimensions of these ellipses in the transverse direction (along the y-axis) are hundreds and thousands of times smaller than their dimensions longitudinally to the path, insofar as meter and decimeter waves are concerned. Thus, the problem can be fairly correctly regarded as pertaining to a plane surface, if the changes of relief in the transverse direction are disregarded. Naturally, in the event of a sharp fluctuation in relief, for example, in a steep ravine etc., these considerations will cease to apply.

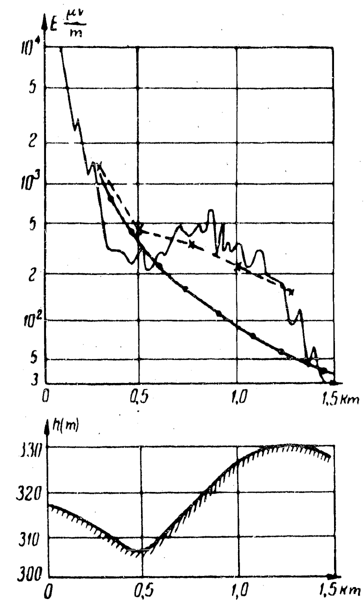

Fig.4

f = 95 mc; vertical polarization;

—— measured value of the field strength;

—●—●— value of the field strength computed by the quadratic formula on the assumption of a plane earth's surface;

●– – ● value of the field strength computed by the above method (earth's surface approximated by segment of sinusoid)

Secondly, it was assumed that reflection from a sufficiently slanted uneven surface proceeds similarly as from a plane surface of the earth, i.e., that the reflection point is extreme and that the angle of incidence at that point equals the angle of reflection. The correctness of this postulate is readily proved (see Bibl.1, pp.219 - 223), if the uneven surface is restricted by the following limitation which always holds true at a sufficiently slanted surface:

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$$\rho \ll \lambda,$$

$$\frac{4\pi}{\lambda} \rho \sqrt{|\varepsilon'|} \gg 1,$$

where ε' is the complex permittivity of the soil.

Third, the segment of uneven surface near the reflection point is replaced by a circle with a radius of ρ [see eqs.(10) and (10')]. Here the resulting error will be negligible, since such a replacement is made near the reflection point, where the greatest influence is exerted on the resultant value of the field strength; the influence of points spaced apart from the reflection point decreases at an increase in their distance from that point. Moreover, the changes in ρ do not greatly affect the field strength so that the value of ρ enters under the sign of the radical.

In the particular case when the dimensions of an uneven surface are considerably smaller than Fresnel's first zone computed for a plane reflecting surface, the results of the computation based on the proposed method compare favorably with the more rigorous diffraction method by L.M.Brekhovskikh (Bibl.3).

Table 1 gives the results of these computations for the case of $h_1 = h_2 = 100$ m, r = 30 km, and length of the uneven square D = 5 km. The obtained maximum deviation of 30% can be considered satisfactory.

Table 1

a)	b)		c)		d)	
A m	λ=0,6 m $E\,\frac{mv}{m}$	λ=6 m $E\,\frac{mv}{m}$	λ=0,6 m $E\,\frac{mv}{m}$	λ=6 m $E\,\frac{mv}{m}$	λ=0,6 m	λ=6 m
1	4,4	4,3	4,2	5,28	5	19
3	3,6	3,7	4,3	5,28	16	30
5	4,0	4,0	4,4	5,28	9	24
8	4,4	4,6	4,6	5,28	5	11
15	5,8	4,9	5,5	5,28	5	8
20	6,3	4,9	5,5	5,29	15	6
30	6,5	5,0	5,2	5,30	25	5

a) Amplitude of irregularity; b) Proposed method of computing; c) L.M.Brekhovskikh's method; d) Deviation in %

In concluding, let us compare the field strength values measured (Bibl.4) over a terrain whose relief is easily approximated by a sinusoid, with the analogous values computed by the above method. Satisfactory agreement is obtained (Fig.4) between the experimental curve and the theoretical curve, plotted on the assumption of a sinusoidal character of the surface. Figure 1 contrasts the experimental data with the computed values.

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CALCULATION OF MULTISTAGE AMPLIFIERS WITH JUNCTION TRANSISTOR

by

I.N.Migulin

On the basis of a generalization of the theory of transistor and vacuum-tube amplifiers, the article furnishes an analysis and computing methods for amplifier circuits with junction-type triodes.

The analysis and calculation of amplifying circuits with junction-transistors can be considerably simplified by using the general theory of vacuum-tube amplifiers. Upon selecting a set of static characteristics, an equivalent circuit and a system of parameters for the transistors, analogous to the theory of vacuum-tube amplifiers (Bibl.1, 2) it is possible to carry out calculations by using the computing methods employed for vacuum-tube circuits.

Figure 1 depicts an equivalent Pi-network for a semiconductor triode. Accordingly, its parameters (Bibl.2):

$$\left.\begin{aligned}
\dot{Y} &= \frac{1}{r_0}\,\frac{g r_0 + i\omega\tau}{1 + i\omega\tau}, \\
\dot{Y}_{fb} &= g_{fb}\,\frac{1 + i\omega\tau_{fb}}{1 + i\omega\tau}, \\
\dot{S} &= \frac{S_0}{1 + i\omega\tau}, \\
\dot{Y}_i &= (1 + \dot{S} r_0)\,\dot{Y}_{ok} + \dot{Y}_{ek}
\end{aligned}\right\} \qquad (1)$$

determine the amplifying properties at both high and low frequencies.

In most radio-engineering computations, which admit an error of several per-cent, the set of parameters (1) can be considerably simplified. In practice, the time constant τ_{fb} of the circuit of internal feedback through capacitance and con-ductance between collector and base, is of the same order as the time constant of the semiconductor triode τ. This fact permits the approximate assumption that and

$$\left.\begin{aligned} \dot{Y}_{fb} &\approx g_{fb} \\ \dot{Y}_i &\approx \frac{1}{R_i} + \frac{i\omega C_{ok} S_0 r_0}{1 + i\omega\tau} + i\omega C \end{aligned}\right| , \qquad (2)$$

where C_{ok} is the output collector capacitance of the triode

Taking account of the approximate equations (2), the complete set of equivalent parameters for the semiconductor triode includes four LF parameters g, g_{fb}, S_o, and R_i, a base resistance r_o, a time constant τ, and an output capacitance C_{ok}. The LF parameters are easily determined from static characteristic. The three other parameters are also determined with relative ease. The resistance r_o equals the input resistance at sufficiently high frequencies*. The time constant τ is determined by the frequency characteristic of the collector current during shorted operation at the output:

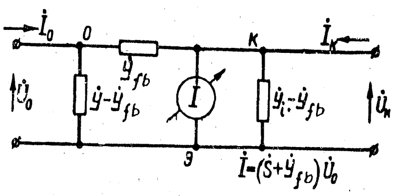

Fig.1

$$\frac{I_\kappa}{I_{\kappa 0}} = \frac{S}{S_0} = \frac{1}{\sqrt{1 + (\omega\tau)^2}} .$$

The output capacitance C is measured by conventional methods. The values of all these parameters for the Soviet-produced P1-series junction transistors are approximately as follows:

$$S_0 = 40 \div 150 \ ma/ \ ; \quad R_i = 20 \div 200 \ \text{kohm}$$
$$\frac{1}{g} = 200 \div 1000 \ \text{ohm}; \quad \frac{1}{g_{fb}} = 0{,}5 \div 5 \ \text{Mohm}$$
$$\tau = 0{,}5 \div 5 \ \mu sec \ ; \quad r_0 = 40 \div 200 \ \text{ohm};$$
$$C_{o\kappa} = 20 \div 40 \ \mu\mu f$$

Considering that the circuit in Fig.1 is analogous to the equivalent circuit of

*In practice, the resistance r_o, with an accuracy not less than 5%, can be measured at frequencies of

$$f = \frac{0{,}5 + 1}{\tau} .$$

a vacuum tube, it can be related to the basic formulas used in the theory of vacuum-tube amplifiers. For this purpose, the symbols $\dot{Y}_{gk}$, $\dot{Y}_{ag}$, $\dot{Y}_{ak}$ and $\dot{S}$ in these basic formulas need merely be correspondingly replaced by $\dot{Y} - \dot{Y}_{fb}$, $\dot{Y}_{fb}$, $\dot{Y}_i - \dot{Y}_{fb}$, and $\dot{S} + \dot{Y}_{fb}$. As a result of this replacement, the following formulas are obtained for three circuits with transistor connection.

The common-emitter circuit:

$$\dot{K} = -\frac{\dot{S}}{\dot{Y}_i + \dot{Y}_n}, \tag{3}$$

$$\dot{Y}_{in} = Y - \dot{K}\dot{Y}_{fb}, \tag{4}$$

$$\dot{Y}_{out} = \dot{Y}_i + \frac{S\dot{Y}_{fb}}{\dot{Y} + \dot{Y}_c}. \tag{5}$$

The common-base circuit:

$$\dot{K} = \frac{\dot{S} + \dot{Y}_i}{\dot{Y}_i + \dot{Y}_n} \approx \frac{\dot{S}}{\dot{Y}_i + \dot{Y}_n}, \tag{6}$$

$$\dot{Y}_{in} = \dot{S} + \dot{Y} + (1 - \dot{K})(\dot{Y}_i - \dot{Y}_{fb}), \tag{7}$$

$$\dot{Y}_{out} \approx \dot{Y}_i - \frac{S(\dot{Y}_i - \dot{Y}_{fb})}{\dot{S} + Y + \dot{Y}_c}. \tag{8}$$

The common-collector circuit:

$$\dot{K} = \frac{\dot{S} + \dot{Y}}{\dot{S} + \dot{Y} + \dot{Y}_i - \dot{Y}_{fb} + \dot{Y}_n} \approx \frac{\dot{S} + \dot{Y}}{\dot{S} + \dot{Y} + \dot{Y}_n}, \tag{9}$$

$$\dot{Y}_{in} = \dot{Y} - \dot{K}(\dot{Y} - \dot{Y}_{fb}) \approx \dot{Y}_{fb} + (1 - \dot{K})\dot{Y}, \tag{10}$$

$$\dot{Y}_{out} = \dot{Y}_i - \dot{Y}_{fb} + (\dot{S} + \dot{Y})\frac{\dot{Y}_c + \dot{Y}_{fb}}{\dot{Y} + \dot{Y}_c}. \tag{11}$$

In these formulas $\dot{Y}_n$ is the load conductance and $\dot{Y}_c$ is the internal conductance of the signal source (or output conductance of the preceding stage).

Equations (3) - (11) permit a computation and analysis of various amplifying circuits with junction transistors.

Multistage Amplifier with Direct Connection between Stages

The circuit of this amplifier is depicted in Fig.2. If the resistances R_2 and R_3 are much greater than the input resistance of a KT_2 triode (which usually

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happens in practice), then it can be fairly reliably assumed that the load of the first stage is constituted by the input conductance of the second. Considering that, at direct stage connection, the amplification factor is not high and that $|Y_{fb}| << |Y|$, it can be assumed that

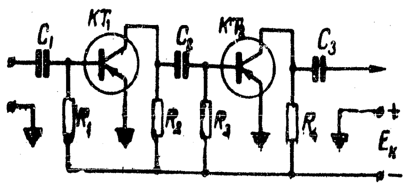

Fig.2

$$\dot{Y}_{in} \approx \dot{Y}. \tag{12}$$

On carrying out a corresponding replacement in eq.(3), and on taking account of the smallness of $\dot{Y}_i$ compared with $\dot{Y}$, we will ultimately obtain

$$\dot{K} = -\frac{\dot{S}}{\dot{Y}} = -\frac{S_0 r_0}{g r_0 + i\omega\tau}. \tag{13}$$

At low frequencies, the amplification factor of a stage equals

$$K_0 = \frac{S_0}{g}. \tag{14}$$

From eq.(13) it is easy to determine the cutoff frequency of the passband, as read from the level of $\frac{1}{\sqrt{2}}$

$$f_{0.7} = \frac{g r_0}{2\pi\tau}, \tag{15}$$

where $gr_0 < 1$ for all triodes. Equation (15) yields the value of the cutoff frequency of a stage. The overall passband of several stages can be found by the same procedure as that used for vacuum-tube resistance-coupled amplifiers.

The efficiency of an amplifier, as evaluated by the product of the amplification factor and the upper cutoff frequency, at direct connection between stages, will be equal to

$$K_0 f_{0.7} = \frac{S_0 r_0}{2\pi\tau} \tag{16}$$

Calculations demonstrate that the same formula determines the cutoff frequency for which amplification is generally possible. An expansion of the passband beyond [redacted] be attained by applying negative feedback and by tolerating a reduction in gain. However, some increase in the efficiency of the amplifier can be achieved by using various corrective networks.

At first glance, eq.(16) leads to the conclusion that the efficiency of a triode is proportional to the base resistance r_0. This conclusion is erroneous, since, actually, the transconductance S_0 and the time constant τ depend in turn on r_0 (Bibl.2)

$$\frac{S_0 r_0}{2\pi\tau} = \frac{A}{2\pi C_{eo}},$$

where $A = \frac{\partial i_e}{\partial u_1}$ is the conductance of the emitter transit in the rectilinear direction and C_{eo} is the diffusion capacitance. If it is considered that the diffusion capacitance is

$$C_{eo} = \frac{Aw^2}{2D}, \tag{17}$$

then, ultimately, the efficiency of the transistor triode

$$K_0 f_{0.7} = \frac{D}{\pi w^2} = f_{cr} \tag{18}$$

is determined by the base thickness w and is numerically equal to the critical frequency f_{cr} (D denotes the diffusion factor of the inserted carriers). As for the power gain, we have, owing to the equality of the input resistance of the stage to the load resistance

$$K_p = K^2. \tag{19}$$

Equations (13), (14), (15), (16), and (18) can be used in computing the preamplification stages in resistors of audio-frequency voltage amplifiers, in video amplifiers, and in pulse amplifiers. In a conventional determination of the passband of a stage it is necessary, in accordance with eq. (18), to select a triode with a critical frequency exceeding at least several times the required band. If the required band is wider than expected from eq.(15), then negative feedback has to be applied.

The load resistance in the collector circuit should be chosen on the basis of the following condition:

$$R_2 = R_n > \frac{10}{g},$$

since, at lower values of R_n the stage gain will decrease and the band will not undergo any notable expansion. The ultimate value of R_n is more accurately determined during the selection of the operating condition by plotting a proper load line.

The selection of the operating conditions of a semiconductor triode can be conducted graphically by means of static characteristics. The methods used are similar to those used in the selection of the operating conditions of a vacuum tube. Figures 3 and 4 depict sets of static characteristics of a semiconductor triode. These diagrams illustrate the plotting of a load line for R_n = 4 kilohms and supply

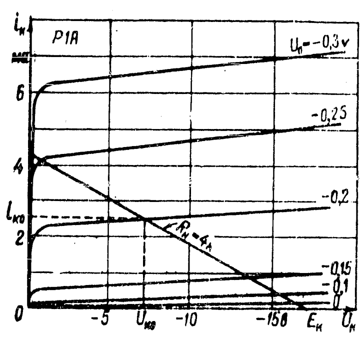

Fig.3

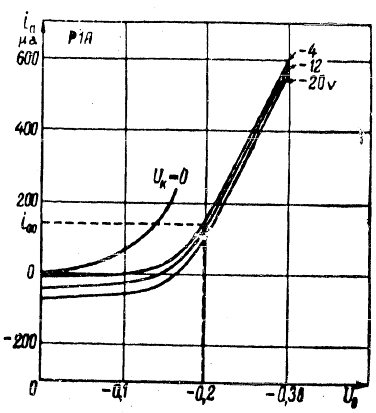

Fig.4

the methods for determining the initial mode. On knowing the initial base current $i_{o\ base}$ it is easy to determine the value of the leakage resistance:

$$R_1 = R_3 \approx \frac{E_k}{i_{o0}}. \tag{20}$$

In computations of separating capacitors it is also possible to apply the analogous formula derived for vacuum-tube amplifiers, upon substituting the leakage resistance in that formula by the input resistance $R_{in} = \frac{1}{g}$. After corresponding transformations we obtain

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[illegible]

[illegible]

Intermediate-[illegible] Amplifier

Maximum possible voltage and power gains are obtained when the output resistance of a stage is matched with the input resistance of the subsequent stage. Figure [illegible] depicts the circuit of an amplifier in which such matching is effected by means of a transforming device with a transformation ratio m.

Considering that the internal conductance of the signal source (or of the preceding stage) is usually very low, we will obtain the following formula for the output conductance at low frequencies:

$$g_{out} = \frac{1}{R_i} + S_0 D_{fb},$$

where $D_{fb} = \frac{\mu_{fb}}{\mu}$ is the reverse grid-through. Here, the following formula is the condition of matching:

$$m_{opt} = \sqrt{\frac{g_{out}}{g}} = \sqrt{\frac{S_0}{g}\left(\frac{1}{\mu} + D_{fb}\right)}. \qquad (21)$$

The amplification factor will be equal to

$$\dot{K} = \frac{m_{opt}\dot{S}}{Y_i + m_{opt}^2 Y}. \qquad (22)$$

At low frequencies, the maximum possible amplification factor is

$$K = \frac{m_{opt} S_0}{2 m_{opt}^2 g} = \frac{\sqrt{\frac{S_0}{g}\left(\frac{1}{\mu} + D_{fb}\right)}}{\frac{2}{\mu} + D_{fb}}. \qquad (23)$$

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Thus, the gain with stage matching, as compared with the preceding case, is

$$\frac{K_{c\,max}}{K} = \frac{\sqrt{g k_1 (1 + \beta D/\beta)}}{2 + \beta D/\beta} < \sqrt{\frac{g R_i}{2}} \qquad (24)$$

times greater. At real values of the parameters of junction-type semiconductor triodes the gain is several times greater.

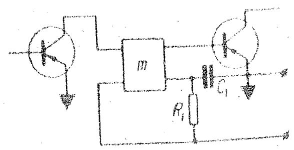

Fig.5

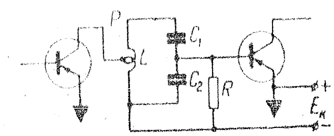

Fig.6

Matching is preferably applied in tuned amplifiers, where it notably increases the stage gain.

Accordingly, it is of interest to determine the cutoff frequency at which gain can still be effected under conditions of interstage matching. Considering that, for high frequencies,

$$\left.\begin{aligned} R_{in} &= r_o \\ \text{and} \qquad R_{out} &= \frac{\tau}{S_0 r_o C_{ok}} \end{aligned}\right\}, \qquad (25)$$

the condition of matching will be

$$m_{opt} = \sqrt{\frac{R_{in}}{R_{out}}} = \sqrt{\frac{r_o}{R_{out}}}. \qquad (26)$$

On substituting this formula into eq.(22) we obtain the amplification factor for the modulus

$$K_{max} = \frac{1}{2}\,\frac{S_0 \sqrt{r_0 R_{out}}}{\sqrt{1 + (\omega\tau)^2}}. \qquad (27)$$

Construing the cutoff frequency as that frequency value at which the amplification factor becomes unity, we will find

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$$f_{n\,red} \approx \frac{S_0 \sqrt{r_0 R_{out}}}{4\pi\tau}. \qquad (28)$$

On equating this formula with eq.(16) it can be easily seen that the cutoff frequency exceeds the critical frequency $\frac{1}{2m_{opt}}$ times, which confirms the possibility of gain at frequencies higher than the critical.

The methods of computing interstage-matching tuned amplifiers with junction transistors is analogous to the methods of computing vacuum-tube SHF amplifiers. For example, when computing an amplifier with a circuit as shown in Fig.6, it is necessary to start from the requirement of ensuring the specified band and maximum gain. On determining the necessary value of equivalent attenuation

$$\delta \geqslant \frac{2\Delta f_{0,7}}{f_0}$$

in accordance with the equivalent circuit in Fig.7, we obtain the following rating formulas:

$$p = \sqrt{\frac{(\delta_e - \delta_k) R_{out}}{2\rho}}, \qquad (29)$$

$$C_2 = \frac{1}{\omega_0 R_{in}} \sqrt{\frac{2R_{in}}{\rho(\delta_e - \delta_k)} - 1} - C_{in}, \qquad (30)$$

$$K_0 = \frac{1}{2} \frac{S_0 \sqrt{R_{in} R_{out}}}{\sqrt{1 + (\omega\tau)^2}} \frac{\delta_e - \delta_k}{\delta_e}. \qquad (31)$$

In these formulas, δ_k denotes the attenuation without taking account of insertion losses, ρ is the characteristic impedance, p is the coupling factor from the collector side, ω_0 is the resonant frequency, C_2 is the matching capacitance, R_{in} and C_{in} are the input resistance and input capacitance as determined from the formula for $\dot{Y}$, $R_{out} \approx R_i$ and $R_e = \frac{\rho}{\delta_e}$.

The characteristic impedance ρ should be so selected as to obtain a positive value of C_2.

Considering that the matching condition (30) depends on ω_0, it is to be satisfied on a higher subband frequency. Here the amplification factor K_0 will depend less on the resonant frequency.

Amplifier with Alternate Stages Connected According to Common-Base and the Common-Collector Circuits

The circuit of an amplifier with alternate stages is depicted in Fig.8. In this diagram, the load of the common-emitter stage is constituted by the input re-

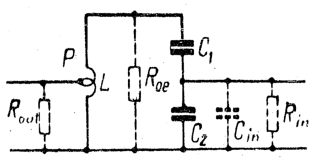

Fig.7

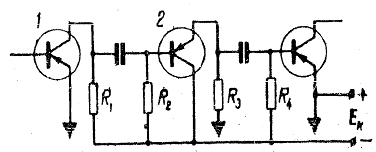

Fig.8

sistance of the common-collector stage, and vice versa. Considering the above and substituting the value $\dot{Y}_n = \dot{Y}$ into eq.(9), let us find the amplification factor and the input conductance of the second stage of this amplifier (Fig.8).

$$\dot{K}_2 = \frac{\dot{S} + \dot{Y}}{\dot{S} + 2\dot{Y}} \qquad (32)$$

and, according to eq.(10),

$$\dot{Y}_{in\,2} \approx (1 - \dot{K}_2)\dot{Y} = \frac{\dot{Y}^2}{\dot{S} + 2\dot{Y}}. \qquad (33)$$

Even if we neglect the conductance $\dot{Y}_i$ compared with $\dot{Y}_{in\,2}$, the amplification factor of the first stage will equal

$$\dot{K}_1 \approx \frac{\dot{S}}{\dot{Y}_{in\,2}}.$$

The total gain of a pair of alternate stages equals

$$\dot{K}_{total} = \dot{K}_1 \dot{K}_2 = \left(\frac{\dot{S}}{\dot{Y}}\right)^2 + \frac{\dot{S}}{\dot{Y}}. \qquad (34)$$

Equation (34) indicates that the gain of a pair of alternate stages practically does not differ from the gain of a pair of stages with identical connection. Consequently, the circuit with alternate stages offers no advantages in the sense of an

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increase in total gain and efficiency.

Conclusions

1. The generalization of the theory of vacuum-tube and transistor amplifiers greatly facilitates the analysis and calculation of the latter type of amplifiers. As a result, the rating formulas are obtained in a sufficiently simple and convenient form. In addition, radio engineers working with transistor amplifiers have to deal with conventional concepts borrowed from the vacuum-tube amplifier theory.

2. In the cases encountered by the author, the discrepancy between computed and actual data was negligible.

3. When using parameters adopted for concrete specimens, the errors do not exceed a few percent. When using averaged-out parameters, the error will be chiefly determined by the extent of the spread of parameters in various specimens.

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CALCULATION OF THE BROAD-BAND GRID TRANSFORMER ACCORDING TO A FIXED INPUT RESISTANCE CHARACTERISTIC

by

Ya.L.Al'terman

The article furnishes a method of computing the optimum parameters of a broad-band grid transformer on the basis of an analysis of a fixed input resistance characteristic. It was assumed that the required input resistance of the transformer is ensured by a shunt consisting of an active resistance connected to the secondary circuit and equal to the internal resistance of the generator, $R_{sh}n^2 = R'_{sh} = R_i$. Using the obtained simple formulas it is possible to determine the maximum transformation ratio for a fixed value of the reflection factor.

Introduction

In computations of a grid transformer for measuring equipment, the requirements made on the frequency characteristic of the transformer are supplemented by the condition that its input resistance Z_{in} should equal a definite value throughout the entire frequency spectrum and should be matched with the internal resistance R_i of the generator. Deviations from such matching in the operating frequency band are estimated by the reflection factor

$$p = \left| \frac{R_i - Z_{in}}{R_i + Z_{in}} \right|. \tag{1}$$

As a rule, the required input impedance of a transformer is ensured by a shunt consisting of a resistance connected to its primary or secondary circuit.

In order for the input resistance of the transformer in the operating frequency

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band to have a definite value and to prevent the deviations from that value from exceeding the set limits, let us calculate the optimum parameters of the transformer on the basis of an analysis of the input resistance characteristic.

For the sake of greater clarity of the basic formulas, we will ignore the resistance in the windings.

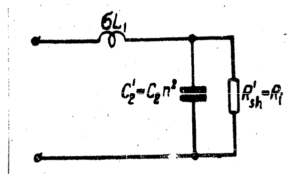

Fig.1

<u>Determining the Maximum Transformation Ratio</u>

The equivalent circuit of a transformer with a shunt connected to its secondary with regard to the form depicted in Fig.1, relative to the upper frequencies of the spectrum.

For this circuit, we have

$$Z_{in} = i\,\omega_v \sigma L_1 - \frac{i\,\omega_v C_2 n^2 R_l^2}{1+\omega_v^2 (C_2 n^2)^2 R_l^2} + \frac{R_l}{1+\omega_v^2 (C_2 n^2)^2 R_l^2}. \qquad (2)$$

Let us designate

$$\omega R_l C_2 n^2 = x, \qquad (3)$$

$$\frac{\sigma L_1}{C_2 n^2 R_l^2} = \alpha. \qquad (4)$$

Then,

$$Z_{in} = \frac{R_l}{1+x^2} + i x R_l \left(\alpha - \frac{1}{1+x^2}\right) \qquad (5)$$

and the reflection factor will be

$$p = \sqrt{\frac{x^2 + (\alpha x^2 + \alpha - 1)^2}{\left(\frac{2}{x} + x\right)^2 + (\alpha x^2 + \alpha - 1)^2}}. \qquad (6)$$

Figure 2 depicts the curves of p = f(x) for various values of α.

Let us represent the graph (Fig.2) of p = f(x) for various values of α in the form of a graph of x = f(α) for various values of p (Fig.3).

As can be seen from the graph in Fig.3, a maximum value of x can be found for every given value of p.

Let us conduct an analysis of eq.(6) to find the optimum relationships. From eq.(6) it can be found that

$$\frac{dx}{d\alpha} = -\frac{2(\alpha x^2 + \alpha - 1)(x^2 + 1)}{4\alpha x(\alpha x^2 + \alpha - 1) + 2x - \frac{8p^2}{x^3(p^2 - 1)}}. \qquad (7)$$

To determine the maximum x we equate the derivative to zero, so that

$$\alpha x^2 + \alpha - 1 = 0$$

and

$$x = \sqrt{\frac{1-\alpha}{\alpha}}. \qquad (8)$$

On substituting the above value of x into eq.(6), we find

$$p = \frac{1-\alpha}{1+\alpha}, \qquad (9)$$

$$\alpha = \frac{1-p}{1+p}, \qquad (10)$$

$$x = \sqrt{\frac{2p}{1-p}}. \qquad (11)$$

Using these simple formulas it is possible to determine the maximum value of x

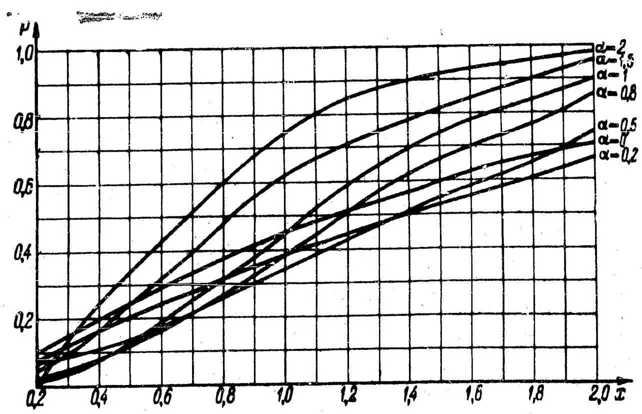

Fig.2

according to a given value of the reflection factor p and, consequently, in accordance with eq.(3), to determine the maximum transformation ratio at a given upper frequency ω_v.

If the shunt is connected to the primary, the equivalent circuit of the transformer for the upper frequencies will have the form depicted in Fig.4.

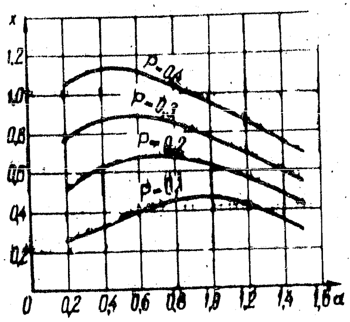

Fig.3

For this circuit, we have

$$Z_{in} = \frac{R_i\left(1 - \omega_v^2 \sigma L_1 C_2 n^2\right)}{1 - \omega_v^2 \sigma L_1 C_2 n^2 + i\,\omega_v C_2 n^2 R_i}$$

In accordance with eqs.(3) and (4)

$$Z_{in} = R_i \frac{1 - \alpha x^2}{1 - \alpha x^2 + ix}$$

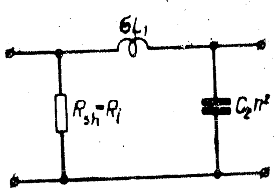

Fig.4

and the reflection factor

$$p = \frac{x}{\sqrt{(1 - \alpha x^2)^2 + x^2}}.$$

Figure 5 depicts the curves of p = f(x) for various values of α.

A comparison of the two versions of shunt connection (Figs.2 and 5) shows that the shunt connection to the secondary - on the condition that the value of p remains

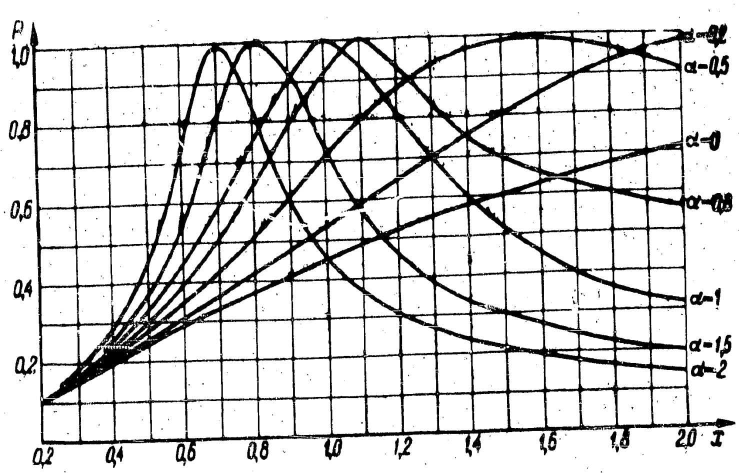

Fig.5

within the range of 0.1 - 0.2 - is more favorable for obtaining the highest value of x.

For example, at $p = 0.2$ and $\alpha = 0.5$, we have:

at a shunt connection to the primary, $x = 0.4$;

at a shunt connection to the secondary, $x = 0.68$.

Appraisal of Frequency Distortions in the Calculation of a Transformer According to a Fixed Input Resistance

Formulas for Upper Frequencies. The frequency distortion factor is determined for the shunt transformer (Bibl.1) according to the following formula:

$$M = \sqrt{(1 - \alpha_{\Pi} x_{\Pi}^2)^2 + x_{\Pi}^2}, \tag{12}$$

where

$$x_{\Pi} = \frac{1 + \frac{R_i}{R'_{sh}}}{1 + \frac{R_i}{R'_{sh}}} x$$

and

$$\alpha_{\Pi} = \frac{1 + \frac{R_i}{R'_{sh}}}{\left(1 + \frac{R_i}{R'_{sh}} \alpha\right)^2} \alpha .$$

Considering that, in our case,

$$R'_{sh} = R_i,$$

Therefore,

$$x_{\Pi} = \frac{1 + \alpha}{2} x \tag{13}$$

and

$$\alpha_{\Pi} = \frac{2\alpha}{(1 + \alpha)^2}. \tag{14}$$

Taking eqs.(10) and (11) into consideration, we obtain

$$x_{\Pi} = \sqrt{\frac{2p}{(1 - p^2)(1 + p)}} \tag{15}$$

and

$$\alpha_{\Pi} = \frac{1 - p^2}{2}. \tag{16}$$

Substituting the values of eqs.(15) and (16) into eq.(12), we obtain the relationship between the reflection factor and the frequency distortion factor for the transformer:

$$M = \frac{1}{\sqrt{1-p^2}}. \qquad (17)$$

Figure 6 depicts the curve for M = f(p).

Formulas for Lower Frequencies. The equivalent circuit of a shunt transformer

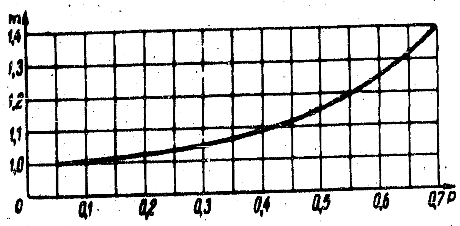

Fig.6

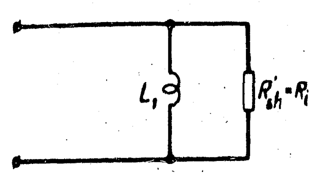

Fig.7

for lower frequencies would have the form depicted in Fig.7.

For this circuit, $Z_{in} = \dfrac{R_i}{1 + \dfrac{R_i}{i\omega_n L_1}}$

On introducing the designation

$$q = \frac{R_i}{\omega_n L_1}, \qquad (18)$$

we obtain

$$Z_{in} = \frac{R_i}{1 - iq}$$

and

$$p = \frac{q}{\sqrt{4+q^2}}. \qquad (19)$$

Whence

$$q = \frac{2p}{\sqrt{1-p^2}}. \qquad (20)$$

It is known that the inductance of the primary winding of a transformer with $R'_{sh} = R_i$ is determined by the following formula:

$$L_1 = \frac{R_i}{2\omega_n \sqrt{M^2-1}};$$

Then, taking eq.(18) into consideration, we obtain

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$$q = 2\sqrt{M^2 - 1}. \tag{21}$$

On equating eqs.(20) and (21), we obtain

$$M = \frac{1}{\sqrt{1 - p^2}} .$$

Thus, the relationship between the reflection factor and the frequency distortion factor in a shunt transformer is the same for both lower- and upper-frequency circuits.

<u>Determining the Highest Value of the Bandwidth in Computing a Transformer According to a Fixed Input Resistance</u>

In accordance with eqs.(3), (4), and (18), we have

$$\frac{\omega_v}{\omega_n} = \frac{x\alpha q}{\sigma} .$$

Then, considering eqs.(10), (11), and (20)

$$\frac{\omega_v}{\omega_n} = \frac{\left(\frac{2p}{1+p}\right)^{\frac{3}{2}}}{\sigma} .$$

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CALCULATION OF QUENCH CIRCUITS

by

F.F.Zhdanov

The article develops further the method of computing quench circuits, previously proposed by the author. Special attention is devoted to determining the limits of applicability of the related rating formulas.

The method of computing quench circuits developed by the author has come into wide use (Bibl.2 - 6). In this connection, it should be noted that the absence of any indication as to the limits of applicability of the rating formulas, derived from this method, often causes difficulties in the use of the formulas and in the interpretation of their specific results.

This leads to the necessity of supplementing the previously proposed method of computing quench circuits by a definition of the limits of applicability of the rating formulas, as was first noted by Cand. Eng. Sci. A.D.Vlasov (Bibl.5). The present article is devoted to resolving this problem and also to further developing the method.

* * *

It is known that the mentioned method of computing quench circuits is based on the requirement for complete elimination of any form of electric discharges on the contacts.

The limiting ratios of current to voltage on the contacts, which determine the condition for the appearance (or disappearance) of electric discharges on contacts at the opening of the latter, are easily established by observation. For silvered contacts these ratios are given in a general form in Fig.1 (curve abcd).

The curve abcd, with an accuracy sufficient for the calculations, can be analytically represented in the form of the following equations:

From $u = 0$ to $u = U_1$

$$i = I_1; \tag{1}$$

From $u = U_1$ to $u = U_2$

$$i = I_0 \frac{u}{u - U_0}. \tag{2}$$

The voltages u on the contacts, ranging from U_1 to U_2, are greater than U_0, as can be seen from Fig.1; consequently, eq.(2) does not yield any indefinite or negative values for the currents i.

In accordance with eqs.(1) and (2), the area of no-discharge on the contacts is encompassed by the following conditions:

at $i < I_2$

$$u < U_2; \tag{3}$$

at $u < U_1$

$$i < I_1; \tag{4}$$

at $U_1 < u < U_2$

$$i < I_0 \frac{u}{u - U_0}, \tag{5}$$

where U_0, I_0, I_1 are constants for a given contact pair, depending on the material, form, and machining of the contacts;

$U_2 = 300$ v is a value constant for any contact;

U_1 and I_2 are determined by the known derivations of eqs.(1) and (2), $u = U_2$, and (2), which yield the following values:

$$U_1 = U_0 \frac{I_1}{I_1 - I_0} \text{ and } I_2 = I_0 \frac{U_2}{U_2 - U_0}.$$

The conditions (3) - (5) represent the basic formulas for computing quench circuits. It is quite obvious that these formulas do not depend on concrete forms of the circuit and, hence, are applicable to the computing of any quench circuit, including circuits consisting of nonlinear elements.

In solving specific problems, the known values are usually: voltage U of the current source, resistance R, and inductance L of the relay winding or of some other

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electromagnetic device.

Considering the above, the wanted parameters of quench circuits and the limits of applicability of the rating formulas are relevantly expressed by these values.

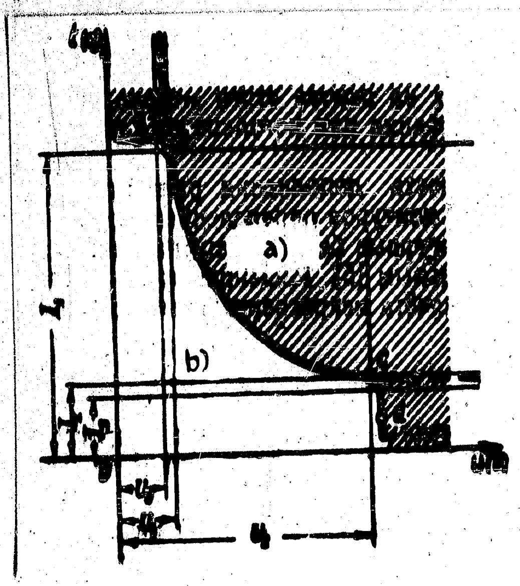


Fig.1

a) Discharge area; b) No-discharge area

The results of the solutions for the most frequently used circuits, as based on the basic rating formulas (3) - (5) and on the assumption that the current in the circuits becomes fixed at the moment of contact opening, are cited in the Table of formulas, given below.

Let us examine in more detail some considerations as to the calculation of a quench circuit consisting of resistance and capacitance (Circuit 3 in the Table).

As is known, the contact voltage in such a circuit at any instant of time after opening of the contacts is determined by the following equation (Bibl.1):

$$u_t = U + Ue^{-nt}\left[\left(\frac{r}{R}-1\right)\cos\omega t + \frac{1}{\omega R}\left(\frac{1}{C}-\frac{r^2+R^2}{2L}\right)\sin\omega t\right],$$

where

$$n=\frac{r+R}{2L}; \quad \omega=\sqrt{\frac{1}{CL}-n^2}.$$

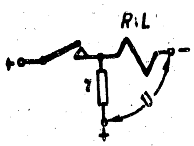
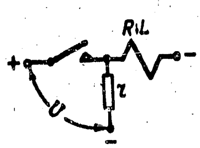
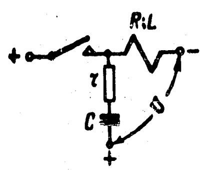
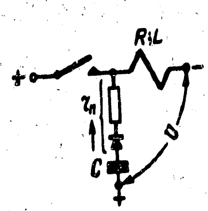
If it is assumed that $r \approx R$, $\frac{1}{CL} \gg n^2$, and that the oscillatory process of change in voltage is weakly damped ($e^{-nt} \approx 1$), then the equation for the maximum contact voltage will have the form

$$U_m = U + \frac{\omega L}{R}U, \qquad (6)$$

where

$$\omega=\sqrt{\frac{1}{CL}}.$$

Table of Formulas for Computing of Quench Circuits

No. of Circuit	Circuit	Formulas	Limits of Applicability of the Formulas
	For any circuit	$u < U_2$	$i < I_2$
		$i < I_1$	$u < U_1$
		$i < I_0 \dfrac{u}{u - U_0}$	$U_1 < u < U_2$
1		$r < \dfrac{U_2}{U} R$	$R > \dfrac{U}{I_2}$
		$r < \dfrac{U_0}{U - I_0 R} R$	$\dfrac{U}{I_1} < R < \dfrac{U}{I_2}$
2		$\dfrac{U}{I_1 R - U} R < r < \dfrac{U_2 - U}{U} R$	$U < U_0$ and $R > \dfrac{UU_2}{I_2 (U_2 - U)}$ or $U_0 < U < U_1$ and $R > \dfrac{UU_1}{I_1 (U_1 - U)}$
		$\dfrac{U - U_0}{I_0 R - U} R < r < \dfrac{U_2 - U}{U} R$	$U_1 < U < U_2$ and $R > \dfrac{UU_2}{I_2 (U_2 - U)}$
3		$\dfrac{U}{I_1} < r < \dfrac{U_2}{U} R$	$U < U_2$ and $R > \dfrac{U}{I_2}$
		$\dfrac{U}{I_1} < r < \dfrac{U_0}{U - I_0 R} R$	$U < U_1$ и $\dfrac{U}{I_1} < R < \dfrac{U}{I_2}$ or $U_1 < U < U_2$ and $\dfrac{U^2}{U_0 I_1 - U I_0} < R < \dfrac{U}{I_2}$
		$C > \dfrac{LU^2}{R^2 [(1+K) U_2 - U]^2}$	$K = 0 \div 3$
4		$C > \dfrac{LU^2}{R^2 [1 + K) U_2 - U]^2}$	$K = 0 \div 3$
		$r_n < \dfrac{U_2 - U (1 - \eta)}{U} R$	$U < U_2$ and $R > \dfrac{U}{I_2}$
		$r_n < \dfrac{U_0}{U - I_0 R} R - (1 - \eta) R$	$U < \dfrac{U_1}{1 - \eta}$; $\eta < \dfrac{U_2 - U_1}{U_2}$ и $\dfrac{U}{I_1} < R < \dfrac{U}{I_2}$ or $U < U_2$; $1 \geq \eta > \dfrac{U_2 - U_1}{U_2}$ and $\dfrac{U}{I_1} < R < \dfrac{U}{I_2}$ or $\dfrac{U_1}{1 - \eta} < U < U_2$; $\eta < \dfrac{U_2 - U_1}{U_2}$ and $\dfrac{U}{I_0} - \dfrac{U_0}{I_0 (1 - \eta)} < R < \dfrac{U}{I_2}$

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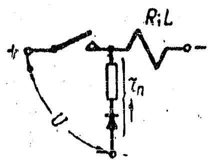
No. of Circuit	Circuit	Formulas	Limits of Applicability of the Formulas
5		$r_n < \frac{U_2 - U}{U} R$	$U < U_2$ and $R > \frac{U}{I_2}$
		$r_n < \frac{U_0}{U - I_0 R} R - R$	$U < U_1$ and $\frac{U}{I_1} < R < \frac{U}{I_2}$ or $U_1 < U < U_2$ and $\frac{U - U_0}{I_0} < R < \frac{U}{I_2}$

Fixed Data

For silvered contact pairs:

$U_0 = 12$ v; $U_1 \approx 17.5$ v; $U_2 = 300$ v; $I_0 = 0.25$ amp; $I_1 = 0.8$ amp; $I_2 \approx 0.26$ amp

For double silvered contacts:

$U_0 = 12$ v; $U_1 \approx 17.5$ v; $U_2 = 300$ v; $I_0 = 0.5$ amp; $I_1 = 1.6$ amp; $I_2 \approx 0.52$ amp

Considering that no air gap can be spanned by a voltage smaller than 300 v, the electrical strength of the gap in separated contacts can be represented by the following formula:

$$U_D = (1 + K) 300, \tag{7}$$

where $K = 0 - 3$ is a coefficient depending on the speed of contact separation.

The condition of the absence of discharges at separated contacts is expressed, in accordance with eqs.(6) and (7), in the form of the inequality $U_m < U_D$, whence

$$C > \frac{LU^2}{K^2[(1 + K) 300 - U]^2}. \tag{8}$$

As indicated by experience, the minimum capacitance of the capacitor depends only to a very small extent on the value of resistance in the quench circuit. Therefore, eq.(8) can be used not only at $r \approx R$ but also at any other value of the resistance of the quench circuit.

When a quench circuit is computed according to Circuit 4 in the Table, the slowly elapsing discharge of the capacitor at the closing of the contacts should be taken into account. If, during the time of open state of the contacts, the capacitor is charged to a voltage U of the current source, and during the time of the closed state is discharged to a value of ΔU, then the residual voltage in the capacitor at the moment of contact opening will equal $U - \Delta U$.

In this case, the voltage on the contacts at the moment of their opening can be represented in the form

$$u = \frac{U}{R} r_n + U(1 - \eta).$$

where: $\eta = \frac{\Delta U}{U}$ is the utilization factor of the capacitor of the quench circuit;

r_n is the total resistance of the quench circuit in the direction of conductance at a current of $\frac{U}{R}$; the valve-element resistance which is part of this total resistance is determined according to the volt-ampere characteristic at the same current.

Subsequently, the procedure for the solution of problems pertaining to circuits with valve elements is analogous to the procedure used for circuits with ordinary spark-quenches.

Here it is necessary to consider, whenever required (when determining the limits of applicability of the obtained formulas) that the value of resistance r_n cannot be negative.

The advantages or shortcomings of a given quench circuit can be appraised only during the solution of specific problems. The principal factors in the selection of a quench circuit are: the voltage of the circuit's power supply source; the current I flowing in the winding of the relay (electromagnet), and the value of the resistance $r(r_n)$ of the quench circuit to the extent to which it affects the rapidity of operation of the relay (electromagnet).

Moreover, in comparing and selecting quench circuits, it is necessary to consider their economic current consumption, simplicity of design, operational dependability of circuit elements, cost, and also simplicity of the circuit wiring and the

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influence of these circuits on the dimensions of the equipment if these dimensions are important to a given design.

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AN ELECTRONIC TELEGRAPH APPARATUS

by

B.P.Terent'yev

The article describes an electronic printing telegraph in which all the basic operations (formation of code sendings at transmission, synchronization, decoding of sendings upon reception, etc.) are realized by means of electron tubes. The diagram of the electronic part of the apparatus is depicted and a brief description of the mechanical part of the mockup of the equipment is furnished.

Introduction

Modern mechanical telegraph equipment is very complicated and requires constant daily attendance and adjustment and cannot operate at high speeds.

The achievements of pulse engineering in the last few years and, in particular, the newly developed pulse telephony systems open wide perspectives for their application in telegraphy. In fact, a study of pulse telephony systems readily shows that they consist in transmission of telephone conversation by telegraph signals. However, the signals of any pulse telephony system differ substantially from telegraphic signals. While the transmission of telegraphic signals occurs at a speed of 150 - 300 signs per minute, the transmission of PM (pulse modulation) signals in telephony occurs at a speed of 8000 - 10,000 signs per second, i.e., the speed of PM transmission is approximately 2000 times higher.

It is obvious that obtaining signals of such a high frequency by means of mechanical devices is totally impossible. Therefore, all operations serving to create pulse signals and to decode them on the receiving end are handled by electronic devices. The handling of telegraph messages by electronic circuits involves no difficulties, since they can produce complex signals at high speeds.

By now, electronics is applied in telegraph equipment, although mostly in frequency- and time-division multiplex devices (Bibl.1) and in retransmitting circuits (Bibl.2).

When designing a telegraph circuit, all operations of coding, decoding, synchronization, distribution, etc. can be adapted to electronic circuitry. The mechanical part of the equipment is left only with the task of printing marks on paper and moving the paper.

At present, the following three methods are chiefly used for sign printing: type wheel, type lever, and the coordinate method.

The two latter methods may have their advantages with respect to quality of printing etc., but they require relatively more power, because they involve a reciprocal motion of comparatively large masses. Moreover, telegraph equipment operating on these two methods includes relatively complex mechanical devices.

In a telegraph apparatus with a continuously rotating type wheel, the motor horsepower can be very low and, if an asynchronous motor and good ball bearings are used, such an apparatus will require practically no attendance.

To facilitate the telegraph operator's work we have adopted the start-stop principle of transmission. At a speed of 382 signs per minute this apparatus can operate in conjunction with the widely employed CT-35 apparatus.

Block Diagram of an Electronic Telegraph Apparatus

Figure 1 depicts the block diagram of an electronic telegraph apparatus during transmission. The formation of coded pulses at transmission is performed exclusively by the electronic part of the circuit. If tape control is not required, the motor of the apparatus can be shut off.

The tapping of a key results in closing of the working contacts (1), (2), (3), (4), (5) (Fig.1) corresponding to the sendings of a given sign. When the key ceases

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to be tapped it causes closing of the contact K_n which connects the start-stop device.

The start-stop device produces brief pulses spaced 22.5 msec apart (at an operating speed of 382 rpm) (Fig.2). The pulses trigger the stages of the time commu-

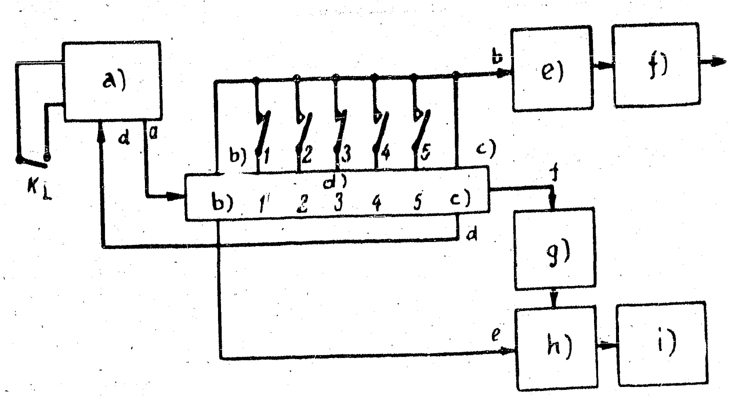

Fig.1

a) Start-stop device; b) Start; c) Stop; d) Time commutator; e) Relay control circuit; f) Transmission relay; g) Delay circuit; h) Control device; i) Electromagnet for blocking of keyboard

tator, so that a single pulse is obtained at each output of the commutator (start 1, 2 ... stop, Fig.2), where also these pulses are spaced 22.5 msec apart. In accordance with the position of the contacts (1), (2), ..., the pulses are conveyed to the transmission relay control circuit. As a result, sendings corresponding to the given sign are obtained at the output of the pulse shaper. These sendings are either directed toward a line or act on the transmission relay.

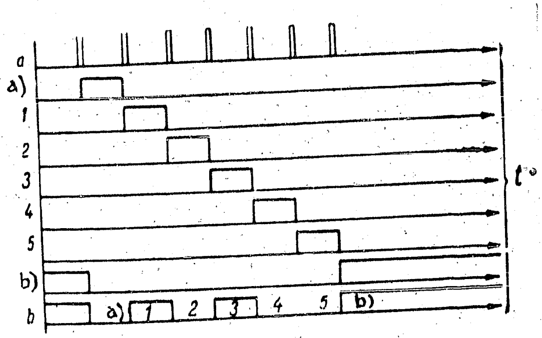

Fig.2

a) Start; b) Stop

The last stage of the time commutator emits not only a pulse going into the wire b but also another pulse going into the start-stop device (wire d, Fig.1) and blocks it (places it on "stop"). The duration of the pulses which, in ordinary telegraph apparatus is determined by rpm of the motor, in an electronic telegraph apparatus is determined by the period of oscillations of the electron-tube oscillator by LC, whose circuit is shunted by another tube having negative feedback. At the "stop" position, this tube is open and its

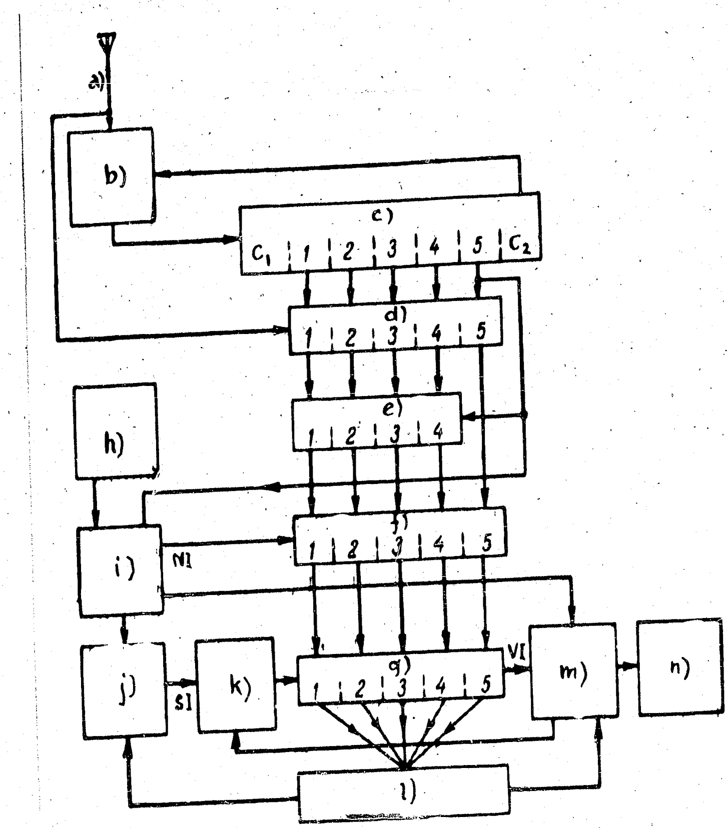

Fig.3

a) Line; b) Start-Stop device; c) Time commutator; c_1) Start; c_2) Stop; d) Selector; e) First tank; f) Second tank; g) Decoder; h) Electromechanical pickup; i) Pulse shaper; j) Recorder-switching circuit; k) Key circuit; l) Circuit for identification of service combinations; m) Decoder-control circuit; n) Print electromagnet

shunting action is so great that the oscillator is inactivated. A start signal closes the shunting tube and the subsequent tube then produces brief pulses out of the received oscillations by means of limitation and differentiation.

In order to block the rise of the key back to its usual position and the opening of contacts 1, 2, ..., until the termination of the transmission of the signal

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sign, and thus to avert a distortion of the signal, a keyboard-blocking electromagnet is provided which is connected by the start pulse (wire e) through the control device. This magnet is disconnected by the "stop" sending through the delay device which determines the minimum length of the "stop" sending.

On receiving, the start signal arriving from the line (Fig.3) connects the start-stop device which switches the time commutator, just as in the transmitting operation. But unlike in transmission, the pulses used in reception are brief and are obtained by differentiation of the leading edges of the pulses leaving the commutator. These pulses are conveyed to a "selector", to which also the signals coming from the line are fed.

The selector consists of five DGTs diodes, closed by some bias voltage. The signals incoming from the line are simultaneously fed to all these diodes. The working pulses from each commutator successively strike each diode of the selector. The closing voltage of the diodes of the selector is higher than the signal and pulse voltages from the commutator, taken separately, but is lower than the total voltage of the latter. If a sending arrives at the instant of appearance of a pulse on some stage of the commutator, the voltages of the sending and the pulse will combine and a current pulse will pass through a given selector diode and arrive at the first tank. The appearance of only one pulse from the commutator or the line does not cause the selector to operate. The tank consists of five unrelated triggers with two stable equilibrium positions each. At the initial moment, they are all in the initial position (left-hand triode closed). The pulses passed by the selector displace the corresponding stages of the tank to their other stable position. Thus, the received combination is, as it were, "recorded" or "remembered" on the first tank. In order to select the optimum "phase" of pulses coming from the commutator, a delayed multivibrator is introduced between the start-stop device and the commutator, in receiving operation, for the purpose of shifting the pulses within the limits of an entire sending. The possibility of such shifting and also the shortness of the sounding pulse, which is of the order of 10 μsec, as well as the high stability of the oscillator frequency of the start-stop device, impart to this apparatus a highly satisfactory correcting ability.

The fifth working pulse of the commutator not only determines the quality of the sending but also places all stages of the first tank in their "initial position". Here, the stages of the first tank, displaced by the previously received sendings into their other stable position, will emit from their outputs voltage pulses which will "displace" the corresponding stages of the second tank. Thus, after the fifth working sending, the accepted combination will already be recorded on the second

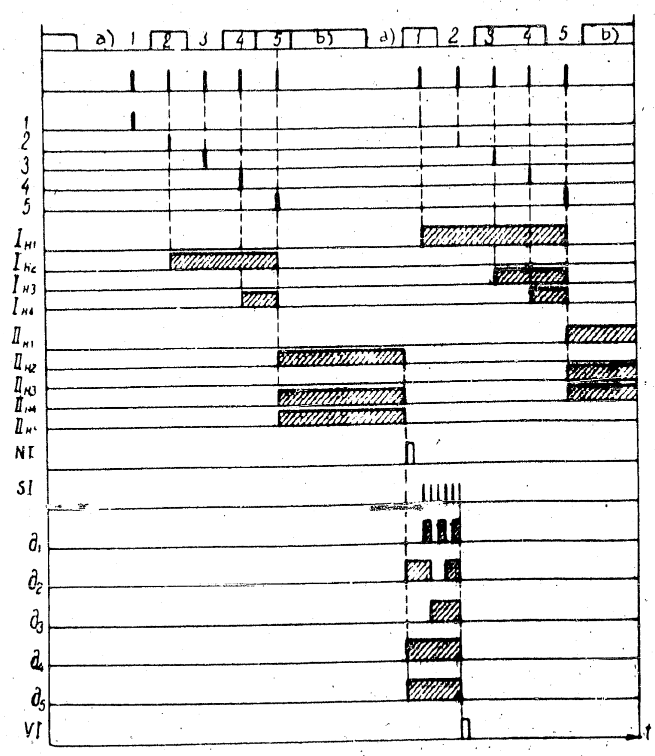

Fig.4

a) Start; b) Stop

tank, while the first tank is by then ready to receive new signals. Hence it is obvious that the electronic telegraph poses no requirements as to the length of the "stop" sending.

The second tank serves to prevent any relationship between the rpm of the type wheel and the speed of transmission. This is achieved in the following manner:

A toothed steel wheel is mounted to the shaft rotating the type wheel. The number and distribution of the teeth correspond to the distribution of type faces on the type wheel. When the teeth pass in front of the electromagnet, a variable emf, arises in the winding of the electromagnet. The pulse shaper uses this emf to produce ("counting") pulses. The phase of the pulses is selected by so displacing the electromagnet that the appearance of the pulses would correspond to the instant of passage of the type faces in front of the printing device.

Fig.5

In addition, the emf of the pickup serves to produce another pulse. This pulse ("initial" NI in Fig.4) continues until the appearance of the counting pulses (SI in Fig.4). The pulse shaper generates the initial pulse and directs it to the second tank. Then, information from the second tank is fed to the decoder (II_{n1} to II_{n5} in Fig.4).

The decoder represents a five-stage binary translating circuit, i.e., five trigger stages with two stable equilibrium positions. If it is conditionally assumed that the first equilibrium position, for example, with the left-hand tube open and the right-hand tube closed, corresponds to zero, and the second equilibrium position corresponds to unity, any number can be "recorded" on these stages according to the binary system.

Here the first stage will designate the number of ones in the number, the second will designate the number of twos, the third - the number of fours, the second --the number of eights, and so forth. In this way, a sign corresponding to a combination of sendings equal to 10101, where the current sending is nominally denoted by unity and the no-current sending by zero, will correspond to the number of

1 + 4 + 16 = 21

The combination "recorded" in the second tank is conveyed to the decoder in stages, i.e., the decoder, as it were, "remembers" the number denoting the ordinal number of the sign (d_1 - d_5, Fig.4). After recording of the accepted combination in the decoder, pulses formed from the emf of the electromagnetic pickup (counting pulses) arrive at the input of the decoder. When the sum total of the number conveyed from the tank and of the pulses arriving from the electromagnetic pickup reaches 32, the decoder assumes its initial position, i.e., the position in which it will record 00000. Then, a pulse arises at the output of the decoder and proceeds toward the decoder control circuit. The control circuit transmits that pulse to the key circuit and blocks it; thus the decoder is inactive until the arrival of a new code combination. Furthermore, the decoder control circuit transmits the output pulse (VI in Fig.4) to the print electromagnet. The electromagnet presses a paper tape against the type wheel and prints the sign.

As can be seen from these explanations, normal reception and decoding of signals will be effected when the frequencies of the oscillators of the start-stop devices at the transmitting and receiving ends are mutually matched, totally independent of the rpm of the type wheel. The latter determines merely the speed of "counting" the signals from the decoder. The sole requirement made on the rpm of the type wheel is that it should be higher than the speed of transmission of the signs, since only one sign can be printed during any single revolution of the wheel. The switching of registers in the apparatus is also effected by purely electronic means.

As can be seen from the block diagram, the code combination is fed from the decoder to the circuit for identification of service combinations, which consists of several circuits of five lock-ons. On reception of a service combination (switching of register, bell, and others) the appropriate lock-on circuit emits a pulse at the output.

In particular, on receiving a register-switching signal, the pulse emitted at the output of the circuit travels toward the register-switching circuit. As noted before, the electromagnetic pickup produces a number of pulses equal to the number of type faces on the type wheel, with the even pulses corresponding to the signs of the alphabet register and the odd pulses corresponding to the signs of the numeral register. Depending on the received combination, the circuit for identification of service combinations connects the register-switching circuit in such a way that the pulse-shaping device would feed either odd or even pulses to the key circuit from the pickup, depending on whether the alphabet- or the numeral-register signs are to be printed.

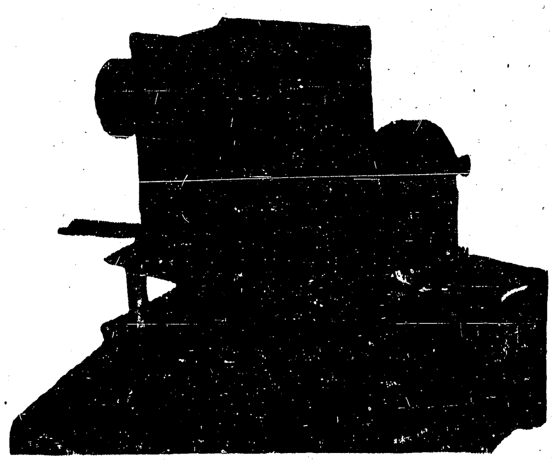

Fig.6

At the same time, the circuit for identification of service combinations blocks the stage of the decoder control circuit controlling the print electromagnet, so that there is no printing of the sign.

The first version of this electronic telegraph was assembled for checking the correctness of operation of the electron-tube circuit; the test model contained 37 dual triodes and one low-power thyratron for controlling the print electromagnet* (Fig.5)

Figure 6 gives a front view of the mechanical part of the mockup. Individual units of various telegraph apparatuses were used here. In particular, a type wheel from a Baudot telegraph was used in the receiving part. The print magnet is located

*A valuable contribution to the development and perfecting of the electronic telegraph has been made by E.D.Demin.

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under that wheel. The wheel and the electromagnet of the electromechanical pickup are visible inside the interlocking frame. The keyboard-blocking electromagnet is shown in the foreground.

Considering that nearly all tubes of this telegraph operate on the "yes – no" mode, the circuit makes low demands on the stability of tube characteristics and power supply sources. The electronic part of the apparatus normally operates at a proportional change in feed voltage by ±25% (150 – 250 v).

The application of second tanks permitted variation in rpm of the type wheel within a very broad range. Correct recording was obtained at a variation of 60 to 500 rpm. The required power equals approximately 180 w; out of this about 140 w is expended on incandescence.

The advantages displayed by this mockup confirm the usefulness of continuing the work, especially with regard to reducing the required power. The replacement of electron tubes by transistor elements will make it possible to reduce sharply the required power.

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OBTAINING HIGH HARMONIC NUMBERS BY MEANS OF A MAGNETIC HARMONIC OSCILLATOR

by

L.T.Kim

The article describes several circuits for achieving high harmonic numbers by means of a complex-loaded magnetic oscillator. With these circuits it is fairly easy to obtain one or two fundamental-frequency harmonics and to greatly suppress the other harmonics.

Magnetic harmonic oscillators are widely used in telecommunication engineering as carrier-frequency sources for multiplex traffic systems. The principal wiring diagram of the harmonic oscillator is depicted in Fig.1; its operation has occasionally been described in the literature [for example, (Bibl.1,2)].

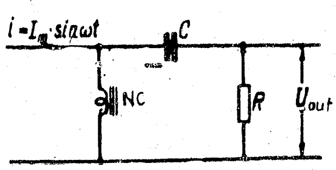

Fig.1

Depending on the mutual ratio of the parameters L_s, R, and C of the discharge circuit (L_s denotes the inductance of the nonlinear coil NC in the saturated state), it is possible to obtain several basic versions of the spectrum distribution for the harmonics of the oscillator output voltage, depicted in Fig.2. As indicated in Fig.2, a harmonic oscillator designed according to the circuit in Fig.1 can yield a fairly uniform spectrum distribution of harmonics.

Sometimes it is necessary to obtain one or two fundamental-frequency harmonics. In this case, several frequency-multiplying stages are usually employed; use of a harmonic oscillator, designed on the basis of the circuit in Fig.1, would be ineffective here since only a small part of the expended power is converted into useful power; furthermore, filtering of the required harmonic is hampered here. However, stage multiplication cannot be applied if the number of the harmonic is a prime number; in this case direct use of the fundamental-frequency harmonic is necessary. We have conducted experiments to obtain high numbers of harmonics by means of a complex-loaded magnetic harmonic oscillator.

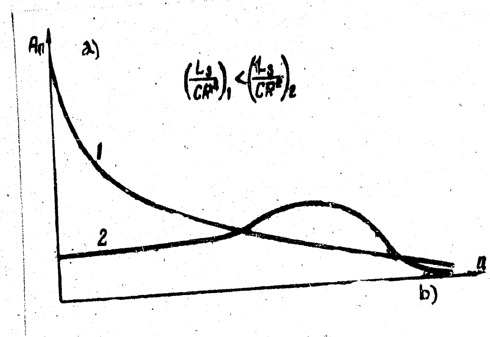

Fig.2

a) Amplitude of the n^{th} harmonic;

b) Harmonic number

Figure 3 shows three versions of the spectrum characteristics of the oscillator output voltage, all obtained at the same fundamental-frequency input power. The first curve corresponds to the circuit in Fig.1. The second curve was obtained on replacing the resistance R by a parallel-connected oscillatory subcircuit tuned for the 33rd fundamental-frequency harmonic with a quality factor of Q = 50. If the parallel-connected circuit is replaced by a series-connected one, we obtain curve 3. Thus, the replacement of the active load by a parallel-connected oscillatory circuit or tank permits obtaining a more than twentyfold gain in the necessary-harmonic voltage. Moreover, while in the case of the active load, adjacent harmonics have approximately the same amplitude as the necessary harmonic, in the case of oscillatory-circuit load these adjacent harmonics have much smaller amplitudes, which facilitates filtering.

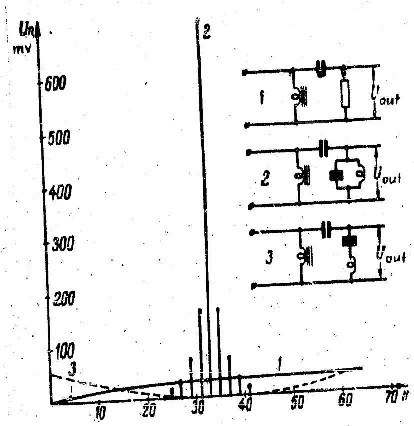

Fig.3

If a circuit with an even higher quality factor is used, the adjacent harmonics can be suppressed still more. Sometimes such filtering suffices and the circuit voltage can be utilized directly. At other times, when it is necessary to effect a higher filtering, more complicated circuits have to be employed. Figure 4 depicts a circuit for obtaining the 21^{st} harmonic of 16-kc frequency.

Figure 5 shows a circuit for obtaining two 8-kc frequency harmonics: the 41^{st} and 43^{rd}.

In these circuits the complex loads consist of band filters. The spectrum distribution of harmonics at the inputs of the filters is analogous to curve 2 in Fig.3; at the output of the filters, the amplitudes of the unused harmonics are at least 7 nepers below the amplitudes of the useful harmonics. The characteristics of the effective attenuation of the filters PF-328, 336, and 344 kc are given in Fig.4.

On a 2-ohm load a useful-harmonic voltage of no less than 2 v is obtained. If it is possible to operate on a high-ohmic load (amplifier, oscillograph), then the output voltage can be increased to 2.8 - 3 v by eliminating a 2-ohm load resistance from the circuit. At lower load-resistances, the voltage decreases and the filtering deteriorates.

The circuit in Fig.5 differs from that in Fig.4 by the fact that the fundamental-frequency amplifier has a balanced output; the harmonic oscillator is also balanced.

As can be seen from Figs.4 and 5, the harmonic filters have Pi-type endings; the frequency characteristics of their input resistance are analogous to the frequency characteristics of K-type filters (Bibl.3).

The input impedance of a given filter has a reactive character when operating on the mean frequency of the passband of an adjacent filter and, in terms of absolute value, amounts to 10 - 12% of the characteristic impedance. As a result, the filters virtually do not affect each other.

The required 16-kc frequency power needed by the oscillator for obtaining the 21^{st} harmonic (Fig.4) amounts to 130 mw. The oscillator in Fig.5 requires 150 mw.

The nonlinear coils of the harmonic oscillators are built up from ferrite rings with a rectangular hysteresis loop; if thin toroidal plates of 78% molybdenum-

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permalloy are used as the core, it is possible to obtain higher harmonic voltages at an unchanged power expenditure.

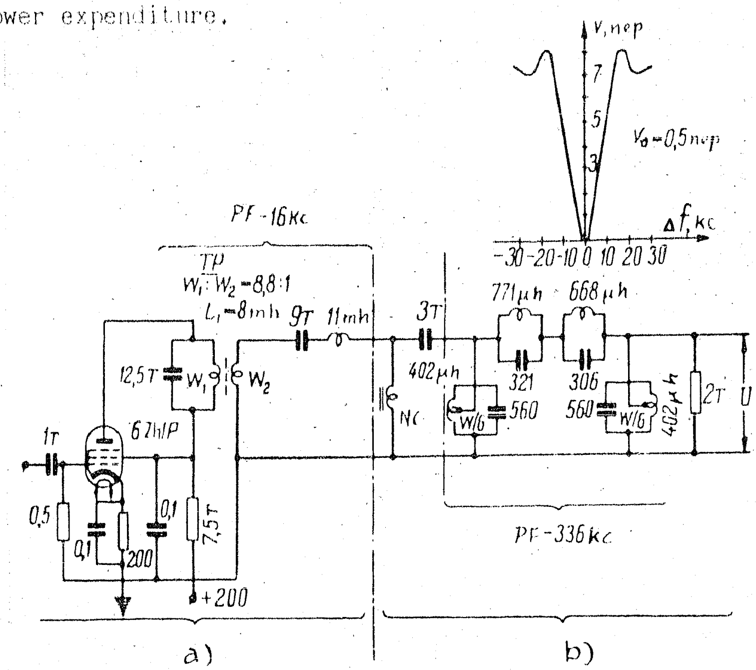

Fig.4

a) 16-kc amplifier; b) Generator of the 21st harmonic

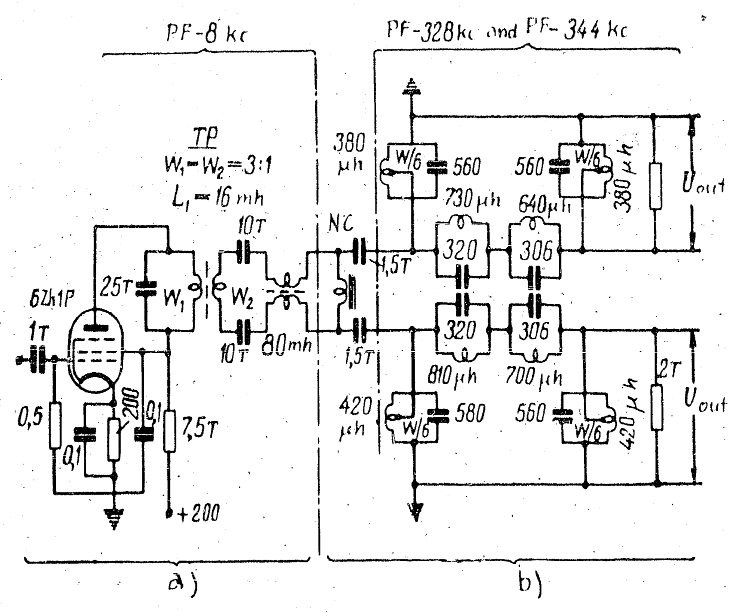

Fig.5

a) 8-kc amplifier; b) Generator of the 41st and 43rd harmonics

It is known that even harmonics are obtained from the magnetic oscillator by using full-wave rectification of the oscillator output voltage. When the even and odd harmonic outputs are loaded with resistances, the spectrum characteristics of both output voltages will have a uniform character. If the active loads are replaced by complex loads it is impossible to achieve a definite predominance of some harmonics over others, which is possible in circuit (2) in Fig.3. Therefore, it is more convenient to obtain frequencies corresponding to even harmonics as the odd harmonics of some other frequency. For example, in Fig.4, the 336-kc frequency which is the 42^{nd} harmonic of the 8-kc frequency is generated as the 21^{st} harmonic of the 16-kc frequency. The latter frequency is easily obtained from the former by means of full-wave rectification.

The above-reviewed circuits for obtaining high harmonic numbers by means of magnetic oscillators can be designed for various powers and frequencies, have small dimensions, and are very stable in performance.

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CHANNEL LOAD DURING ON-DUTY TALK BY THE TELEPHONE OPERATOR

by

V.M.Belous

The article points out the unsatisfactory technical operation of group systems with respect to prevention of overloads. It describes a method for protecting group channels from the high voltages arising during on-duty talk by telephone operators.

The present article is part of a larger study of group-channel load in high-frequency multiplex telephone systems. Unfortunately, such loads in such systems is a problem that has been little explored. The available data indicate that the formulation of measures for a more rational group-channel load is one of the major ways of increasing the lengths of repeater sections and, consequently, of reducing the construction and operating costs of long-distance intertoll trunks.

In connection with the above, the problem of channel loads due to on-duty talk by telephone operators acquires great importance.

As evidenced by the data taken, on-duty talk by telephone operators involves considerable power which can be dangerous from the viewpoint of possible overloading of a system.

Observations established that the power developed during on-duty talk by telephone operators using identical talking sets depends on many factors.

The principal such factors are: the manner and conduct of talk characteristic of a given operator; matching with the telephone operator of the other station; number of channels serviced by the operator; duration of continuous on-the-spot duty.

The first factor, which affects the magnitude of the power developed, is self-evident and requires no explanation.

The three other factors require an explanation.

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If the work of a telephone operator in one station is matched with the work of a telephone operator in another station, the on-duty talk is conducted without increasing the tone. At unmatched work of operators, the normal tone is violated during on-duty talk and there even occur isolated "shrieks". The less channels are serviced by an operator the calmer will be the tone of her speech. At the beginning of her daily shift, the operator talks with a raised voice for some time before achieving a rhythm in her work. When the end of the shift approaches the operators become fatigued, and most of them also become somewhat irritable, which affects the volume of speech.

All the above-described peculiarities of the work of telephone operators have to be taken into account when investigating the channel load during on-duty talk.

For the purpose of determining the probable values of the voltages arising during on-duty talks by a telephone operator, the curves of the voltages arising during the work of five operators were photographed.

The voltage curves on the photographs differ by a small value (0.2 - 0.1 neper),

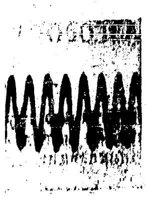

Fig.1

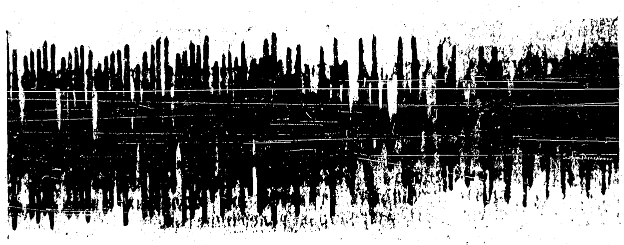

Fig.2

which justifies citing only a few characteristic curves.

The photographing of the voltage curves was conducted at points with a level of measurement equal to -1.3 neper*. The comparison of the curves was effected by juxtaposing them to the scale curve shown in Fig.1 (voltage with a frequency of 8 cycles and a level of 1.6 neper).

*Which corresponds to the locus of connection of the limiter.

Figures 2 and 3 depict photographs of the curves of the voltages arising at measuring-level points of 1.3 neper during on-duty talks by several different telephone operators on a single station. The peak value of the voltage levels for the curves in Figs.2 and 3 equals approximately -0.05 neper.

Obviously, during such on-duty talk by an operator, a voltage with a level of -0.25 neper will arrive at the input of an individual converter in the K-24 and V-25 systems. A voltage with such a level on the input of an individual converter is dangerous from the viewpoint of overloading the system and increasing the nonlinear distortions.

In the K-24 and V-12 systems, the amplitude limiter should ideally be represented by an individual converter; as indicated by past measurements, this is a poor amplitude limiter.

Figure 4 shows the amplitude characteristic of an individual V-12 converter.

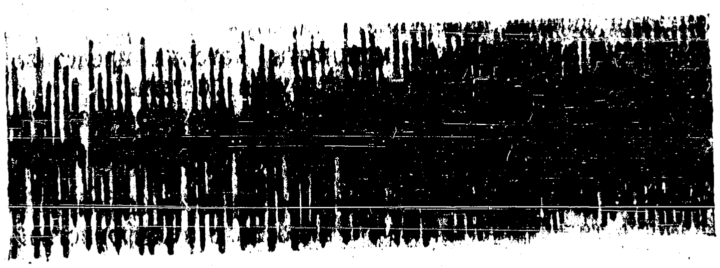

Fig.3

The curve is approximately linear up to a level of -0.3 - 0 neper at the input of the converter, which does not suffice for limiting the amplitudes of the voltages arriving at the input of the group channel.

At present, the Soviet Industry for channel-separating equipment is manufacturing amplitude limiters of a quality sufficient for total protection of group channels from overload.

The amplitude limiter is installed in the circuit at the points with a measuring level of -1.3 neper.

The on-duty talk of the operator in the other station (Fig.3) was tape-recorded and was then fed from the tape recorder, at a level of -0.05 neper, to the amplitude limiter. The curve of the voltage arising at the output of the amplitude limiter is depicted in Fig.5.

The peak value of the voltage curve at the output of the amplitude limiter equals approximately -1.25 neper.

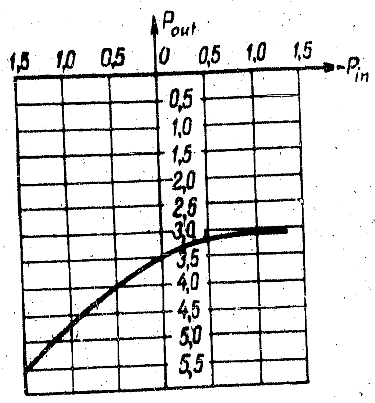

Fig.4

Photographing the curves of the voltages arising during subscriber talk established that the level at the points with a relative zero level varies approximately from -1.1 to -2.5 neper*. Consequently, the maximum value of the level at the input of the limiter will be approximately -2.4 neper. Therefore, the voltages arising during transmission of subscriber talk are not limited and the limiter does not introduce nonlinear distortions.

In view of the existing standard for the number of channels to be serviced by one telephone operator, simultaneous on-duty talk by three operators can be expected

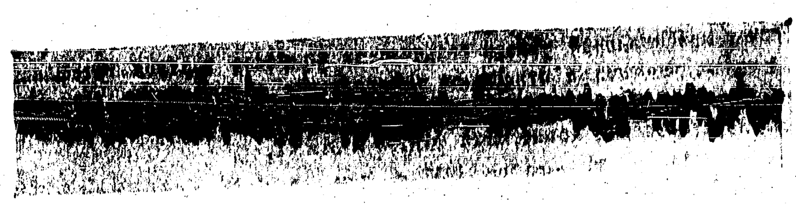

Fig.5

for a twelve-channel system, simultaneous on-duty talk by six operators can be expected for a twenty-four channel system, etc.

According to statistical data, the number of conversations passed by a single channel at peak load is six, and the standard for the duration of on-duty talk by the operator is nine seconds.

*The voltages were photographed during transmission of speech by ten different subscribers.

Thus, when a telephone operator services four channels, the duration of on-duty talk by the operator will be:

$$9 \times 6 \times 4 = 216 \text{ sec} = 3.6 \text{ min}$$

In servicing 12 channels, the duration of on-duty talk will be 10.8 min and on servicing the 24 channels of the K-24 system, this duration will be 21.6 min.

As can be seen, the performance of a telephone system worsens during a considerable interval of time.

The amplitude limiter may serve as an effective protection of group systems from overload during on-duty talk. However, insufficient attention has been paid to the protection of group systems from overload.

At the present, no amplitude limiters are used in toll telephone offices, with a considerable number of group systems and with a total number of channels amounting to 300 and more. This indicates that the struggle against noise and crosstalk in group systems is not waged efficiently, and that the technical operation of telephone systems is not satisfactory.

When there is no amplitude limiter, it is expedient to connect an attenuator to the operator's talking set.

Observations conducted by the author indicate that, on connection of a 0.5-neper attenuator to the talking set of the telephone operator, the latter voiced no complaints about any deterioration in the quality of audibility.

Article received by the Editor 25 September 1956.

THE INTERNATIONAL TELEPHONE AND TELEGRAPH CONSULTATIVE COMMITTEE (CCITT)

by

P.A.Frolov

The article describes the structure of the unified CCITT and the principal resolutions on questions of telephony and telegraphy adopted by the 18^{th} Plenary Assembly of the CCIF and the 8^{th} Plenary Assembly of the CCIT.

In December of 1956 in Geneva (Switzerland), the 18^{th} Plenary Assembly of the International Telephone Consultative Committee (CCIF) and the 8^{th} Plenary Assembly of the International Telegraph Consultative Committee (CCIT) were held. These were the last to be convened separately by both organizations since, in accordance with a resolution of the International Telecommunications Union (ITU), they merged into a single International Telephone and Telegraph Consultative Committee (CCITT). The merger was decided after unanimous approval by the parties concerned - the members of the International Telecommunications Union. This unification of the CCIF and the CCIT into a single CCITT will no doubt facilitate further progress in wire communications engineering. The development of an international network of voice-frequency telegraphy, subscriber telegraphy (Telex service), and phototelegraphy necessitates a maximum correlation of the technical aspects of telephone and telegraph channels, especially with respect to standardization of equipment. Experience shows that lately much of the work of the investigating commissions of the CCIF was duplicated by the investigating commissions of the CCIT, and vice versa.

During the period from 15 to 20 December 1956 the First Plenary Assembly of the new CCITT convened in Geneva, where it elected a new director, adopted the new statutes of the CCITT, elected chairmen and vice chairmen of investigating commissions, examined and approved the questions to be investigated in the next few years, mapped out a study program for the investigating commissions, and reviewed the question of providing technical assistance to countries with underdeveloped communications.

The newly elected director of the CCITT was M.Rouvier (France), Director General of the French PTT.

Considerable discussion was evoked by the question of the structure of the new CCITT: after examination of several proposals, the structure depicted schematically in Fig.1 was finally adopted.

This flow sheet indicates that the CCITT will consist of only 12 investigating commissions (the structure of the CCIF and the CCIT contained 22 such commissions), one commission for the development of an international telegraph network, and one telephone laboratory.

The flow sheet also shows that the commissions are preferably set up in a "mixed" form, i.e., they will investigate problems pertaining to both telephone and telegraph engineering.

The most important among the investigating commissions are: Commission No.1, concerned with general problems of transmission (telephone, telegraph, radio broadcasting and television); Commission No.2, concerned with the general problems of operation and rate-scheduling of wire communications; Commission No.3, concerned with the problems of connecting radio-relay lines to wire lines; Commission No.4, concerned with the problems of the technical operation of wire communication facilities; Commission No.8, concerned with the problems of telegraph equipment and of facsimile and phototelegraphy; and Commission No.11, concerned with the problems of telephone commutation.

The new structure of the CCITT requires a more careful and coordinated preparation of data by the Soviet scientific-research organizations, but on the other hand it will accelerate the solving of the problems posed for study.

The following Soviet delegates are among the governing members of the investi-

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gating commissions of the CCITT:

S.V.Borodich, senior researcher at the Scientific-Research Institute of the Ministry of Communications, Vice-Chairman of Investigating Commission No.3, concerned

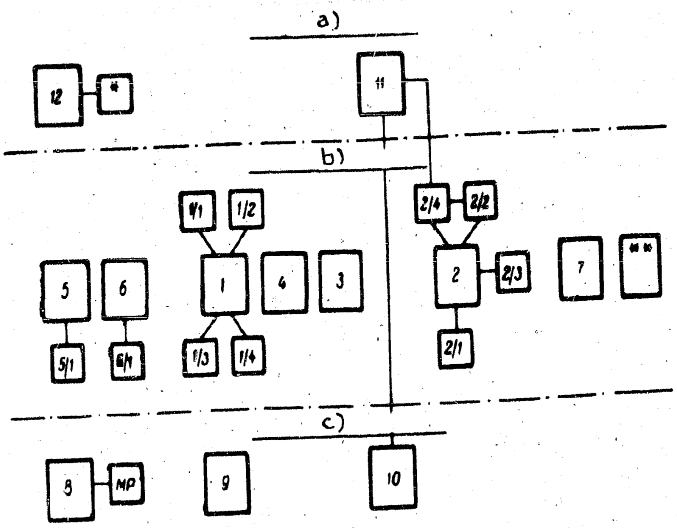

Fig.1 - Structure of the CCITT

a) Telephone commissions; b) Mixed commissions; c) Telegraph commissions;

1. General problems of transmission (1/1 - Specifications of long-distance lines; 1/2 - Ulilization of lines for telephony; 1/3 - Utilization of lines for telegraphy; 1/4 - Utilization of lines for broadcasting and television); 2. Operation and rate-scheduling (2/1 - Telegraph operation and rate-scheduling; 2/2 - Telephone operation and rate-scheduling; 2/3 - Net costs of communication services; 2/4 - Semiautomation and automation, operation); 3. Radio-relay communications; 4. Technical operation of electrocommunications services; 5. Noise and interference prevention (5/1 - Review of directives); 6. Corrosion prevention (6/1 - Review of recommendations); 7. Terminology; symbols; 8. Telegraph equipment, facsimile, phototelegraphy; MP - Radiophototelegraphy; 9. Quality of telegraph transmission, channel operation; 10. Telegraph commutation and telex; 11. Telephone commutation and signaling; 12. Quality of telephone transmission; * Laboratory; ** Plan for the development of international network

with the problems of connecting radio-relay lines to wire lines.

M.I.Mikhaylov, Professor, Chief of the Laboratory of the All-Union Scientific Research Institute for Communications in the Ministry of Communications, Vice-Chairman of Investigating Commission No.5, concerned with the problems of protecting communication installations from noise and from dangerous effects of high-voltage lines.

V.N.Amarantov, Chief Engineer of the Scientific-Research Institute of the Ministry of Radio-Engineering Industry, Vice-Chairman of Investigating Commission No.8, concerned with the problems of terminal telegraph equipment and facsimile telegraphy.

S.A.Vasil'yev, Head of the Laboratory of the All-Union Scientific-Research

Institute for Communications in the Ministry of Communications, Vice-Chairman of Commission No.11, concerned with the problems of telephone commutation and signaling.

The First Plenary Assembly of the CCITT actively discussed the question of providing technical assistance to countries with underdeveloped communications systems. The delegates noted that so far the assistance extended to such countries proved insufficient; this requires better familiarization with the needs of these countries. Further, the delegates pointed out that it would be expedient to publish technical information on various engineering problems and practical directions on design, construction, and operation of communications media; they noted the inadequacy of the funds allotted for this purpose by the International Telecommunications Union (ITU).

The First Plenary Assembly of the CCITT approved a document drafted by a study group on the principal forms of the realization of technical assistance. Further, the Assembly acclaimed the resolution of the Plenary Assembly of the International Radio Consultative Committee (CCIR) concerning the organization of a provisional commission which would explore this question more thoroughly and draft concrete proposals and recommendations for submission to the Administrative Council of the ITU at its next session.

The First Plenary Assembly of the CCITT was preceded by meetings of all investigating commissions of the CCIF and the CCIT, and also by the final plenary assemblies of the CCIF and the CCIT which examined and approved the reports of the investigating commissions and of the budgetary sections.

The delegates of the final plenary assemblies of the CCIF and the CCIT commemorated by a solemn ceremony the long, active, and useful labors of their directors, Messieurs Valenci and Townsend who retired because of age.

In all, the sessions of the 22 investigating commissions of the CCIF and the CCIT took care of about 300 questions on the agenda; however, no conclusive reports were prepared on most of these questions and they were deferred for further study.

The purpose of the activities of the CCITT consists in drafting standards and

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recommendations for communication systems so that these systems, developed and constructed in various countries, could be easily and without obstructions connected to a general international telephone and telegraph network. In addition, the CCITT is concerned with developing standard rules for general and technical operation of various communication systems.

Accordingly, any question which has been fully studied should be decided upon unanimously by the delegates of all countries, in a resolution specifying concrete electrical standards for communication channels and equipment or technical directives and recommendations.

The 300 questions examined during the sessions of the investigating commissions of the CCIF and the CCIT should be classified into groups pertaining, respectively, to general and technical operation of telephone and telegraph communications, rates and net costs of communication services, qualitative indexes of all forms of communication and radio broadcasting, methods of protecting communication installations from various sources of interference, standardization of new communication systems, definition of the characteristics of telephone, telegraph, phototelegraph, radio, and television channels, organization of a Pan-European network of semiautomatic dialing, and also of a direct-link telegraph network.

A brief description of certain questions which are most important from our viewpoint and of the adopted recommendations is given below.

New Coaxial System with a Frequency Band to 12 Mc

The frequency spectrum standardized by the CCIF for the 960-channel system (with an upper frequency of 4028 kc) will not be altered and the channels for further multiplexing will be located in the bands above 4028 kc.

In the new system, the basic group will be represented by a tertiary group of 300 channels each, which occupies a frequency band from 812 to 2044 kc and consists of 4, 5, 6, 7, and 8 secondary groups.

Altogether, it is intended to apply six tertiary groups spaced at 88-kc intervals in the frequency band above 4028 kc.

The frequency interval between the 16th secondary and the 1st tertiary groups is to be 304 kc. All carrier frequencies of the tertiary groups are 440-kc multiples. The new system should provide for the possibility of simultaneous transmission of television programs and telephone conversations.

The frequency allotment for this new system, as recommended by the CCIF, differs from the frequency allotment adopted in the Soviet Union for the 8.5-mc coaxial system. However, joint operation of these systems is feasible at tertiary groups, by transfer to the 812 - 2044 kc tertiary group.

During the examination of this system there arose the question of the pertinency of altering the equipment types of radio-relay lines and, in particular, of altering their capacity from 240 to 300 (and multiple 300) channels so as to cause them to correspond to the basic tertiary group of the coaxial system.

The other questions to come under study were calculation of wideband systems on the basis of nominal standard circuits and a more precise postulation of these nominal standard circuits. It was resolved to assume that the primary or secondary channel groups connect with each other on the ends of a homogeneous segment according to the random-flight law.

With respect to standard coaxial and balanced-line cable circuits the general rule should be observed that the mean psophometric power of noise (during any hour) for any homogeneous-segment circuit should not exceed $3\ \mu\mu w$/km, proceeding from the assumption that the absolute power-level of the mean signal (speech currents plus signaling currents) during that hour equals -15 db at the point of the relative zero level.

Standardization of Coaxial Cables

The pulse method was recommended as the basic method for determining the degree

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of the nonhomogeneity of factory lengths of coaxial cables and of the assembled repeater sections.

The shape of the pulse for such cases is square of sine with a width of no more than 0.1 μsec at testing of factory lengths and no more than 0.2 μsec at testing of cables on assembled sections.

The echo attenuation (logarithm of the ratio of the amplitude of the basic pulse to the amplitude of the reflected pulse) was assumed equal to:

a) For factory length - 50 db for 100% measurements and 54 db for 95% measurements;

b) For repeater sections - 54 db.

During a session of the CCIF Investigating Commission No.3, the Soviet delegation described a new method for evaluating the nonhomogeneity of coaxial cables by measuring the frequency-weighted energy of reflected signals, and as a result the Commission posed for study the following new question: "Which Methods of Transient-Mode Measurements are most Descriptive of the Homogeneity of Factory Lengths and Repeater Sections of Coaxial Lines Designed for Transmission of Television?".

New Three-Channel Aerial Line Multiplexing System

The CCIF adopted as the basic three-channel system, a system operating in the frequency spectrum of 4 - 31 kc, with a channel bandwidth of 0.3 - 3.4 kc (cf. Green Book, Vol.III, pp.104 - 106).

During a session of the CCIF Investigating Commission No.3, it was found that the recommended system does not satisfy the requirements of a number of countries (including the USSR), since the frequency band up to 6 kc is allocated to other purposes (single-channel systems, official communications, superacoustic telegraphy).

In this connection, the following new question has been posed: "What Should be the Characteristics of a Three-Channel System that would Satisfy the Requirements of Government Administrations Rejecting as Inapplicable the Originally Recommended

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System?".

Appraisal of Quality of Transmission in the Presence of Noise

The CCIF Investigating Commission No.4 recommends that the so-called method of "opinions" be used for appraising the quality of transmission in the presence of noise. In this method, the subscribers themselves are polled as to the quality of transmission by the five-ball polling system. The attenuation of circuits connected between subscriber sets and the level of line noise are taken into account. Special Tables have been drafted for transmission.

Plan for Semiautomatic Telephone Network in Europe

The CCIF has approved a plan for organizing a semiautomatic telephone network in Europe, which was drafted in accordance with the proposals of various Government administrations. This plan should be materialized by the year 1960.

Through-switching centers are expected to be set up by 1960 in the following eight European cities: Belgrade, Stockholm, Frankfurt/Main, Copenhagen, Milan, Prague, Vienna, and Zurich.

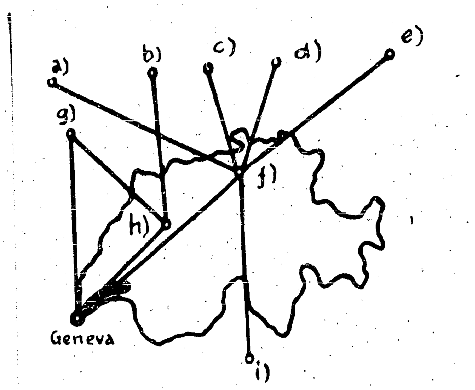

Fig.2

a) London; b) Brussels; c) Amsterdam; d) Stockholm; e) Frankfurt/Main; f) Zurich; g) Paris; h) Berne; i) Milan

The plan envisages the organization of direct semiautomatic connections between Moscow and the following cities: Sofia, Helsinki, Warsaw, Bucharest, Budapest, Prague, Belgrade, Tirana and Berlin. It also provides for the organization of through-switching connection with Vienna via Warsaw and with Athens via Belgrade.

Simplified CCIF-standardized single-frequency semiautomatic dialing equipment of the S type can be installed on all direct-connection offices, and the CCIF-

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standardized two-frequency equipment of the 2F type can be installed on all through offices.

To materialize this plan the ministries of communications and of radio-engineering industry should accelerate the development of simplified international semiautomatic equipment.

It is pertinent to note that semiautomatic dialing has begun to be developed in Europe. Figure 2 gives a schematic layout of the international automatic dialing system employed in Switzerland where the work of the international consultative committees on telephone and telegraph was conducted.

Standardization of the FM Voice-Frequency Telegraphy System

The discussion of this question indicated that the delegates of several countries had previously voiced their opposition to the FM system on the grounds that, compared with AM, it is more complicated and, principally, that it causes the voice-frequency telegraphy channels to be more sensitive to shifts in carrier frequencies.

At present, this principal obstacle has been overcome. Several methods for increasing the channel stability have been developed. These methods are: application of circuits for compensating frequency shifts up to ±30 cycles, to be effected individually in every receiver set; application of a control channel on a 300-cycle frequency, which is transmitted along r-f channels together with telegraph-channel frequencies and, upon reception, is equated to the frequency of the local 300-cycle oscillator.

The discussion ended in the adoption of a recommendation on the standardization of a 24-channel FM voice-frequency telegraphy system, operating at a velocity of 50 bauds and a 120-cycle interval between mean frequencies.

On motion by the Soviet delegation, the following new questions were posed for study: possibility of increasing the speed of phototelegraph transmission by applying phase correction of telephone channels and by applying unbalance-type transmis-

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sion of a single frequency sideband; question of the methods and equipment for automatic through-handling of phototelegrams. The discussion of the latter question revealed that a number of countries tend to employ a broad variety of "facsimile" equipment (simplified phototelegraph equipment installed directly in the premises of subscribers and serving to obtain photocopies of newspaper and periodical articles, information material, announcements, etc.) for the transmission of ordinary telegrams and documents.

This question was posed for study.

Organization of a Direct-Connection European Telegraph Network

In compliance with the proposals of various national administrations, the CCIF drafted a plan for the organization of a European direct-connection telegraph network and compiled a roster of the institutions in European countries to be included in that network. The European telegraph network will consist of both national systems and of systems especially isolated for this purpose.

It was resolved that this new service be called "Gentex".

Investigation of the Stability of Residual Attenuation in Telephone Channels of the European Network

The CCIF Investigating Commission No.9 analyzed the results of all five series of experiments on varying the residual channel attenuation in the European network, conducted in the years 1949 - 1956. The Commission established that the stability of this attenuation in the European network had increased and the number of individual fluctuations in this attenuation had decreased.

It was demonstrated that these fluctuations do not, as a rule, occur at night or during lunch or supper hours when the staffing of telephone offices is greatly reduced.

The sixth consecutive series of tests on the stability of residual attenuation in telephone channels is scheduled for 1957.

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During their sessions, the investigating commissions formulated new questions which should be explored in the next few years.

The CCIF Investigating Commission No.3 posed only one question concerning the tendency to a decrease in the established 0.8-neper value of residual attenuation in telephone channels; in the event of a positive reply, this would be supplemented by the question of what should be the new plan of the European and worldwide network from the viewpoint of transmission.

The CCIF Investigating Commission No.5 posed for study only one new question concerning a rational distribution of the frequency spectrum, the number of channels, and the area of application of the multichannel radio system, based on tropospheric and ionospheric radio-wave propagation.

The following new questions posed for study also deserve mention: characteristics of the international dial telephone systems; automation of the measurements of residual channel attenuation; utilization of "facsimile" service for transmission of telegrams and business documents; objective methods of appraising the quality of telephone transmission; characteristics of channels for the transmission of computing data (in connection with the widespread development of automatic-computer plants in the USA and Europe); protection of r-f channels from radio station interference; replacement of lead sheathing of cables and wires by thermoplastics; revision of telegraph rules; and many others.

In concluding, it should be pointed out that the work of the investigating commissions and plenary assemblies of the CCIF, CCIT, and CCITT was conducted in a businesslike manner: all questions were resolved without dispute, and the resolutions were adopted unanimously. This is further enhanced by the good organization of the meetings with respect to equipment and seating arrangements, rapid multigraphing of documents, and good servicing of the delegates. The activities of the commissions and assemblies were shared by 500 delegates representing 50 countries - members of the International Telecommunications Union.

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The activities of the Soviet delegation have increased considerably. In the last few years, numerous expert replies and expositions concerning the questions studied were presented. It would be desirable to have the Soviet radio and electric engineering industries participate even more actively in this important work in the future.

The sessions of the investigating commissions of the CCIF and CCIT, held in November - December of 1956 in Geneva, are of great technological interest, and their output of certain data, documents, and recommendations will be of use in the practical work of the Soviet Union with respect to the development of an international telephone and telegraph network.

In view of the great number of alterations and corrections introduced by the investigating commissions, the Plenary Assembly of the CCITT adopted the resolution of issuing new volumes of the Green Book (Vols. I, II, and III, new edition) in 1957.

Article received by the Editors 26 February 1957.

LETTER TO THE EDITOR

Considering that a comparison of the efficiency of various forms of communication is of great practical interest, I request that you print the following remarks relative to the comparison of some forms of communication conducted by Professor A.A.Kharkevich in his "Outline of General Communications Theory" (GITTL Moscow, 1955, Section 18, pp.51 - 55).

The calculation of approximate data on the amount of information that can be conveyed during unit time by means of various forms of communication has been presented by Professor A.A.Kharkevich in three Tables. In the first Table, the initial transmission parameters are selected and the following value is computed:

$$\frac{I}{T} = 2F \log_2 m,$$

where F is the width of the signal frequency spectrum;

m is the number of discernible gradations of the signal strength.

The second Table calculates the specific content of a signal, i.e., the amount of information per unit volume

$$\nu = \frac{T}{V},$$

where the volume of the signal is

$$V = FTH = FT \log_2 \frac{P}{P_n},$$

Here, P and P_n are the signal power and noise power, respectively; the dynamic range H for the case of transmission by the APM mode (under the condition that the probability of error due to interference in the form of white noise will be approximately 10^{-3}) will equal

$$H = \log_2 \frac{P}{P_n} \approx \log_2 7(m-1)(2m-1).$$

Finally, the third Table computes the signal content as expressed in the number of words per minute in relation to the signal volume, i.e., the following value:

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$$a = \frac{A\,wds/min}{FH}.$$

Noteworthy is the negligible efficiency of phototelegraph transmission indicated in the last column of that Table, as compared with the efficiency of telegraph transmission, considering that

$$\frac{a_{telegr.}}{a_{phototel.}} = \frac{0{,}33}{0{,}0019} = 175.$$

The above numerical result is in strong contrast to the customary representations about the efficiency of phototelegraph transmissions of text, not to speak of the transmission of more complex images. For instance, the number of words that would be needed for the transmission of the entire information contained in a chart with a great number of convolute curves, superscripts, etc., to a recipient who is totally unfamiliar with the chart would hardly coincide with the low content of such transmission as deducible from that first Table.

Let us examine some corrections which should be introduced into the calculations done according to that first Table, confining ourselves to the case examined by Professor Kharkevich. If a phototelegraph blank is filled with small typewritten characters through a single interval, the width of a letter, considering the spacings between the letters, equals (2.6 - 2.85) mm, while the height of the letter, considering the spacing between the lines, equals approximately 4.3 mm, and the spacing between words equals one letter, whereas N = 92 mean Russian words of 7.9 letters each can be placed per 1 square decimeter of the area of the blank.

A part of the blank (10 - 20%) is unproductively wasted, since the margins are stapled together, used for official notations, and so forth. Let us take this into account by introducing the coefficient α_1. The transmission occurs with pauses for the replacement of blanks. The duration of these pauses does not exceed 1 to 1.5 min, which, even for small-size blanks, reduces the productivity by up to 15% (α_2 = 0.85). With such data and at a linear velocity of scanning of v_{lin} = = 0.33 m/sec and a feed rate of δ = 0.266 mm, the operating speed of transmission

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equals

$$A = a_1 a_2 \frac{60N}{S} v_{lin} \delta =$$

$$= 0,8 \cdot 0,85 \frac{60 \cdot 82}{104} \cdot 330 \cdot 0,266 = 33 \text{ words/min},$$

while the frequency band equals

$$F \approx \frac{v_{lin}}{2\delta} = \frac{330}{2 \cdot 0,266} = 620 \text{ cycles}$$

A change in the scanning rate corresponds to a change in A and F; therefore, if it is assumed, as was done in Table 1, that $F = 3 \times 10^3$ cycles, the speed of transmission will be approximately 160 words/min instead of 60 words/min.

Table 1

Form of Communication	F cycles	m	$\log_2 m$	$\frac{1}{T}$	H	FH		A words/min	a
Telegraph (Baudot)	40	2	1	80	4.4	1.8×10^2	0.45	60	0.33
Phototelegraph	3×10^3	12	3.6	2.2×10^4	10.7	3.2×10^4	0.68	60	0.0019

Now let us resolve more accurately the question of the dynamic range, i.e., of the magnitude of m. In principle, only two gradations of optical density are necessary for transmission of text. For example, radio transmission is sometimes effected by applying the frequency-keying mode, as in telegraphy. Consequently, it can be assumed that m = 2.

In accordance with the fact that the frequency band selected for phototelegraph transmission is taken from the calculation of the passing capacity of only the first harmonic of the briefest photosignal pulses, the value of F for a Baudot multiplex-diode system telegraph should be taken as F = 24 cycles (46.7 bauds) and for a CT-35 type telegraph at F = 22 cycles (44 bauds). It is known that the following standards for speed of transmission have been established for a first-discharge telegraph operator at a communications load of over 75% and at work with a monitor also fulfilling auxiliary operations: for a Baudot telegraph – 1150 words/hr per simplex,

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and for CT-35 telegraph - 1600 words/hr, which yields A = 38 words/min for the Baudot and A = 27 words/min for CT-35. (We have though to consider that first-discharge telegraph operators are high-grade experts, and a considerable part of the operations is conducted by second-and third-discharge operators; therefore, the mean actual speeds of transmission will be still lower.)

The calculations done according to such corrected data are contained in Table 2.

Table 2

Form of Communication	F cycles	m	$\log_2 m$	$\frac{1}{T}$	H	FH	ν	A words/min	a
Baudot Telegraph	24	2	1	48	4.4	1.06×10^2	0.45	38	0.36
CT-35 Telegraph	22	2	1	44	4.4	0.97×10^2	0.45	27	0.28
Phototelegraph	620	2	1	1240	4.4	27.3×10^2	0.45	33	0.012

It can be concluded from the last column in Table 2 that

$$\frac{a_{telegr(Baudot)}}{a_{photot el.}} = \frac{0,36}{0,012} = 30,$$

$$\frac{a_{telegr(CT\text{-}35)}}{a_{photot el.}} = \frac{0,28}{0,012} = 20.$$

This result is well known and is obtained from the following simple ratio:

$$\frac{a_{telegr}}{a_{photot el}} = \frac{F_{photel.} A_{telegr.}}{F_{telegr.} A_{photot el.}}.$$

It is also well known that such a considerable excess has not yet been realized. For example, according to the standards of the Comite Consultatif International, a telegraph channel occupies (on taking account of the interval for channel filtering) a 120-cycle band. The allowable speeds for singleband phototelegraph transmission by a telephone channel are of the order of 0.7 - 0.8 m/sec. At these data, we have

$$\frac{a_{telegr.}}{a_{photot el.}} = \frac{(2\,700 - 300) \cdot 38}{120 \cdot 70} = 11.$$

In the event of a letter size smaller than typescript, or in the event of singleband transmission by a corrected communication channel, this ratio becomes even smaller (of the order of 4 - 6).

In evaluating the above-obtained results, it is pertinent to explore further the following two questions:

1. Table 2 indicates that the specific content (ν) is identical for any transmission.

This result is by no means unexpected, since at the accepted definition of

$$\nu = \frac{I}{V} = \frac{2FT\log_2 m}{FTH} = \frac{2\log_2 m}{H} = \frac{2\log_2 m}{\log_2 \frac{P}{P_n}},$$

the coefficient ν is a function of the levels only and cannot characterize the comparative information value of various transmissions if these transmissions are all effected on identical levels. Therefore, the footnote to Table 2 (p.53) in Professor Kharkevich's book: "we should have obtained the highest specific content for telegraphy, but we did not, since we cited practical figures for the spectrum width but computed the excess of the signal by theoretical formulas" is not a sufficiently precise footnote.

2. The possibility of conducting calculations according to Table 1 becomes quite doubtful when the value of phototelegraph information is to be calculated not according to the number of words but according to some other standards. Even at black-and-white transmission it is possible to identify on the blank the characteristics of individual handwriting and of other notations on diagrams, sketches, complex circuit schematics, etc. At halftone transmission this is supplemented by information on brightness resulting in the creation of a visual image which sometimes is almost impossible to be described by words. Apparently the concept of the "content of phototelegraph transmission" has to be further evolved and more precisely defined, if only by analogy to the fact that the concept of speech transmission differentiates between semantic content, voice characteristics, etc.

B.Kisel'gof
Cand. Engl.Sci.

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REGARDING THE LETTER BY B.Z.KISEL'GOF

The tentative figures I cited were not designed for practical engineering calculations. It may be that the figures cited by the author of the letter are more exact and detailed than mine. However, my qualitative conclusions remain unaffected. One of these conclusions is that the possibilities of phototelegraphy are very poorly exploited in the transmission of printed text. I do not think that this quite explicit postulate could be contested, and it is hardly worthwhile to examine (as was done by the author of the letter) black-and-white phototelegraphy for the purpose of weakening the effect of the comparison with ordinary telegraphy, considering that the actually used phototelegraph systems are of the halftone type.

Neither should the matter be obscured by the mention of "a visual image which sometimes is almost impossible to be described by words". This may perhaps be a problem to a poet but not to the communications engineer who, on the basis of the fundamental concepts of the information theory, will say that the amount of information in an image does not exceed n log m, where n is number of elements of dissociation and m is number of brightness gradations.

A.Kharkevich

BOOKS IN 1957

Concise information on the technical literature scheduled for publication in 1957 which, in the Editors' opinion, is of interest to Elektrosvyaz' readers.

Svyaz'izdat Publishing House

Smirnov,V.A.: Foundations of Ultrashort Wave Radio Communications. Edition: 15,000 copies.

Ayzenberg,G.Z.: Ultrashort Wave Antennas. Edition: 17,000 copies.

Orlovskiy,E.L.: Theoretical Foundations of Phototelegraphy. Edition: 5,000 copies.

Chistyakov,N.I., Sidorov,V.M., and Mel'nikov,V.S.: Radio-Receiving Devices. Edition: 25,000 copies.

Iodko,Ye.K.: The Economics and Organization of Radio Communication and Broadcasting Enterprises. Edition: 10,000 copies.

Koshcheyev,I.A.: Foundations of the Theory of Electrocommunications. Part III, "Nonlinear Systems". Edition: 15,000 copies.

Krize,S.N.: Audio-Frequency Amplifiers. Edition: 25,000 copies.

Solov'yeva,A.G.: Foundations of Telephony and Attended Telephone Offices. Edition: 15,000 copies.

Kristal'nyy, V.S., Falunin,A.F., and Ivanova,A.A.: Toll Telephone Offices. Edition: 10,000 copies.

Klykov,S.I.: Telegraphy. Part III, Phototelegraph Communications. Edition: 10,000 copies.

Komarov,B.S.: The Power Supply of Wire Communication Enterprises. Edition: 15,000 copies.

Team of Authors: Engineering-Technical Manual on Electrocommunications. Radiosvyaz', Issue No.8. Edition: 20,000 copies.

Efimkin,V.I., Minenko,Yu.G., et al: Color Television. Edition: 2000 copies.

Il'ina,N.N., and Yulovskiy,P.V.: Short-Wave Radio Transmitting Devices. Edition: 2000 copies.

Tarakanova,M.S., Amarantov,V.N., et al: Supersonic Telegraphy Equipment NT-ChM-4. Edition: 6000 copies.

Polyak,M.U.: The KRR-30/60 Short-Distance High-Frequency Telephone System. Edition: 5000 copies.

Voznesenskiy,B.N., and Zaynchkovskiy,Ye.A., et al: Semiautomatic Toll Telephone Communications Equipment. Edition: 6000 copies.

Communications Engineering Abroad. Translations of Articles on Radio Relay Antennas

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and Masts. Edition: 4000 copies.

Communications Engineering Abroad. Translations of Articles on Phototelegraph Communications. Edition: 5000 copies.

Authors Collective: Communications in the Land of Socialism (in honor of the 40th anniversary of the Great October Socialist Revolution). Edition: 20,000 copies.

ERRATUM

In S.M.Gerasimov's article on "The Investigation of the Self-Oscillatory Mode in a Junction-Transistor Oscillator", published in No.3 of Elektrosvyaz', the author showed incorrectly the slope of curve x_{eo} in Fig.4 (p.26). Actually, at $\varphi_{dr} = \frac{\pi}{2}$ the x_{eo} should equal $-\frac{1}{S_{mean}}$ and at $\varphi_{dr} = \frac{3\pi}{2}$ the x_{eo} should equal $\frac{1}{S_{mean}}$. Figure 4 is an illustrative one. The conclusions and formulas cited in that article are all correct.

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CALCULATION OF A COMMUNICATION LINE WITH LONG-DISTANCE TROPOSPHERIC SCATTERING OF ULTRASHORT WAVES

The utilization of long-distance tropospheric scattering of ultrashort waves, combined with the use of high-power transmitters, sensitive receivers, and spaced receiving antennas, makes it possible to realize wideband communication at large distances between stations.

The aim of the calculation of such a communication line is to consider the characteristics of long-distance scattering for the purpose of selecting equipment parameters of a type that will ensure a certain reliability of communication.

Here it is necessary to take account of the magnitude of attenuation of the signal power during propagation in free space and beyond the line of sight, the total gain of transmitting and receiving antennas, the loss of antenna gain due to scattering, power losses in feeders and in connecting wirings of transmitter and receiver, the noise factor of the receiver, the established signal-to-noise ratio, etc.

On considering all these data the following power (in decibels per watt) is necessary for ensuring communication on a route with a length of d during 50% of the time:

$$P_m = L_{fs} + L_{oh} - G_s + L_c + L_l + F_n + (s/n) + 10 \lg B - 204, \qquad (1)$$

where $L_{fs} = 22 + 20 \log (d/\lambda)$ is the signal attenuation in free space;

L_{oh} is mean value of attenuation beyond the line of sight;

G_s is total gain of the transmitting and receiving antennas;

L_s is the antenna-gain decrement determinable from a graph, depending on the relationship to the width of the antenna lobes and the angular distance between stations, at an effective earth's radius of 8000 km;

L_l is the total loss in feeders and connections, approximately equal to

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$5 \log f_{mc} - 10$;

F_n is the noise factor of the receiver, approximately equal to $3.6 \log f_{mc} - 1.9$ (for $f \geq 100$ mc);

s/n is the excess of the signal over the noise level, at a given form of communication (from 10 db for a printing telegraph to 40 db for high-quality television);

$10 \log B - 204$ (where B is the intermediate frequency band of the receiver in cycles) takes account of the power of the thermal noise of the antenna at the receiver input for $T = 289°K$ (+15°C).

Equation (1) takes the mean values of signal attenuation beyond the line of sight L_{oh} and computes the power values ensuring communication during 50% of the time.

To ensure communication during larger time intervals (90%, 99%, and more) it is necessary to compensate the interference of slow and rapid fading by using greater power. Statistics on slow fadings makes it possible to determine the depth of such fading for different distances in relation to the mean signal level, depending on the percentage of time of the signal excess. The depth of slow fading decreases at an increase in length of the path. For example, at 99% reliability, 11 db should be added to the mean attenuation for 300 km and only 5 db should be added for 650 km. The interference induced by slow fading usually does not decrease at spaced reception.

Rapid fading is considered in an analogous manner, but here the value of the compensating power depends on the specific method of spaced reception.

To ensure reliability of communication during q% of the time, the following power is necessary:

$$P_q = P_m + ML_q + MS_q, \qquad (2)$$

where P_m is determined from eq.(1);

ML_q is the power for compensating the slow fading;

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MS_q is the power for compensating the rapid fading;

The values of ML_q and MS_q are determined from the graphs of fading distribution, taking account of the specific method of spaced reception used.

Determination of the mean value of attenuation beyond the line of sight, L_{oh} is preferably to be done according to empirical formulas of the following type:

$$L_{oh} = A + B(d - d_0), \qquad (3)$$

where A is the attenuation beyond the line of sight at $d = d_0$ (d_0 is usually 100 miles), and B is the rate of attenuation.

The B constant depends on the mean refracting properties of the atmosphere along a given path.

Measurements indicate that a value of B = 0.13 db/mile is typical for overland routes in the northern part of the USA and in Southern Canada. In the subtropics and on overseas routes, the value is

B = 0.11 db/mile under dry arctic conditions:

B = 0.15 db/mile.

At B = 0.13 db/mile, A = 57 db. For other B values, the magnitude of A must still be precisely defined.

Equations (1) and (2) can be applied not only to a determination of the required power but also to a determination of communication-line parameters (at a given transmitter power) such as antenna gain, receiver sensitivity, or some other appropriate combination of parameters of designed or applied equipment.

The same article also tabulates the equipment parameters for frequencies of 400 and 2000 mc and distances from 100 to 600 miles at various degrees of communication reliability.

(Electronics, Buyers Guide Issue, 1956, R = 18).

A TRANSISTORIZED SINGLE-CHANNEL HIGH-FREQUENCY SYSTEM

The "Telettra" Co. has recently completed the development and started series production of equipment for single-channel systems for multiplexing of cable and aerial lines of the TP system.

This equipment makes use of the conventional methods of frequency conversion; the currents of one of the sidebands are transmitted to a line. The terminal equipment was so designed as to realize either reception of the upper sideband and transmission of the lower sideband, or vice versa. The effectively transmitted frequency band of a single channel ranges from 200 to 2000 cycles. There are five versions of the distribution of the line spectrum of the system in a band of 4.5 to 29.5 kc, so that up to five single-channel TP-system sets can independently operate on a single circuit. The values of the virtual carrier frequencies of the equipment are: 7, 12, 17, 22, and 27 kc.

The equipment uses semiconductor triodes, which makes it possible to reduce the power demand to its minimum and to reduce the dimensions of the equipment. The magnitude of the current consumed by a single terminal equipment set has been successfully reduced to 10 ma(at a voltage of 24 v). The gain of transmitting and receiving amplifiers amounts to approximately 30 db which ensures compensation of the attenuation on an aerial-line segment of 100 to 150 km length (depending on the variant of line spectrum). At ± 30% fluctuations in the feed voltages, the magnitude of the amplifier gain remains nearly unchanged. The stability of the carrier frequencies amounts to approximately 20×10^{-6} (per $1^{\circ}C$), which is achieved by using quartz stabilization of the oscillator frequency (in every single-channel set). The divergence of the carrier frequencies at the terminal stations does not exceed 2 cycles.

The residual attenuation of the channels of the system at 800-cycle frequency was assumed equal to 6 db. The increase in the residual attenuation at 200-cycle

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frequency relative to that value does not exceed 2.5 db, and at 2000-cycle frequency this difference is below or equal to 1.5 db. At variations in temperature within the range of (-20°C) to (+45°C) the order of fluctuations in the value of residual attenuation equals several tenths of a decibel.

The value of transient attenuation on the near and far ends is not less than 65 db. The transmission of dialing pulses is effected by a relay system. Considering that the system can operate on cable lines as well as on aerial lines, the directional filters are designed for load resistances of 600 or 150 ohms.

With respect to design, a single-channel set is constructed in the form of a hermetically closed box measuring 390 × 280 × 120 mm; the box has a removable lid covering a face panel; a pointer indicator and a few switching elements are located on the panel.

The use of this TP-type equipment is particularly interesting under field conditions and in cases where several telephone connections must be rapidly organized for a comparatively short period of time, if the requirements made on the qualitative indexes of the channels are not too high.

DEVELOPMENT OF ELECTROCOMMUNICATIONS IN ITALY

Because of the events of World War II, the Italian long-distance communication system, consisting of underground cables with an overall length of 3500 km and a number of multiplex aerial main lines, was almost totally destroyed.

During the years 1949 and 1950 and the first half of 1951, considerable work was done on restoring the long-distance line system on a modern technological basis. The backbone of the newly created cable network was a main line intersecting the Italian Peninsula from north to south and extending toward Sicily. Several other main lines branched off from this backbone in the dense-traffic areas of northern Italy. The aerial line network differed from the cable network by its greater density, especially in the northern and central regions of the Peninsula. In these

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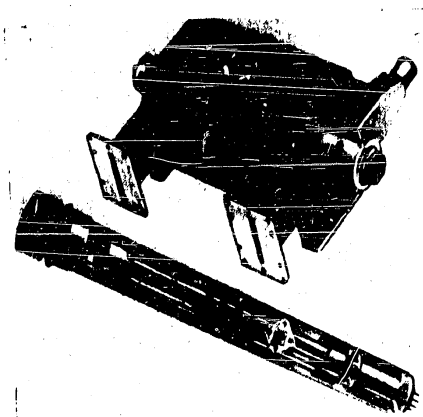
regions, the majority of the main lines multiplexed by multichannel systems (systems with more than 12 channels) was concentrated.

In this connection, a plan for creating a new wideband long-distance network, formed by coaxial and radio-relay lines, was drafted. The overall length of the new coaxial network exceeds 4500 km and its cost is 55 billion Lira. The construction work of this network was scheduled for completion by 1957: by the middle of 1956, main lines with an overall length of 1000 km were ready for operation. The plan envisages the construction of two long-distance main lines running from north to south (along the Tyrrhenian and Adriatic seacoasts) and four transverse main lines. Several other such lines are being constructed in Sicily. The new main-line network will have outlets toward the main-line networks of France, Switzerland, Austria, Yugoslavia, and Tunis.

It is also intended to take a number of measures for modernization and development of telephone communications; in particular, broad development of subscriber telegraphy is scheduled.

A NEW TRAVELING-WAVE TUBE

The German firm Telefunken has developed and initiated the series production of traveling-wave tubes of the TL4 and TL6 type, designed for power amplification within the 2000 - 4000 mc range at a bandwidth of up to 30 mc. These tubes are used in radio relay equipment for transmission of up to 600 telephone conversations or one television program in a single trunk.



Traveling-wave tubes can also be used in measurement engineering for power amplification within a wide frequency band; for example, in the terminal stage of a variable-tuned transmitter, for qualitative

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measurements of standing waves, for tuning of resonance [illegible] during fil-
ter passbands, controlling the tuning of antennas, and the like.

A traveling-wave tube (TWT) amplifier (Fig.1) consists of a traveling-wave tube, a tube base, and the coils of an electromagnet for focusing the electron beam. The base contains, in addition, a waveguide input and output and tube leads. The concentration of the beams is effected by a magnetic field. The evacuated glass envelope of the TWT contains an electron gun, a helix, and a collector.

Specifications of TWT of the TL4 and TL6 Types

The external parameters of both tube types are identical. Tube length is 265 mm, and diameter of the glass envelope is 27 mm.

Operational Parameters of the Tubes

	Tube TL4	Tube TL6
Frequency	2000 mc	4000 mc
Voltage at plate	–	-80 – +80 v
Plate voltage	–	650 v
Voltage along the helix	–	1250 – 1150 v
Voltage on the collector	–	1300 v
Collector current	–	20 ma
Magnetic field	–	400 – 500 cycles
Maximum power at the output end . .	950 – 760 mw	4 – 1.4 w
Gain	60 db	25 – 36 db
Noise factor	12 – 4 db	27 db
Bandwidth (matching m 0.97)	35 – 45 mc	30 mc

(Radioschau, No.10, 1956, p.268).

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Class $21a^1$, 32_{34}. No.104456. Yu.V.Ivanov and S.I.Katayev. Single-Beam Color Television Picture Tube.

Class $21a^1$, 33_{40}. No.104425. G.V.Braude. Method of Correcting the Aperture Distortions Induced in Television Signals by Camera Tubes and Picture Tubes.

Class $21a^1$, 32_{35}. No.104426. K.A.Fedorova. Electron-Beam Camera Tube.

Class $21a^1$, 32_{35}. No.104424. G.V.Braude. Television Camera Tube.

Class $21a^1$, 32_{31}. No.104588. A.G.Lapuk. Signal Plate for Camera Tubes.

Class $21a^1$, 34_{31}. No.104592. V.V.Odnol'ko. Device for Reproduction of Color-Television Images.

Proposal for a device for reproducing color-television images with dotwise color switching and with a graticulated-screen tube; the tube screen receives projections from a conductor together with vertical strips of different-colored luminophores, thus forming a projected raster serving for the reception of voltage pulses; these pulses, on being mixed with the video signal, ensure the opening of the beam during the time intervals when they coincide with the luminophore, corresponding to the color of the transmitted field; the pulses shut off the beam during transverse transit over the vertical strips of luminophores of other colors.

Class $21a^1$, 34_{53}. No.104594. V.I.Peskovskiy. A Method of Illustrated Radio Broadcasting.

Proposal for a method of illustrated radio broadcasting involving the transmission of fixed images by electronic-television methods with a number of frames per second high enough to ensure the possibility of operation in the medium- and long-wave ranges, with the images received on a cathode-ray tube with long afterglow. Accordingly, for the purpose of reducing the number of radio transmitters used and the frequency band occupied, and also for the purpose of ensuring the possibility of using the existing radio receivers for reception, the transmission of the audio part of a program and of illustrations is effected consecutively by a single radio transmitter on a single carrier frequency. To eliminate mutual interference between audio and video channels, the video reception channel is automatically disconnected by control signals from the transmitter during transmission of the audio part of the program, and the audio reception channel is analogously disconnected during transmission of the video part of the program.

It is proposed that video synchronization signals be used in the capacity of signals for disconnecting the audio reception channel.

To disconnect the video reception channel after the transmitted image frame is completed, it is proposed to close with a beam the circuit of the supplementary electrode of the cathode-ray tube of the receiver, used for switching to the audio channel.

Class $21a^1$, 32_{04}. No.105255. G.N.Bogorodskiy, I.M.Baynberg, I.L.Koblents, A.D.Kirpicheva, R.A.Kudryavtsev. Phototelegraph Transmitter with Junction-Type Scanning.

Class $21a^1$, 7_{01}. No.105187. P.G.Tager. Trigger Device.

Class $21a^1$, 32_{11}. No.105189. P.G.Tager. Method for Through-Handling of Phototelegrams.

It is proposed that, in through-handling of phototelegrams by magnetic recording, such magnetic recording of the received phototelegrams be utilized for indirect control (i.e., without converting the phototelegrams into images) of the transmitting device, for the purpose of reducing distortions.

Class $21a^1$, 32_{35}. No.105014. S.B.Gurevich and R.Ye.Bykov. Method for Inserting the Level of Black into the Signal Formed by the Television Tubes Transmitting the Mean Component of the Signal.

Class $21a^1$, 33_{40}. No.105090. G.V.Braude. Method of Correcting Aperture Distortions.

Class $21a^1$, 33_{40}. No.105181. G.V.Braude. Device for Correction of Aperture

Distortions in Television Receivers.

Class $21a^3$, 16_{01}. No.104372. V.N.Arapov. Monopolar Push-Button Dial.

Class $21a^3$, 16_{32}. No.104586. G.I.Zernov and A.V.Lebedev. Device for Suppression of Radio Noise Arising in the Pulse Contact of a Telephone-Receiver Dial.

Class $21a^3$, 17_{20}. No.104593. A.V.Lebedev. Relay Pulse Oscillator.

A relay pulse oscillator utilizing the charge of the capacitor with voltage pulses for obtaining a greater duration of intervals between pulses and greater degree of control is based on a circuit utilizing a method of delaying the relay operation by means of the pulse charge of the capacitor.

Class $21a^3$, 66_{01}. No.104371. M.A.Pimenov. Check Point for the Step-by-Step Dial System.

For the purpose of automating the efficiency testing of multiple fields it is proposed to use three relays connected in series to cable conductors a, b, and c through intermediate decade-step selectors whose brushes automatically traverse all decades of a field and are triggered at the disruption of any of these conductors or of the connection between them or between them and the main unit.

FOREIGN PATENTS

From the Editors: With respect to the concise information published here on some foreign patents in the field of antenna engineering, the readers can obtain photostats (in the original languages, with sketches) of these patents on applying in person or sending a money order to the All-Union Patent Engineering Library of the Committee for Inventions and Discoveries at the Council of Ministers of the USSR.

Address of the Library: Moskva-Tsentr, Proyezd Serova 4, Pod'yezd 7a.

Phone: B8-64-52.

Photostats of patents can also be obtained through the purchasing office of the All-Union Institute of Scientific and Technical Information serving the USSR Academy of Sciences and the Gostekhnika.

Address of the Institute: Moskva D-219, Baltiyskiy Pos. d. 42b.

Phone: D7-00-10, dob.51.

US Patent, Class 250-33, No.2 724 052, 15.11.55. J.M.Boyer (Douglas Aircraft Company). Wideband Radio Antennas.

The invention describes a wideband antenna consisting of a rotating body designed for the radiation of waves of lower frequencies and having a circular radiation pattern in the horizontal plane. To prevent major changes in the radiation pattern in the horizontal plane during transition to higher frequencies, the surface of the antenna is topped with a quarter-wave groove, bisecting the surface into two parts with mutual synphase excitation thus yielding a vertical pattern that differs little from the pattern at lower frequencies. Diagrams of several variants of the invention are included.

Patent German Federal Republic, Class 21a4, 74, No.936811, 22.12.55. H.Larsen (Siemens and Halske A.-G.) A Horn Device for the Excitation and Reception of Electromagnetic Waves Propagating Along Surface-Wave Lines.

The diameter of the transition horns should be twice as large as the radius of the surface-wave field. Field-concentrating transient devices are applied to reduce the dimensions of the horn. These devices are installed between the horn and the line and are represented by dielectric bodies of streamlined form. The form and cross-sectional area of these bodies and of the adjacent segments of the conductor are so selected as to obtain at the same time a matching of the characteristic impedance of the horn and of the single line. Diagrams of several variants of the invention are included.

Canadian Patent No.512143, 19.04.55. W.E.Kock (Western Electric Company, USA). Aperture Antennas.

Description of a horn antenna whose input aperture is divided into cells by metal partitions. The size of the cells is much below $\lambda/2$. The partitions are so distributed that greater energies are directed toward the peripheral cells than toward the central cells. Phase shifters are installed in some cells to control the form of the radiation pattern.

Patent German Federal Republic, Class $21a^4$, 46_{02}. No.936400, 15.12.55. G.Piefke (Siemens and Halske A.-G.) Funnel and Horn Radiator Arrays for Short and Very Short Electromagnetic Waves.

Description of several variants exponential horns of circular cross section, fed by waveguides or coaxial lines. An equation is given for the construction of the longitudinal profile of the horn and for determination of its basic parameters. Horn versions with a dielectric rod insert, forming an extension of the internal conductor of a coaxial line, are described.

Canadian Patent No.516922, 27.09.55. F.V.Gosline (Pioneer Specialty Company, USA). Retractable Aerial.

Description of a retractable bayonet antenna for radio broadcasting consisting of telescoping hollow rods. The antenna can be placed in operating condition by means of a flexible rotary coupling.

US Patent, Class 250-33. 65, No.2 714 659, 02.08.55. E.Johnson and R.F.Kolar (Radio Corporation of America) Broad-Band Unidirectional Antenna.

Description of an antenna system consisting of a corner reflector formed by a number of metal rods and a broad-band biconical dipole, with its axis directed parallel to the rods of the reflector.

US Patent, Class 343-843, No.2 732 551, 24.01.56. J.D.Kraus (Battelle Development Corp.). Spherical Cage Antenna.

Description of an ultrashort-wave antenna system consisting of metal rings (with a diameter of $\sim \frac{\lambda}{4}$) in a uniform array similar to the meridians of the terrestrial globe about an axis constituted by a coaxial feeder line and its extension. Half of the rings, in the lower part of this arrangement, are attached to the external conductor of the coaxial line, while the other half are attached to the internal conductor which projects upward by a length of $\sim \frac{\lambda}{4}$. The first group of rings is isolated from the internal conductor and the second group, from the external conductor of the coaxial line. This antenna is characterized by its small linear dimensions ($\sim \frac{\lambda}{4}$), considerable bandwidth (up to 10%), uniform radiation pattern in the horizontal plane approximating a circle (in view of the absence of radiation in a vertical direction), some gain over the quarter-wave rod (1 - 2 db), the possibility of varying the value of input resistance of the antenna by varying the number and diameters of the rings, and the possibility of varying the radiation pattern of the antenna by inclining its axis relative to the horizontal plane.

US Patent, Class 250-33, No.2 719 919, 04.10.55. K.Enslein (Stromberg-Carlson Company). Built-In Antenna System.

Description of the antenna system of a television set designed for reception of transmissions on two channels with bands of 54 - 88 and 174 - 216 mc, respectively. The antenna is formed by a system of U-shape dipoles located above the chassis of the television set and capable of rotating in the horizontal plane. The dipoles are connected to the input of the receiver by a flexible two-wire line. The elements of the antenna and of line length are so selected as to eliminate antenna switching when tuning to another channel. The description includes diagrams of dipoles and of their connection and an overall view of the television set with antenna. Also described are all parts of the antenna and the dimensions of the dipoles.

Australian Patent No.165043, 22.09.55. (N.V.Philips' Gloeilampenfabrieken). Antenna with Loop and Ferrite Rod.

Description of a radio receiver antenna consisting of a small loop placed inside the radio receiver and a r-f transformer with ferrite-rod core. The axis of the core rod forms a right angle with the plane of the loop.

US Patent, Class 250-33, No. 2715 184. E.C.Cork (Electric and Musical Industries, Ltd., Great Britain). Aerials.

Description of a rod antenna, with several nodes along the rod, constituted by the capacitive load of the antenna. The largest bulge is in the middle of the rod. One end of the antenna is connected to a receiver. By varying the position of the nodes it is possible to vary the input resistance of the antenna. The passband of the antenna is 8% of its working frequency.

Canadian Patent No.516559, 13.09.55. W.A.Cummings (National Research Council). Non-Resonant Antenna.

Description of an antenna consisting of two zigzag conductors arranged stepwise along the antenna axis in such a manner that the angle of one conductor is included between two angles of the other.

US Patent, Class 250-33, No.2 712 602, 05.07.55. E.G.Hallen (Telefonaktiebolaget L.M.Ericsson, Sweden). Reflection-free Antenna.

Description of a device consisting of set of capacitances and serving to reduce reflection from the ends of a clamp-driven antenna. The device suppresses the reflected wave interfering with the current wave forming in the antenna.

US Patent, Class 250-33, 67, No.2 719 922, 04.10.55. R.S.Johnson (Zenith Radio Corp.). Core-Tuned Loop.

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Description of a loop with convenient control of the magnitude of inductance, thus facilitating the coupling between antenna circuits and heterodyne. The loop is wound with insulated wire, with the adjacent turns lying in parallel planes. The loop circuit has the form of a rectangle and is so shaped as to form a cylindrical depression. The corresponding part of every turn has a form permitting a nearly complete encompassing of the transverse cross-sectional area of the cylinder. Inside the resultant cavity a complete cylinder is placed, in which the tuned ferromagnetic core is free to move.

Patent German Democratic Republic, Class 21a^4, 46$_{02}$, No.10503, 08.10.55. W.Kriebel. Exciters for Meter- and Decimeter-Range Antennas (Antennas in the Form of a Concave Spherical Reflector or Rectangular Horn).

Description of exciter antennas consisting of metal spirals wound on a radial frame or stretched rubber cords. The spirals form a grid with cells whose dimensions are determined by the wavelength. Antennas of this type can be easily disassembled, which facilitates their transport.

Canadian Patent No.511984, 19.04.55. J.R.Winegard. TV Antenna Array and Director Therefor.

Description of a wideband TV receiving antenna of the director type containing two dual active vibrators for the maximum and minimum frequencies of the band and several alternately-arrayed passive directors for these two extreme frequencies.

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BIBLIOGRAPHICAL AND ABSTRACT MATERIAL ON ELECTROCOMMUNICATIONS

The present era is characterized by the colossal strides of science and engineering which, in particular, is manifested in the tremendous increase in the quantity of the published scientific and technical literature.

As of the present, no expert is physically able to even skim over the entire mass of the literature being published on his own specialty. However, our scientific research workers must keep abreast of all new developments in their particular specialties.

The only way out of this situation apparently is a regular examination of bibliographical and reference publications on these specialties.

The present article supplies a brief survey of Soviet and foreign bibliographical and reference sources on electrocommunications.

Public Records or State Bibliography

Bibliographical information on newly published Soviet literature on any specialty and, in particular, on problems of electrocommunications, is published in the weekly issues of all All-Union Book Board's "Knizhnaya Letopis" (Book Annals) and "Letopis Zhurnal'nykh Statey" (Annals of Periodical Literature).

Books, brochures, catalogs, test and measurement standards, patents, programs, instructions, methodological works, and other literature on problems of communications, are included in Chapter XIX ("Communications") of "Knizhnaya Letopis". It is also advisable to peruse the literature listings included in Chapter XVI on "Engineering Industry", and Chapter XV on "Physical and Mathematical Sciences".

It is worthwhile to peruse the weekly mass bibliographical bulletin of the All-Union Book Board, "Novye Knigi" (New Books) which contains systematic lists of new books for all branches of knowledge (inclusive of belletristic literature).

The articles and documentary materials published in Russian-language journals, transactions, proceedings, and periodic anthologies issued in the USSR are listed in "Letopis Zhurnal'nykh Statey".

Once every three months supplements to these annals are published: a name index (authors, editors, illustrations), and a geographical index.

Reference Information. The publications of the All-Union Book Board comprise only the literature issued in the USSR; moreover, no annotations are provided, i.e., there are no brief explanatory notes on the literature.

The reference journals issued by the All-Union Institute for Scientific and Technical Information of the USSR, Gostekhnika, and USSR Academy of Sciences differ from the above-mentioned bibliographical publications in that they selectively comprise not only Soviet but also foreign literature and, as a rule, furnish annotations or a brief abstract for every title, aside from secondary materials.

"Elektrotekhnika" Reference Journal. Literature on electrocommunications is cited in the Sections "Radiotekhnika" and "Elektrosvyaz'" of the Elektrotekhnika reference journal.

The "Radiotekhnika" Section comprises literature on the following group of problems: theoretical radio engineering; technology of radio parts and equipment; radio stations and radio transmitters; radio broadcasting; antennas and radio-wave propagation; radio receivers and amplifiers; pulse engineering; television; radar and radio navigation; ultrashort-wave and SHF engineering; ultrasonic engineering and hydrolocation; electroacoustics and sound recording.

The "Elektrosvyaz'" Section classifies material on the following topics: information theory; long-distance communications; telephony; telegraphy; communication lines and cables; noise and its prevention.

Electron tubes, electron-beam devices, transistors, etc. are reviewed in the "Elektronika" Section.

The literature on primary elements and accumulators is surveyed in the second Section, "Chemical Sources of Energy", of this journal. The Section "Electrotechnical Materials" surveys the literature on magnetic and electricinsulating materials, conductors, semiconductors, and capacitors.

All forms of remote control and telemetering have a theoretical basis in common with electrocommunications, since they are based on the transmission and reception of various kinds of electric signals. Therefore, electrocommunication workers may be directly interested in the literature surveyed in the Section "Automation and Telemechanics" and also in the Subsections surveying literature on automation and telemechanics.

Reference Journals "Fizika" and "Matematika". The theoretical literature discussing the physical and mathematical foundations of electrocommunications is correspondingly collated in the reference journals "Fizika" and "Matematika". Aside from papers of a general nature that are of interest in the solving of electrocommunication problems, these journals list annotated periodical literature relating to the theory and practice of electrocommunications: for instance, information theory, algebra of relay systems, computers, etc.

Bibliographical Journals and Bulletins. Less extensive but more topically oriented are the bibliographical journals and guides to the literature published by various scientific-research institutes and institutions of higher education.

Of considerable interest is the "Annotated Guide to Radioelectronics Literature" whose publication was started twelve years ago by the Sovetskoye Radio Publishing House. This guide is issued twice a month. As of 1957, it will comprise not only foreign but also domestic literature.

The Bureau of Scientific and Technical Information of the Ministry of the Radio Engineering Industry publishes a guide to technical literature which has the special value of including data on departmental publications.

A guide to technical literature is also published by the Central Bureau of Technical Information of the Ministry of the Electrical Industry.

As of 1954, the State Science Library has begun to publish quarterly a Bibliographical Index of Current Literature. One of the series of this Index surveys Soviet literature on problems of energetics and power industry.

In particular, an electrocommunication specialist will find, in this Index, material on the following problems: history of radio engineering and wire communications; radio systems and lines; broadcast engineering; radio networks; applied acoustics; image transmission; wire communications; general and theoretical problems; aerial and cable lines and wire-communication networks; telegraph equipment and offices; telephone equipment and offices; high-frequency, multichannel, and wire communications; long-distance communications; power sources for communication media utilization of new sources and forms of energy.

Foreign Sources of Current Bibliography.

A brief survey of Soviet sources of current bibliography was given above. Let us enumerate briefly the foreign sources of current bibliography.

Poland issues quarterly a "Survey of Polish Technical Literature" which includes not only technical but also scientific literature as well as critical reports on articles and books published outside Poland.

In Hungary there is the quarterly "Survey of the Hungarian Technical Press", which contains a special Section on "Electrical Engineering and Communication Engineering" including, in particular, a listing of the tables of contents of the Hungarian journal "Communication Engineering" (Magyar Hiradastechnika).

Czechoslovakia issues a bibliographical guide to periodical and book literature.

The German Democratic Republic issues a monthly reference journal "Technisches Zentralblatt. Abteilung Elektrotechnik". The principal topics of the journal concern electrical engineering, but radio-electronic literature is also surveyed.

In the Capitalist countries, the following technical journals with extensive bibliographical sections on radio-engineering and electrocommunication literature are of interest: the British journal "Wireless Engineer" (as of 1957, "Electronics and Ra American journal "Proceedings of the Institute of Radio Engineers" (PIRE), which once a month reprints the bibliography cited in "Wireless Engineer", the French journal "Annales des Telecommunications", which surveys much periodical literature on long-distance, wire, and radio communications, and finally the West German journal "ETZ".

The following are outstanding among the foreign reference and bibliographical journals on electric engineering, radio engineering and electrocommunications: British, "Electrical Engineering Abstracts. Section B. of Science Abstracts" and German, "Physikalische Berichte. Electronische Rundschau".

Retrospective Bibliography

Beside current bibliography there exists the so-called retrospective or cumulative bibliography, which is very useful whenever it is necessary to survey literature published over a long period of time.

The foreign sources of retrospective bibliography include the subject guide published annually in March by the British journal "Electronics and Radio Engineer" ("Wireless Engineer"): "Index to Abstracts and References", which is reprinted annually in April by the American journal PIRE.

Up to 1954, the April issues of PIRE contained a general annual survey of radio-engineering literature. The bibliographical yearbook "Engineering Index", issued in June - July, is also of interest.

Such bibliography is also contained in the book and periodical bibliographies cited in the Great Soviet Encyclopedia, and in the surveys published in journals and issued in the form of separate books (for example: Scientific Literature on Transistors. Bibliography 1929 - 1952. Moscow - Leningrad, USSR Academy of Sciences Press, 1955, 631 pages).

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