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TRANSLATION

ELECTRIC COMMUNICATIONS (ELEKTROSVYAZ')

BY VARIOUS AUTHORS

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ON RADIO DAY

Next Radio Day, our country may mark a number of substantial achievements in the field of progress in domestic radio engineering and its most diverse applications.

The further broadening of the radio-frequency ranges which can be made accessible and useful and the simultaneous striving for a more efficient utilization of the ranges already accessible remain the main trends of modern radio development. This is due to the fact that the demand for radio channels of various purposes is increasing at a sharply rising rate. A very short time ago, it seemed that the capacity of radio waves in the centimeter range was inexhaustible. However, in practice, it was found that various services must already compete for their place in the air and in the centimeter range of wavelengths.

Presently, radar, radio-relay communication links, sound and television broadcasting on ultrashort waves are predominant among the domains of radio engineering. A short survey cannot throw sufficient light upon all the achievements and all the most important trends of these domains; therefore let us consider the main ones only.

Television is developing quite intensively in our country. To date, the number of television broadcasting centers has grown to 24; moreover, a considerable number of them are in the building stage. At present, throughout the RSFSR alone, 28 television broadcasting centers and eight relay television stations are being built. Little more than one year has elapsed since the Twentieth Congress of the Communist Party of the Soviet Union (CPSU) decided to bring the number of television broadcasting centers in the Soviet Union to not less than 75 by the end of the sixth Five-Year Plan. But it is already obvious now that, as a result of the population's exceptional interest in television and of the local agencies' energetic efforts, the Twentieth Congress assignment will be overfilled. The number of television sets is growing at a correspondingly rapid rate. At the end of 1956, a total of 1,350,000 television sets were registered, whereas there were only 60,000 television sets in

1951. The industry is turning out many new types of television sets with rectangular picture tubes of large size.

The new types of television sets are continuously being improved with respect to their parameters. Thus, by the use of new tubes and better circuits, the video channel sensitivity of the television receiver has been increased to 200 μv as compared to the previous 500 and 1000 μv. A substantial improvement in the new television sets was the introduction of automatic amplification control and of automatic focusing; this simplifies the operation of the receiver and decreases the number of tuning elements. New, even better types of television sets are being developed and put into use.

In 1956, a notable step forward was made as far as the development of a color television system is concerned. After a long technical discussion, the field sequential version was rejected; the dot sequence system compatible with black-white television was approved for final development. Laboratory installations of such a system were publicly demonstrated in Moscow and Leningrad. The task of the scientific collectives which are working in this interesting field is to make color television available to the working masses at large, as soon as possible.

While noting the undoubted achievements in the field of television, the fact must be stressed at the same time that a large number of technical problems associated with the further development of television broadcasting are still being solved at quite a slow pace by the workers of the radio-engineering industry.

First of all, this concerns making accessible and useful the range of 174 - 230 mc in which seven additional television broadcasting channels are planned.

The designing of radio transmitters and television sets operating in this range must be completed and their production in adequate numbers organized as soon as possible, since new television broadcasting centers cannot be built to operate on the five existing channels.

The situation is much worse as to carrying out the Instruction of the Twentieth

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Congress of the Party concerning the development of ultrashort-wave broadcasting in the European part of the USSR. The plan for the building of ultrashort-wave broadcasting stations was disrupted in 1956 due to the fault of the radio-engineering industry. The development of new types of equipment with tetrodes has been delayed considerably. The situation as far as the creation of a receiving ultrashort-wave FM network is concerned is particularly bad. So far, a negligible number of radio receivers which can operate in the ultrashort-wave range have been produced. The ultrashort-wave broadcasting range is also absent in most of the television sets produced. The solution of this design problem cannot be considered successful even in television sets which are equipped with the above-indicated range.

Apparently, a number of workers of the industry still lack an understanding of the role and of the prospects of ultrashort-wave broadcasting. Actually, however, an analysis of the trends in development of modern broadcasting engineering quite definitely shows that, simultaneously with the broadest possible development of television in the course of the next 10 - 15 years, ultrashort-wave FM broadcasting must gradually become the basic means of sound broadcasting. Accordingly, all planning for radio broadcasting development must be based on the prospects of erecting a network of radio stations combining television broadcasting centers and ultrashort-wave FM transmitters, at first for two, and later for 4 - 5, sound channels. Therefore, the recent practice of designing and constructing television broadcasting centers without simultaneous installation of ultrashort-wave FM transmitters for sound broadcasting should be considered incorrect.

From the above considerations, it naturally follows that all new types of television sets must definitely be provided with the ultrashort-wave FM range. The fastest possible development of the most rational designs of band switches and of tuning controls is the most urgent task of the designers.

Certain achievements in the field of radio-relay communications development in the past year may be noted. Several radio-relay links are already built and additional ones are being built at the present time on the basis of the system introduced by the Scientific Research Institute (SRI) of the Ministry of Communications. Ryazan and Stalinogorsk are now receiving Moscow television programs by means of radio-relay links. The Institute completed the development of the new, completely modern radio-relay communication system R-60, and one of the large-scale plants has begun organizing the production of this system which will play an important role in carrying out the instruction of the Twentieth Party Congress. The task of the designers and of the engineers is to ensure a proper development of the work undertaken to build a number of radio-relay links based on the new system.

The collectives of the SRI of the Ministry of Communications have also done considerable work in developing a more powerful radio-relay communication system designed so that every band be multiplexed with 240 telephone channels or one television channel. Laboratory models of such a system were tested successfully in experiments and, once its production will have been organized, this system must become the basic type of equipment for main radio-relay communication systems. The honor-bound duty of the collectives of the plant and of the SRI is to complete the work undertaken in organizing the production of such an important system as soon as possible.

The vacuum-tube industry has an important role in the large-scale introduction of radio-relay communications. A radio-relay system includes a number of new types of radio tubes. The qualitative indexes of some of them are technically not satisfactory. The insufficient life of these tubes is particularly a subject of alarm; the guarantee is 500 hours, whereas corresponding radio tubes of foreign firms operate from 5000 - 10,000 hours. The rapid failure of the radio tubes may be the basic cause of communication failures and may increase the operating costs considerably.

The creation of a modern radio-relay links requires the creative efforts of many collectives. The multiplexing unit is a rather important part of the total system. The fact that the same methods of multiplexing of channels are used in radio-relay links and in cable communication lines is characteristic of modern communica-

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tions technique. The standard specifications for the communication channels are the same in both cases. Thus, the development of radio communications is inseparable from the general development of electrical communications. The past year has brought a number of substantial achievements with respect to the improvement of multiplexing systems; this will help in a larger measure the creation of modern multichannel radio-relay systems. Let us mention here the development of the new 12-channel system V12-2 and the completion of the development of the 60-channel system K-60 for multiplexing of cable lines as well as of radio-relay links. The creation of a simplified 60-channel multiplexing system of small size, KRR-30/60, suitable for short cable lines as well as radio-relay links of the type R-60 is a great success. Finally, the equipment of the experimental section of the coaxial cable with a 900-channel multiplexing system is very important. On the basis of the experiments carried out with this system, the final models of the multiplexing system for coaxial cables and for powerful radio-relay links will be built.

The entire above-described complicated pattern is a result of the creative cooperation of the workers of the radio-engineering industry and of the Ministry of Communications and is an important step toward elimination of the lag in the field of modern electrical communications engineering. However, in order to capitalize on these achievements, the production of the necessary number of the indicated models of modern communication engineering must be ensured. We insistently request the Minister of the radio engineering industry, V.D.Kalmykov, that he ensure the production of all above-mentioned types of systems and give special consideration to the question of the qualitative indexes and of the lives of radio tubes. We also invite the workers of the vacuum-tube industry to come forward on the pages of our magazine and to report on the reason for our lagging behind the world standards in this important field.

We have already noted the close correlation of modern radio communications with the general electric communications system. Such a coordination becomes more and more necessary as far as broadcasting is concerned. In order that the transmission of programs to the many broadcasting stations and to the radio rebroadcasting and receiving stations be of high quality, the network of intercity broadcasting channels must be broadened in every way, by means of cables and radio-relay links. For this purpose, a special unit is being built which combines three standard telephone channels into a wide broadcasting channel. Such channels, in particular, must by all means be made available when ultrashort-wave FM broadcasting stations are built.

The resolutions of the Twentieth Congress of the CP specify also the building of a network of main lines for exchanging television programs between the largest television centers. Radio-relay links are the basic technical means of solving this problem. In the course of the sixth Five-Year Plan, such long-distance links as Leningrad - Tallin - Riga - Vilnyus - Minsk; Moscow - Kharkov - Dnepropetrovsk - Simferopol; Moscow - Kazan - Sverdlovsk; and a number of others will be built. Many short distance radio-relay links will be built to increase the effective range of a number of television centers.

Twenty years ago, Prof. P.V.Shmakov had already proposed to extend the effective range of television centers by means of aircraft relaying. Unfortunately, no practical steps toward realizing this proposition have been taken to date. This current year, the collectives of the Leningrad and of the Odessa electric communications engineering institutes have vigorously undertaken the realization of this interesting proposition. Let us hope that their work will be completely successful.

A new trend in the development of radio communications has been to utilize the tropospheric scattering of ultrashort-waves in order to build radio-relay links in which the distance between intermediate stations would be more than 200 km. Such radio-relay links may be very efficient for communications over sparsely populated and difficultly accessible regions. Presently, experimental links which utilize tropospheric scattering are being erected. This type of work must be boosted by all means.

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One is forced to admit that the development of research in the field of ionospheric scattering of ultrashort-waves is mediocre, and it is hoped that by next year's Radio Day our scientific organizations will have achieved more substantial results.

Many data on long-distance television reception have shown how insufficiently the laws of propagation of radio waves have been investigated so far. In connection with this, the organization of the International Geophysical Year beginning in July 1957 may be welcomed. The thorough investigation of the magnetic and other properties of the earth and of the atmosphere surrounding it and their interaction with the sun and the adjoining cosmos will no doubt provide many valuable data for the further investigation of radio facilities.

Soviet specialists of all branches of radio may boast of new and remarkable achievements in the field of technical progress on Radio Day 1957. At the same time, it must be stated that the rate and scale of the development of certain important branches of radio engineering and of electric communications lag behind the general rate of development of the basic branches of the national economy of the Soviet Union. The further considerable expansion of the radio-engineering industry and of its scientific and engineering basis is a prime requisite.

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THE MOST EFFICIENT USE OF CODING SYSTEMS

by

V.I.Siforov

The properties of coding systems, which operate under conditions of interference and various statistical properties of the transmitted letters, are considered. A general expression is derived, for the probability that one letter of the telegram will be distorted. It is proved that, for given probabilities of the distortion of letters and for a given rate of transmitting messages, there exists a most efficient set of statistical properties of the letters. Quantitative relationships which characterize the various performances of the coding system are derived and they are subjected to a comparative evaluation.

Introduction

Lately, new trends in the theory of coding have emerged. They are reflected in part in the papers presented at the Second Symposium on Information Theory at Cambridge (U.S.A.) in September 1956, in which Soviet scientists took part.

The papers read by member of the academy A.N.Kolmogorov on the theory of the transmission of continuous messages, of the American scientist C.Shannon on the traffic capacity of a communication channel with noise at zero error; by D.Huffman (U.S.A.) on the linear circuit theory of error - correcting code systems; by V.I.Siforov on the theory of coding systems with small errors, and by other authors were devoted to these new trends.

In our work, an attempt was made to determine the total combination of parameters which reflect the basic properties of coding systems related to the class of binary coding systems.

The present article gives an account of the results of the investigations carried out by the author, which were devoted to the most efficient utilization of coding systems, i.e., to a utilization which would ensure obtaining the best possible characteristics of the signal transmitting system under conditions of interference, when the selected code is used.

Probability of the Distortion of the Transmitted Message

Let us assume that the letters a_1, a_2, $a_3, \ldots, a_N$ are transmitted through a system which transmits discrete messages. When the binary coding system is used, each of these letters is a code combination which consists of several elements; each combination being able to assume only two possible meanings.

When the transmission system is subject to interference, the letters being transmitted will be distorted. For instance, while the letter a_1 is fed to the input of the system, the letter a_3 may appear at the output of the system.

Let us denote the probability of distortion of the letters a_1, a_2, $a_3, \ldots$, a_N by y_1, y_2, $y_3, \ldots$, y_N, respectively. This probability, under given conditions of interference in the channel transmitting the messages, apparently is dependent upon the coding system selected.

Let p_1, p_2, $p_3, \ldots$, p_N be the probabilities of the appearance, at the input of the transmission system, of the letters a_1, a_2, a_3,.1., a_N, respectively.

In accordance with this notation, the probability that the letter a_i will appear at the input of the systems will be equal to p_i, and the conditional probability that this letter, having appeared at the input, will undergo distortion will be y_i. It follows, from the multiplication theorem of probability, that the probability of the letter a_i appearing at the input and that this letter will undergo distortion is equal to $p_i y_i$.

Since any of the letters a_1, a_2, $a_3, \ldots$, a_N may appear at the input of the system, then, in accordance with the addition theorem of probability, the probability y of the distortion of any of the letters being transmitted through the system may be represented as a sum of the products $p_i y_i$ at all possible values of i from 1 to N, i.e.,

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$$y = \sum_{i=1}^{N} p_i y_i. \tag{1}$$

Let the message transmitted or "telegram" consist of M letters. The probability of an undistorted transmission of each of these letters is equal to 1-y, and the probability of an undistorted transmission of the entire telegram, in accordance with the multiplication theorem of probability, will be equal to $(1-y)^M$.

The probability y_T that the telegram being transmitted will be subject to any kind of distortion will be

$$y_T = 1 - (1 - y)^M \tag{2}$$

or

$$y_T = My - \frac{M(M-1)}{2} y^2 + \frac{M(M-1)(M-2)}{6} y^3 - \dots$$

If

$$My \ll 1,$$

then

$$y_T \approx My. \tag{3}$$

Equation (3) shows that when the probability of distortion of each letter is sufficiently small, the probability of distortion of the telegram in its entirety is approximately proportional to the number of letters it contains.

From eq.(3) it also follows that the quantity y, determined by the relationship (1), is the probability of distortion of the telegram with respect to one letter.

<u>The Most Efficient Statistical Properties of the Message Source</u>

The statistical properties of the message sources are characterized first of all by the total combination of the probabilities $p_1, p_2, \dots, p_N$ of appearance of the letters $a_1, a_2, \dots, a_N$. It is known from the information theory (Bibl.2) that the number of messages referred to one letter which enter the system (entropy)

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will be

$$H = -\sum_{i=1}^{N} p_i \log_2 p_i. \qquad (4)$$

When the probabilities $y_1, y_2, \ldots, y_N$ of distortion of the transmitted letters are small, the number of messages transmitted through the system with respect to one letter will be approximately expressed by eq.(4).

The probability y that the transmitted telegram will be distorted, with respect to one letter depends on the probabilities $y_1, y_2, \ldots, y_N, p_1, p_2, \ldots, p_N$, in accordance with eq.(1), and consequently also upon the entropy H.

For given values of the probabilities $y_1, y_2, \ldots, y_N$, corresponding to the selected coding system, and for a given value of the entropy H, the probability y will evidently depend on the magnitudes of $p_1, p_2, \ldots, p_N$, i.e., on the statistical properties of the message source. For a definite total combination of $p_1, p_2, \ldots, p_N$, the specific probability y that the telegram transmitted will be distorted is minimized. In other words, for every coding system and for a given rate of message transmission, there exists a most efficient set of statistical properties of the message source, for which the probability with respect to one letter that the telegram will be distorted is minimized.

In order to discover this most efficient set of statistical properties of the message source and the minimum specific probability that the telegram will be distorted with respect to this set of properties, let us turn to the system of equations

$$\left.\begin{aligned} y &= \sum_{i=1}^{N} p_i y_i \\ \sum_{i=1}^{N} p_i \ln p_i &= -H \ln 2 \\ \sum_{i=1}^{N} p_i &= 1 \end{aligned}\right\} \qquad (5)$$

Here, the first expression coincides with eq.(1), the second follows from eq.(4), and the third expresses the fact that any one of the N possible letters always appears at the input of the considered coding system.

In the system of equations (5) y_i = const, H = const, p_i = var, and y = var.

In accordance with the method of determining relative maxima and minima, known from mathematics (Bibl.3), let us consider the function

$$\Phi=\sum_{i=1}^{N}p_i y_i+\lambda_1\left[\left(\sum_{i=1}^{N}p_i\ln p_i\right)+H\ln 2\right]+\lambda_2\left[\left(\sum_{i=1}^{N}p_i\right)-1\right],$$

where λ_1 and λ_2 are certain unknown quantities.

If the partial derivatives of the function Φ with respect to all the variables p_i, are equated to zero, while the quantities p_i are considered independent and λ_1 and λ_2 constant, we will obtain

$$y_i+\lambda_1(\ln p_i+1)+\lambda_2=0\ (i=1,\ 2,\dots\ N). \qquad (6)$$

This relationship together with the second and third expressions in the system (5), will form a system of N + 2 equations in N + 2 unknowns p_1, p_2,..., p_N, λ_1 and λ_2. The quantities p_1, p_2,..., p_N obtained as a result of solving this system correspond to the most efficient set of statistical properties of the message source, which will ensure a minimum specific probability of the telegram being distorted.

If each of eqs.(6) is solved for p_i, we obtain

$$p_i=e^{-\left(1+\frac{\lambda_2}{\lambda_1}\right)}e^{-\frac{y_i}{\lambda_1}}.$$

If the quantities p_i at all values of i from 1 to N are summed up, and the third expression of the system (5) is taken into consideration, we obtain

$$e^{-\left(1+\frac{\lambda_2}{\lambda_1}\right)}\sum_{i=1}^{N}e^{-\frac{y_i}{\lambda_1}}=1.$$

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If the factor $e^{-(1+\frac{\lambda_2}{\lambda_1})}$ is eliminated from this and the preceding equation and if we assume that

$$z=\frac{1}{\lambda_1}, \tag{7}$$

then

$$p_i=\frac{e^{-y_i z}}{\sum_{i=1}^{N} e^{-y_i z}}. \tag{8}$$

It can be seen from eq.(8) that a definite total combination of the probabilities $p_1, p_2, \ldots, p_N$ and, consequently, as it follows from eqs.(5), a definite entropy

$$H=-\frac{1}{\ln 2}\sum_{i=1}^{N} p_i \ln p_i \tag{9}$$

and a definite minimum probability of the telegram being distorted with respect to one letter

$$y_{min}=\sum_{i=1}^{N} p_i y_i \tag{10}$$

correspond to each and every value of the parameter z.

By means of eqs.(8), (9), and (10), for given values of the probabilities $y_1, y_2, \ldots, y_N$, the functions $H(z)$ and $y_{min}(z)$ can be found and, consequently, the relationship $y_{min}(H)$ can be determined; we will call this relationship the characteristic of optimum utilization of the coding system. This characteristic expresses the dependence of the lowest possible specific probability of telegram distortion upon the number of messages relative to one letter which the telegram contains.

Figure 1 shows the characteristic of optimum utilization of the coding system for the case where $y_1 = 10^{-3}$, $y_2 = 4.10^{-4}$ and $y_3 = 10^{-4}$. The specific probability of distortion y_{min}, during a decrease in entropy H from its maximum value H_{max} = 1585 to zero, falls smoothly from 5×10^{-4} to 1×10^{-4}.

Assuming that z = 0 in eqs.(8), (9) and (10), we obtain

$$\left.\begin{array}{l} p_i = \frac{1}{N} \\ H = \log_2 N \\ y_{min} = \frac{1}{N}\sum_{i=1}^{N} y_i \end{array}\right\}, \qquad (11)$$

and, as $z \to \infty$, we obtain

$$\left.\begin{array}{l} p_i \to 1 \text{ for } i = k \\ p_i \to 0 \text{ for } i \neq k \\ H \to 0 \\ y_{min} \to y_k \end{array}\right\}, \qquad (12)$$

where y_k is the lowest probability of all the differing probabilities $y_1, y_2, \ldots, y_N$.

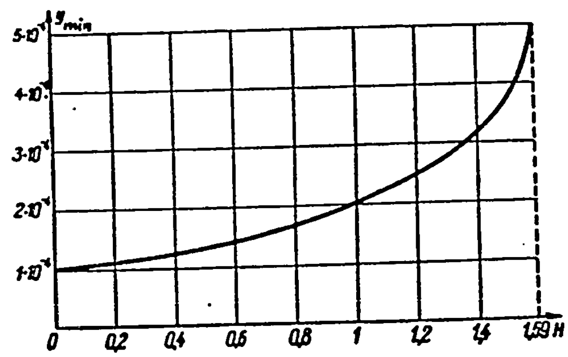

Fig.1

The expressions (11) and (12) characterize the extreme operating conditions of utilization of the coding system. From these expressions, it follows that, as the entropy H decreases from its highest value $H_{max} = \log_2 N$ to zero, its minimum probability of telegram distortion decreases from the arithmetic mean value of the probability y_i of all the letters being distorted to the minimum values of all these probabilities.

Transmission of a Given Number of Messages with Minimum Probability of Distortion

One of the basic parameters of the coding system is the relative probability of a decoding error (α) by which we mean the ratio of the probability of telegram distortion with respect to one letter (y), to the number of messages contained in this telegram, also with respect to one letter (H), i.e.,

$$\alpha = \frac{y}{H}. \qquad (13)$$

The relative probability of a decoding error α is thus the probability of error during transmission of one message unit. Since the quantity y has the dimension $\frac{1}{\text{letter}}$, and the entropy H the dimension $\frac{\text{binary unit}}{\text{letter}}$, the quantity α has the dimension $\frac{1}{\text{binary unit}}$, i.e., is expressed in reciprocal binary units.

If the telegram transmitted contains M letters, then the probability of its distortion, in accordance with eq.(3), will be $y_T = My$, and the number of messages which it contains D will be

$$D = HM. \tag{14}$$

If this is taken into consideration, eq.(13) may be written as

$$y_T = D\alpha. \tag{15}$$

The lower the relative probability α, the lower will be the probability of telegram distortion y_T for a given number D of messages which it contains. If the probability y_T is considered to be given, which corresponds to securing a definite degree of reliability of an error-free telegram transmission, then the decrease in the parameter α permits an increase in the quantity D. In other words, the operating conditions of the coding system, which corresponds to low values of the parameter α, secures a large number of messages contained in the telegram (D) for a given degree of reliability of an error-free transmission.

For a given probability $y_1, y_2, \dots, y_N$ of the transmitted letters being distorted, the relative probability α will be a function of the quantities $p_1, p_2, \dots, p_N$ and H. For H = const, in accordance with eq.(13), the parameter α will have its lowest value at $y = y_{min}$, i.e.,

$$\alpha_{min} = \frac{y_{min}}{H}, \tag{16}$$

where y_{min} is expressed by eq.(10) and is a function of the entropy H.

As the entropy H decreases from its maximum value H to zero, the specific probability y_{min}, as shown in the previous Section, also decreases from the arithmetic mean value of the probabilities $y_1, y_2, \ldots, y_N$ to the lowest of these probabilities. Then, the parameter α_{min} decreases at first, reaches its minimum possible value $\alpha_{min\ min}$, then increases again, and, as $H \to 0$, finally tends toward infinity.

The operating conditions of a coding system for which the relative probability

$$\alpha = \alpha_{min\ min} \tag{17}$$

is a condition which ensures transmission of the largest possible number of messages contained in the telegram, for a given probability of the latter being distorted. This operating condition also gives the lowest possible probability of the telegram being distorted, for a given number of messages contained in the telegram.

The quantities y_{min} and H entering the formula for α_{min}(16) are functions of the parameter z. The value of this parameter which corresponds to the optimum performance at which the relative probability α attains its minimum possible value $\alpha_{min\ min}$, is determined from the equation

$$\frac{d\alpha_{min}}{dz} = 0.$$

If eq.(16) is taken into consideration, this equation may be expressed in the form

$$\frac{dy_{min}}{dz} H - \frac{dH}{dz} y_{min} = 0. \tag{18}$$

If eqs.(9) and (10) are differentiated with respect to z, and if eq.(8) is taken into consideration while assuming

$$\left.\begin{aligned} A &= \sum_{i=1}^{N} e^{-y_i z} \\ B &= \sum_{i=1}^{N} y_i e^{-y_i z} \\ C &= \sum_{i=1}^{N} y_i^2 e^{-y_i z} \end{aligned}\right\}, \tag{19}$$

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we obtain

$$\left.\begin{aligned} y_{\min} &= \frac{B}{A} \\ H &= \frac{1}{\ln 2}\left(\frac{Bz}{A} - \ln A\right) \\ \frac{dy_{\min}}{dz} &= \frac{B^2 - AC}{A^2} \\ \frac{dH}{dz} &= \frac{z}{\ln 2}\left(\frac{B^2 - AC}{A^2}\right) \end{aligned}\right\}. \qquad (20)$$

If these expressions are substituted in eq.(18), we will have

$$A(B^2 - AC)\ln A = 0$$

or, since in the case considered

$$\left.\begin{aligned} A &\neq 0 \\ B^2 - AC &\neq 0 \end{aligned}\right\},$$

i.e.

$$\ln A = 0$$

or

$$A = \sum_{i=1}^{N} e^{-y_i z} = 1. \qquad (21)$$

If the root of this transcendental equation is denoted by z_0 and if eqs.(8), (9), (10), (16), (19), and (20) are taken into consideration, we will have

$$\left.\begin{aligned} p_i &= e^{-y_i z_0} \\ y_{\min} &= \sum_{i=1}^{N} y_i e^{-y_i z_0} \\ H &= \frac{y_{\min} z_0}{\ln 2} \\ \alpha_{\min\min} &= \frac{\ln 2}{z_0} \end{aligned}\right\}. \qquad (22)$$

By means of eqs.(21) and (22), all the basic quantities which characterize the the optimum utilization of the coding system can be determined; this optimum performance secures the largest possible number of messages which the telegram may contain, for a given probability of it being distorted.

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Referring to the example given in the preceding Section, let us assume $y_1 = 10^{-3}$, $y_2 = 4\times10^{-4}$, and $y_3 = 10^{-4}$. Applying eqs.(21) and (22), we will obtain for the optimum performance: $z_0 = 3500$; $p_1 = 0.03$; $p_2 = 0.25$; $p_3 = 0.72$; $y_{min} = 2.03\times10^{-4}$; $H = 1$; $\alpha_{min\,min} = 2.03\times10^{-4}$.

Comparison of the Various Operating Conditions of a Coding System

Figure 2 shows the dependence of the relative probability $\alpha_{min} = \frac{y_{min}}{H}$ upon the entropy H for various numerical values of the probabilities y_1, y_2, and y_3 which correspond to the example given above.

Point A on the curve in Fig.2 corresponds to the highest possible rate of message transmission or to the highest entropy H (operating condition A). Here, all the probabilities $p_1, p_2, \ldots, p_N$ are equal, i.e., the letters are transmitted with equal frequency. Using this operating condition, secures the transmission of the largest possible number of messages through the coding system in a given period of time.

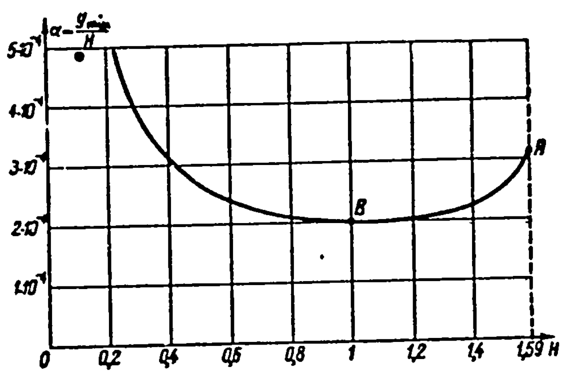

Fig.2

Point B on the curve in Fig.2 corresponds to the performance which provides for the largest possible number of messages contained in each telegram for a given probability of it being distorted (operating condition B). Using this operating condition, secures the lowest possible probability of each telegram being distorted for a given number of messages that it contains. Thus, class B operation is the operation which provides for the highest possible quality of transmission of each telegram.

Table of the Basic Parameters of a Coding System

Performance	H	y_{min}	$\alpha = \frac{y_{min}}{H}$
A	1,58	$5\cdot10^{-4}$	$3,16\cdot10^{-4}$
B	1	$2,03\cdot10^{-4}$	$2,03\cdot10^{-4}$

The Table gives the basic parameters of a coding system for class A and B operation, applicable to the numerical values of the probabilities $y_1 = 10^{-3}$, $y_2 = 4\times10^{-4}$ and $y_3 = 10^{-4}$.

This Table indicates that, as the operation changes from class A to B, the number of messages transmitted during the period of time allowed for all the telegrams to be transmitted, drops 1.58 times. On the other hand, changing from class A to class B operation permits increasing the number of messages that it contains by $\frac{3.16\times10^{-4}}{2.03\times10^{-4}} = 1.56$ times, for a given probability that every telegram will be distorted.

It also follows from the given example that, even when the difference between the probabilities that the various letters will be distorted, is relatively large, selecting the most efficient set of statistical properties results only in negligible improvement of the operation of the coding system. Actually, for a ratio of extreme probability values $\frac{y_1}{y_2} = \frac{10^{-3}}{10^{-4}} = 10$, the relative probability α of an error in decoding can be decreased 1.56 times by selecting the most efficient set of statistical properties of the letters being transmitted.

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THEORETICALLY OPTIMUM COMMUNICATION SYSTEM

by

A.A.Kharkevich

From the theoretical point of view, selecting a communication system consists in selecting a method of communication (i.e., a code) and a method of reception. An optimum system is one that provides maximum efficiency for a given minimum distortion or, conversely, one that has the lowest minimum distortion for a given efficiency. It is shown that selecting an optimum system consists in a certain variation problem, whose formulation includes the distribution of distortion probabilities. Several examples are given to illustrate the formulation of the problem.

Many recent papers devoted to general communication theory discuss the problems associated with the building of optimum coding systems and with ideal reception. In this respect, it should be noted that these problems, generally speaking, cannot be considered separately. The selection of a code (i.e., a method of communication) and the selection of a method of reception are two aspects of a single problem, which consists in building an optimum communication system. This problem has a definite answer if the operating conditions of the system are given, in particular, if the characteristic of the distortion acting on the system is given.

Let us agree on a definition for an optimum system: let us call optimum a system which possesses the highest efficiency for a given minimum distortion or, conversely, possesses the lowest minimum distortion for a given efficiency.

Let us give a quantitative definition for both of the above-mentioned properties; let us express the minimum distortion by the probability q of a correct reception; let us characterize the efficiency by the amount of information for the coding combination, i.e., by the quantity log N, where N is the total number of code combin-

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ations (assumed to be equally probable).

Now, let us attempt to formulate the problem, in geometric terminology, which is already sufficiently familiar to us. First, let us consider the number of code combinations. If the usual condition that the energies of all code combinations are the same is imposed,

$$E_i = \sum_{k=1}^{n} x_{ik}^2 = E = \text{const},$$

where i is the number of combinations, n the number of signs (symbols, coordinates), the problem reduces to distributing either a given or maximum number of code points over the surface of an n-dimensional sphere of radius $\sqrt{E}$.

With respect to the minimum distortion, this depends also on the method of reception. The problem may be formulated as follows: As a result of imposing a distortion, the point of the received signal is displaced with respect to the point of the transmitted signal. The operation of the receiver consists in identifying the received signal with the i^{th} transmitted signal when the point of the received signal is located inside a certain n-dimensional region Q_i, which is the region of correct reception for the i^{th} signal. For such a formulation of the problem, the probability q_i of a correct reception of the i^{th} signal is the probability that the point of the received signal, during transmission of the i^{th} signal will fall into the region Q_i. The configuration and arrangement of the regions Q_i determine the method of operation of the receiver. A receiver which ensures the lowest possible minimum distortion may be called ideal; however, it must be emphasized that this property is not absolute; as many ideal receivers as the number of various possible operating conditions could be listed.

The probability that the point of the received signal will fall into the region Q_i depends upon the multi-dimensional probability density of distortion, represented by some function $\varphi(x_1, x_2, \ldots, x_n)$. Now, we can formulate the problem of determining the optimum system for the version in which N is given as follows: maximize the quantities

$$q_i = \int_{Q_i} \varphi(x_1 - x_{i1},\ x_2 - x_{i2},\ \ldots,\ x_n - x_{in})\, dx_1,\ dx_2, \ldots, dx_n \qquad (2)$$

provided, in addition, that eq.(1) is valid. The equality of the probability of a correct reception of all signals may be taken as a second additional provision

$$q_1 = q_2 = \ldots = q_N = q. \qquad (3)$$

Here, q_i must be maximized by varying the limits of the region Q_i (selecting a method of reception) and by varying the coordinate x_{ik} of the code points (selecting a code). Consequently, the problem is a combination of a variation problem with a moving limit and of the problem of determining the conditional extreme of a function in n variables. Obviously, this is not at all a simple problem. But we are not attempting to find methods for solving the problem in its general form. The general formulation of the problem is given here only in order to show that it is a single problem. Expression (2) indicates directly that, for a given code (i.e., when x_{ik} are given), a receiver of maximum efficiency (i.e., a most favorable configuration of the region Q_i) may be found and vice versa. The distribution of distortion (i.e., the function φ) is assumed to be given. However, a case is conceivable in which Q_i and x_{ik} are given, and in which the distortion characteristic which will ensure the lowest (or, possibly, the highest) minimum distortion under the given conditions is to be determined. In short, the code, the method of reception, and the distortion characteristic are mutually associated and their interrelationship is expressed by eq.(2).

We will attempt to clarify the significance of the above-mentioned general relationships by means of a few very simple examples, whose consideration presents no mathematical difficulties.

Example 1. Let there be two signals only (N = 2), spaced at a distance d in n-dimensional space. With respect to the distortion density distribution, let us assume that it is spherosymmetric, i.e., that it depends only on

$$r=\sqrt{x_1^2+x_2^2+\ldots+x_n^2}.$$

This is a very logical assumption which signifies that all the possible directions of the distortion vector are equally probable. Let φ be the decreasing function r (in particular, this is a property of the normal distribution). For the probability of correct reception, we have

$$q_1=\int_{Q_1}\varphi(x_1-x_{11},\ x_2,\ \ldots\ x_n)\,dx_1,\ dx_2,\ \ldots dx_n,$$
$$q_2=\int_{Q_2}\varphi(x_1-x_{21},\ x_2,\ \ldots\ x_n)\,dx_1,\ dx_2,\ \ldots\ dx_n$$

(we assume that the points of both signals lie on the axis X_1). Assuming

$$q_1=q_2=q,$$

let us take

$$x_{11}=\frac{d}{2},\quad x_{21}=-\frac{d}{2}.$$

Due to the symmetry expressed by the equality q_1, the regions Q_1 and Q_2 have mirror symmetry with respect to the hyperplane AA', which is normal to the segment d and divides it in half, as shown in Fig.1, where the region Q_2 is crosshatched. An

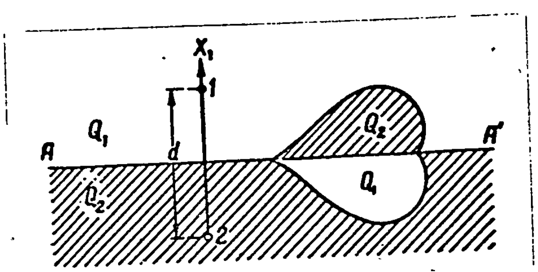

Fig.1

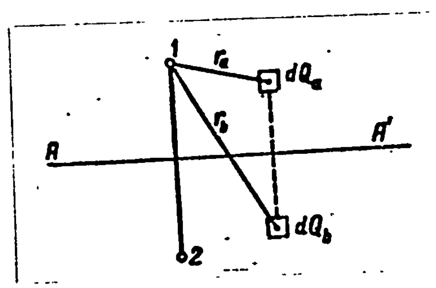

Fig.2

elementary discussion will show that, in order to maximize $q_1 = q_2$, the plane of symmetry AA' must be taken as the boundary of the regions Q_1 and Q_2. Let us consider the pair of volume elements dQ_a and dQ_b, placed symmetrically with respect to AA', as shown in Fig.2. Let the signal 1 be transmitted and let dQ_b pertain to Q_1. However, since $r_b > r_a$, the probability of a correct reception q_1 would increase if dQ_b were replaced by dQ_a. Thus dQ_b should be associated with Q_2, and dQ_a with Q_1. If such a discussion is applied to any pair of symmetrical volume elements, we come to the conclusion that if the distribution density of the distortion decreases with decreasing distance (in accordance with what law is immaterial), the ideal receiver will be one which associates the signal received with the nearest possible signal. This is the ideal receiver according to Kotelnikov.

The optimum code remains to be found. However, in the given simple case, it is hardly necessary to prove that the minimum distortion decreases with increasing d and that, consequently, assuming that condition (1) is valid, the points of the signals must be placed at the ends of the diameter of the sphere of the signals. The distance will then be $d = 2\sqrt{E}$.

Example 2. Let, as before, the distribution density of distortion decrease with increasing distance, but let us now take $N = 8, n = 3$. The points of the signals are placed on a three-dimensional sphere. If a binary code is selected, the code points will place themselves at the vertices of a cube. The regions of correct reception will be octants, i.e., the interiors of trihedral angles, formed by the planes normal to the edges of the cubes and bisecting these edges. However, the binary code is actually the most advantageous code; its geometric presentation is obtained if one of the faces of the cube is rotated through 45^o in its plane. The figure so obtained is an irregular decahedron (8 triangles, 2 squares) with 16 equal edges which are longer than the edge of the cube inscribed in the same sphere. The region of correct reception is located inside the tetrahedral angle, as shown in Fig.3. Let us note that the polyhedron considered, though irregular, is symmetrical in the sense that if any vertex is placed at a given point, the figure may coincide with itself by rotating it about the center. Therefore, all the regions Q_i are equal and, consequently, the probabilities q_i are also equal.

<u>Example 3.</u> Let us again assume $N = 2$ and let, for the sake of simplicity, $n = 2$ (which allows representation of the entire picture in one plane), but now let the

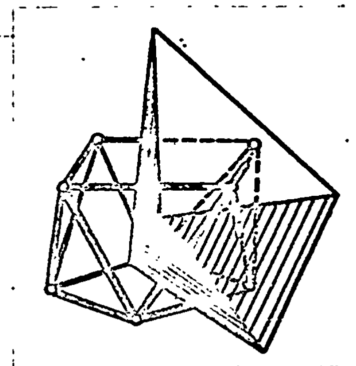

Fig.3

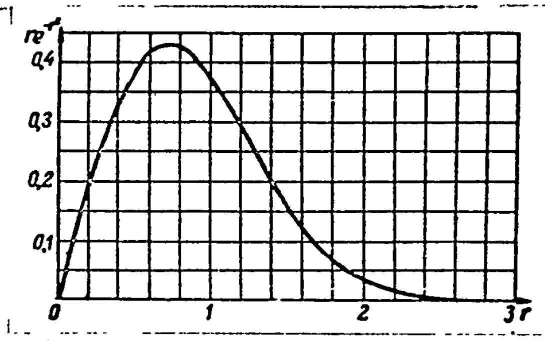

Fig.4

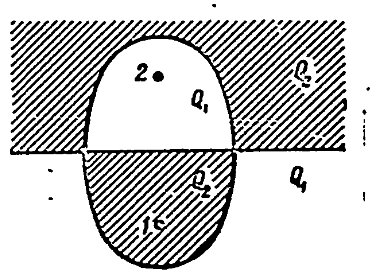

Fig.5

probability density of distortion decrease at a rapid rate. Let, for instance,

$$\varphi(r) = Cr\, e^{-r^2}$$

(Fig.4). Let us discuss the location of the boundary of the correct reception region along the same lines as in example 1, let us formulate the conclusion slightly differently, namely, each element of volume dQ should be associated with the region of correct reception of that signal, which upon reception has the highest probability of falling into dQ. In other words, the boundary of the region of correct reception must be the locus of the points where the probability density for two signals becomes equal. In the case considered, this gives

$$\left[\left(x_1 - \frac{d}{2}\right)^2 + x_2^2\right]^{1/2} \exp - \left[\left(x_1 - \frac{d}{2}\right)^2 + x_2^2\right] = \\ = \left[\left(x_1 + \frac{d}{2}\right)^2 + x_2^2\right]^{1/2} \exp - \left[\left(x_1 + \frac{d}{2}\right)^2 + x_2^2\right],$$

or

$$\frac{x_1 d}{\operatorname{th} 2x_1 d} - r^2 + \frac{d^2}{4} = 0,$$

where

$$r^2 = x_1^2 + x_2^2.$$

The boundary of the region of correct reception is shown in Fig.5. The contour of the boundary obviously depends upon d.

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UHF OSCILLATOR MODULATOR

by

E.P.Korchagina

A two-tube network of an UHF oscillator, permitting frequency control, is described. In this network, both tubes deliver power to the load and both participate in frequency control. The frequency deviations produced by the tubes will be cumulative if the feedback factors of both tubes are complex conjugate numbers. The modulating voltages are supplied to the tubes 180° out of phase. This circuit allows larger frequency deviations than the reactance-tube circuit and gives a better stability of the carrier frequency, at fluctuations in the supplied voltages.

Introduction

Two-tube networks used for frequency control are discussed below. Single-tube networks, in which control is effected by means of varying the grid currents, are not considered.

Frequency control in two-tube networks may be achieved by means of two types of duty. The type of duty, in which one of the tubes operates as a generator, while the second tube (the reactance tube) is used for frequency control, is widely known. The second type of duty, in which both tubes are used for frequency control and in which both deliver power to the circuit, is described in Mansfeld's paper (Bibl.1). We will refer to this type of duty as the oscillator-modulator duty. Let us discuss the problem of using both types of duty UHF oscillators.

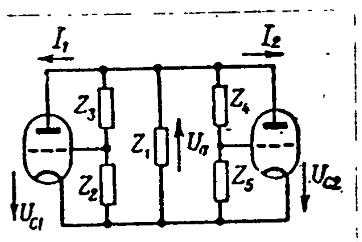

Fig.1

Basic Relationships

Let us consider the conditions of the steady-state operation of the two-tube oscillator shown in Fig.1. The notations and the positive directions of the currents and voltages are shown in Fig.1.

If the inertia of the electrons and the plate reaction are neglected, the following expressions for the amplitudes of the first harmonics of the tube plate currents may be written:

$$\overline{I}_1 = S_1 \overline{U}_{c1} = S_1 \overline{K}_1 \overline{U}_a,$$
$$\overline{I}_2 = S_2 \overline{U}_{c2} = S_2 \overline{K}_2 \overline{U}_a,$$

where $\overline{K}_1 = \frac{\overline{U}_{c1}}{U_a}$, $\overline{K}_2 = \frac{\overline{U}_{c2}}{U_a}$ are the feedback factors of the first and of the second tube respectively;

$S_1 = S\gamma_1(\theta_1)$ and $S_2 = S\gamma_1(\theta_2)$ are average transconductances;

S is the steepness of the statistical characteristic of the plate current;

$\gamma_1(\theta) \quad \frac{1}{2\pi}(2\theta - \sin 2\theta)$ is the scanning factor for the plate current first harmonic;

θ is the cutoff angle of the plate current.

If we let $\overline{Z}$ denote the total impedance in the plate circuits of the tubes, then the plate voltage will be

$$\overline{U}_a = (\overline{I}_1 + \overline{I}_2)\overline{Z}.$$

If the values of the currents are substituted, the following condition for the existence of steady-state operation is obtained:

$$1 = \overline{Z}(S_1 \overline{K}_1 + S_2 \overline{K}_2). \qquad (1)$$

If we suitably substitute the real and imaginary terms in the right-hand and left-hand sides of eq.(1), taking into consideration that

$$\overline{K}_1 = K e^{i\varphi_{K1}}; \quad \overline{K}_2 = K e^{i\varphi_{K2}}; \quad \overline{Z} = \frac{R_{oe}}{1 + ia},$$

where

$$\alpha = -\operatorname{tg}\varphi_a \approx \frac{2\Delta\omega}{\omega_0\delta},$$

ω_0 is the resonant frequency of the plate circuit formed by the impedances Z_1, Z_2, Z_3, Z_4, Z_5;

δ is the plate circuit damping;

φ_a is the phase shift between the total current of the two tubes and the voltage in the circuit;

R_{oe} is the resonant resistance of the plate circuit;

φ_{k1} and φ_{k2} are the phase angles of the feedback factors.

Then we will obtain the amplitude balance and phase balance equations in the following form:

$$R_{oe}(K_1S_1\cos\varphi_{k1} + K_2S_2\cos\varphi_{k2}) = 1, \quad (2)$$

$$\operatorname{tg}\varphi_a = -R_{oe}(K_1S_1\sin\varphi_{k1} + K_2S_2\sin\varphi_{k2}). \quad (3)$$

These equations, strictly speaking, are valid only in the case when the impedances in the plate circuits of the tubes form a single-circuit modulating system. In the UHF range, two-circuit oscillators with a common grid are used. In this case, the expression $\overline{Z} = \frac{R_{oe}}{1 + i\alpha}$ is approximate and is valid only when the cathode-grid circuits are considerably detuned.

Usually, these conditions are satisfied so that eqs.(2) and (3) may be considered adequate for evaluating the operation of UHF oscillators. The frequency ω_0 in such a case is understood to be one of the communication frequencies which undergoes self-oscillation.

Let us now consider the behavior of the network operating as an oscillator with a reactance tube.

Single-circuit oscillators are used in the ranges of long and short waves. In this case, the feedback factor phase angle of the oscillator tube is $\varphi_{k1} = 0$, and the phase angle of the feedback factor of the reactance tube can be given a value of

$\varphi_{k2} \approx 90^\circ$. Under these conditions, the equations for phase and amplitude balance will have the form

$$S_1 K_1 R_{oe} = 1, \qquad (4)$$

$$\mathrm{tg}\,\varphi_a = -R_{oe} K_2 S_2 \sin \varphi_{k2}. \qquad (5)$$

From the expressions obtained, it follows that the change in the transconductance of the reactance tube controls the frequency and does not affect the oscillation amplitude.

The performance of an oscillator with a reactance tube is less efficient in the UHF range than on long waves, for two reasons. First, due to the effect of the interelectrode capacitances, the phase angle of the feedback factor of the reactance tube differs from 90°. This causes a decrease in the frequency deviations due to the change in transconductance of the reactance tube. Secondly, UHF oscillators are usually designed as two-circuit networks so that the phase angle of the feedback factor of the oscillator tube is not equal to zero. In this case, as follows from the amplitude-balance equation (2), a change in transconductance of the reactance tube will be accompanied by a change in the equivalent transconductance of the oscillating tube, due to a change in the oscillation amplitude. A change in the transconductance of the oscillating tube, in turn, will cause a change in the frequency of the self-oscillations, as follows from the phase-balance equation (3). The magnitude of the resulting decrease in frequency will be essentially dependent on the sign of the feedback factor phase angles of the oscillating tube and of the reactance tube. If the respective phase angles of the self-excited oscillator tube and of the reactance tube are of the same sign, the change in frequency caused by the oscillator tube will decrease the resultant frequency decrement. If $K_1 = K_2$ and $\varphi_{k1} = \varphi_{k2}$, then the resulting frequency will be equal to zero. In the networks that are generally used for UHF, the frequency change caused by the oscillator tube decreases the resultant frequency decrement.

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It follows from the above statements that the duty of an oscillator with a reactance tube is no less useful in the UHF range than in that of long waves only because of poor phase angle ratios of the reactance tube; but also because the feedback factor phase angle of the oscillating tube impairs the operation of the network. Physically, this can be explained by the fact that when the feedback factor of the oscillating tube is a complex number, this tube in addition to having a negative impedance which perpetuates self-oscillations, also introduces into the circuit a reactance whose magnitude varies with changes in the oscillation amplitude.

By virtue of the above reasons, the use of reactance tubes does not give good results in the UHF range.

Let us now switch to consideration of the behavior of the network as an oscillator-modulator. For a summation of the frequency deviations produced by both tubes, the feedback factors must be complex conjugate numbers:

$$|K_1| = |K_2| = |K|; \quad \varphi_{\kappa 1} = -\varphi_{\kappa 2} = \varphi_{\kappa}.$$

The steady-state performance eqs.(2) and (3) in such a case take the form

$$1 = R_{oe} K (S_1 + S_2) \cos \varphi_{\kappa}, \qquad (6)$$

$$\operatorname{tg} \varphi_a = -R_{oe} K (S_1 - S_2) \sin \varphi_{\kappa}. \qquad (7)$$

Equations (6) and (7) show that, in the process of frequency control, an increase in the transconductance of one tube must be accompanied by a decrease in the transconductance of the other tube. Therefore, the modulating voltage must be supplied to the tubes 180° out of phase.

The mean transconductance may be changed by varying the plate voltage or the control grid voltage. Grid modulation is of great practical interest since it requires lower voltages for frequency control.

The relationship between the mean transconductance and the shift is determined, as is generally known, by the function $\gamma_1(\cos\theta)$ which has a nonlinear section at

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small cutoff angle. If it is assumed that the relationship between the mean transconductance and the shift is approximately linear and that both tubes have approximately the same mean transconductance S, in the absence of a modulating voltage, then the application of a modulating voltage will cause the same change in transconductance ΔS for both tubes. Under these circumstances,

$$S_1 = S + \Delta S; \quad S_2 = S - \Delta S.$$

If the values of S_1 and of S_2 are substituted into eqs.(6) and (7), we will have

$$1 = 2R_{oe}KS\cos\varphi_k, \qquad (8)$$

$$\mathrm{tg}\,\varphi_a = -KR_{oe}2\Delta S\sin\varphi_k = -\frac{\Delta S}{S}\,\mathrm{tg}\,\varphi_k. \qquad (9)$$

Equation (9) shows that, if the above assumptions are valid, the frequency deviation will be proportional to the transconductance increment ΔS. Then, the oscillation amplitude which is determined by eq.(8) is independent of the modulating voltage and is determined by the steepness at the initial performance level. Naturally, sustain the oscillations, the feedback phase angles of both tubes must be less than 90°. The smaller the phase angle φ_k, the larger will be the radio-frequency power and the lower will be the frequency variation which is obtained in the oscillator-modulator network.

It must be emphasized that the oscillator-modulator duty has much in common with the flip-flop frequency modulator. When the oscillator-modulator network is symmetrical, a change in the voltages supplied has no effect on the frequency since such a change will be cophasal for both tubes.

If the nonlinearity of the function $\gamma_1(\cos\theta)$ is taken into account, the oscillation amplitude changes somewhat in the process of frequency control.

The relationship between the frequency and the modulating voltage becomes nonlinear and the maximum frequency deviation decreases as well.

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A comparative evaluation of the r-f power and of the frequency deviations which are obtainable by means of an oscillator with a reactance tube and of those obtainable by means of an oscillator-modulator with a single-circuit oscillator network. The oscillator-modulator was designed with the nonlinearity of the function $\gamma_1(\cos\theta)$ taken into consideration. The irregularity of the power in the frequency control process depends on the choice of the cutoff angle at the initial operating conditions. A cutoff angle of $\theta = 80^\circ$ was selected at which the irregularity of the power does not exceed ± 10%. The phase angle of the reactance tube feedback factor was assumed to be equal to 90°; therefore the power of an oscillator with a reactance tube remains unchanged in the process of frequency control. It was assumed that one tube was used as much as the other.

The calculations showed that, in order to obtain the same amount of power from the oscillator-modulator as from the oscillator with a reactance tube, it will be practical to give the oscillator-modulator a feedback factor phase angle of the order of 40°. The maximum frequency deviation produced by the oscillator-modulator is then 35% larger than the frequency deviation produced by the reactance tube.

It follows from the above comparison that the oscillator-modulator duty has certain advantages in the long-wave and short-wave ranges for which single-circuit self-excited generators are used.

As shown above, no satisfactory results can be obtained in the UHF range by using a reactance tube; therefore, an investigation of oscillator-modulator duty as applied to UHF generators is of practical interest. It should be mentioned that the obtained expressions which characterize the behavior of an oscillator-modulator with a multicircuit self-excited generator network are valid as first approximations only.

The phase angle of the feedback factor in a multicircuit generator is determined by a detuning of the circuits at the frequency of self-oscillations; therefore, in the frequency control process, the phase angle will not remain constant as it did in a single-circuit generator network. A special investigation is necessary in

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order to evaluate the operation of an oscillator-modulator in the UHF range. However, the above relationships illustrate the operating principle of the network and make it possible to formulate the basic requirements to be met by the UHF oscillator-modulator.

Equations (6) and (7) for an oscillator-modulator were obtained for parallel connection of the plate circuits of both tubes; evidently, the equations remain valid for a flip-flop plate circuit connection. Since parallel connection of the tubes is undesirable for the UHF range, the UHF oscillator-modulator must be designed on the basis of the flip-flop circuit. The feedback factor phase angle of UHF generators is determined by the sign of the detuning of the cathode-grid circuit; therefore, in order to obtain complex conjugate feedback factors, one of the tubes must be operating in accordance with the equivalent circuit of a capacitive Hartley oscillator and the other in accordance with an equivalent circuit of an inductive Hartley oscillator. If the design of the existing tubes is taken into account, the oscillator-modulator must be built in accordance with the common-grid circuit. The modulating voltages must be supplied to the tubes 180° out of phase; therefore, the tube grid circuits must be DC isolated.

Description of a UHF Oscillator-Modulator

The layout of an oscillator-modulator with ceramic tubes is shown in Fig.2. The oscillator is designed on the basis of a common-grid circuit. The plate-grid circuit is common to both tubes and consists of a coaxial line whose terminals are connected to the plate-grid capacitance of the tube. The oscillating voltages across the plates and the grids of both tubes are 180° out of phase. The load is connected by means of a coupling loop (4). The blocking capacitors (5) protect the plate cylinder from the DC plate voltage.

The tubes have distinct cathode-grid circuits which are tuned by means of movable pistons providing high frequency short-circuiting. The mechanism which controls the position of the piston is located inside the cathode cylinder and is not shown in Fig.2.

The tube grid cylinders are DC isolated by the blocking capacitor (1). The modulating voltage is supplied to the grid cylinders through the opening in the plate cylinder located at the voltage node.

Feedback is achieved by means of the two loop couplings (2), which are attached to the cathode-grid cylinders. The loop couplings must not short-circuit the cathode and the grid cylinders in direct current; therefore, they must be connected to the grid cylinders across the blocking capacitors (3).

The oscillator-modulator duty requires that the phase angles of the tube feedback factors are of opposite sign. For this purpose, the cathode-grid circuit impedances must be of different sign at the generating frequency. Oscillations can exist under these circumstances if the loop couplings are oriented in such a way that the voltages applied to the cathode-grid circuits are 180° out of phase. The manner in which the loop couplings should be oriented in order to sustain oscillations is shown in Fig.2.

The air for cooling the tubes is introduced through the opening located in the middle of the plate cylinder.

To illustrate the operation of the UHF oscillator-modulator, its equivalent

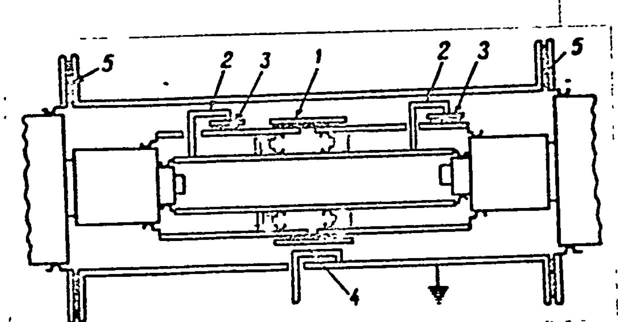

Fig.2

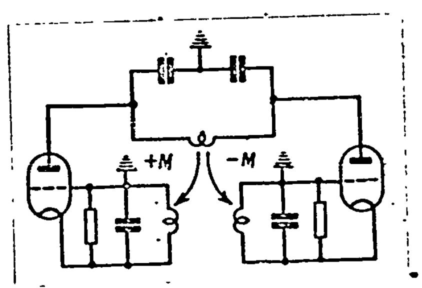

Fig.3

circuit is shown in Fig.3. This circuit shows that each of the tubes is connected to the network of a two-circuit self-excited generator. A characteristic feature of these networks is the fact that they have a common plate-grid circuit. The connection with the cathode-grid circuits is of the inductive type; the emf induced in the cathode-grid circuits is 180° out of phase. In such a three-circuit network there exist three coupling frequencies. One of these frequencies is higher than the plate-grid frequency while the other is lower. The third frequency coincides with the plate-grid circuit frequency, when the circuit is symmetrical. Only at this frequency can correct phase relationships be obtained for both parts of the network. In order that the feedback factors be complex conjugate numbers, the cathode-grid circuits must be tuned to both sides of the resonant frequency. Then, the reactances which they introduce into the plate-grid circuit will have different signs, and if the network is symmetrical, the generated frequency will coincide with the plate-grid circuit frequency.

Experimental Testing of the UHF Oscillator-Modulator

The experimental testing was carried out for an oscillator design on the basis of Fig.2. The relationships between the modulating voltage supplied 180° out of phase to the tubes, and the frequency, and between the modulating voltage and the power, were measured. The frequency was measured by means of the resonance wavemeter VST-2D. The power was measured by means of a photometric power meter, connected to the oscillator by means of a cable of known attenuation.

Figure 4 gives graphs to illustrate the operation of the network. The general character of the relationships corresponds to the calculations made on the basis of the approximate equations. A certain asymmetry of the graphs is visible. The network permits complete blocking of the tube which operates as an inductive Hartley oscillator. When the tube operating as a capacitive Hartley oscillator is blocked, the oscillations are disrupted. The asymmetry of the graphs indicates an asymmetry of the network. The experiment was carried out with GI-7B tubes, whose plate-cathode capacitance is not equal to zero. Therefore, the tube operating as an inductive Hartley oscillator requires a more powerful external feedback than the capacitive Hartley oscillator. The graphs shown in Fig.4 correspond to the case when

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one tube has two feedback loop couplings and the other three. The network operates with cathode bias, and the frequency of the generated oscillations is equal to 489 mc.

The experimental testing showed that the oscillator-modulator circuit can be used for frequency control in UHF oscillator. The total frequency deviation is of the order of one percent for a power irregularity of the order of ± 10%.

It is essential to point out that the tuning of cathode-grid circuits, necessary for normal operation of the oscillator-modulator, corresponds to tuning for maximum load power. This is due to the fact that a retuning of the cathode-grid circuits causes a change in the modules and the phase angles of the feedback factors. With the usual UHF oscillator parameters, the maximum power given up to the load corresponds to the maximum control impedance under which the phase angle of the feedback factor is close to 45^{o}. Such a feedback phase angle is completely sufficient to make the oscillator-modulator operation practicable. Thus, each of the UHF oscillator-modulator tubes must be tuned for maximum load power, and frequency control may be achieved without decreasing the power supplied by the tubes. The low value of the r-f power given in the data of Fig.4 is due to the fact that we had tubes in which the thermoelectric grid currents, forbidding normal use of the tubes on current were significant.

The above-described oscillator-modulator design can be used on wavelengths greater than 60 cm. As the wave becomes shorter, difficulties arise in tuning the cathode-grid circuit of the oscillator operating as an inductive Hartley oscillator. The point is that such a circuit must have a length shorter than the resonant length so that its impedance can be of the inductive type at the self-oscillation frequency. On waves shorter than 60 cm, the necessary length of the GI-7B tube cathode circuit is so small that it cannot be constructed in practice. Shorter waves require changeover to the use of cathode line overtones, which cannot be done in the described design. In such a case, the use of a design in which cathode-grid circuits are arranged on different sides of the plate-grid circuit, is preferable.

It should be mentioned that, in discussing the oscillator-modulator operation, we idealized the problem and neglected the inertia of the electrons. Due to the inertia of the electrons, the slope of the plate current becomes complex, which causes the steady-state phase balance to change. The electron inertia affects the oscillator operation in various ways, depending on the oscillator circuit. In the capacitive Hartley oscillator, the phase angle of the slope compensates the feedback phase angle; as a result, at small angles, the condition of self-excitation becomes easier to satisfy. In the inductive Hartley oscillator, the phase angle of the slope and phase angle of the feedback have the same sign; this makes self-excitation more difficult. Apparently, the asymmetry in the oscillator-modulator operation, which was pointed out above, is not merely due to the asymmetry of the circuit. The asymmetry of the circuit, which is due to the plate-cathode capacitance of the tube, may be eliminated by a suitable choice of the size and number of feedback loops. This problem was given considerable attention but complete elimination of the asymmetry of operation was impossible. Apparently, the obtained asymmetry of operation is indicative of the effect of the electron inertia. As the wave is shortened, the transconductance will increase and the operation asymmetry will increase.

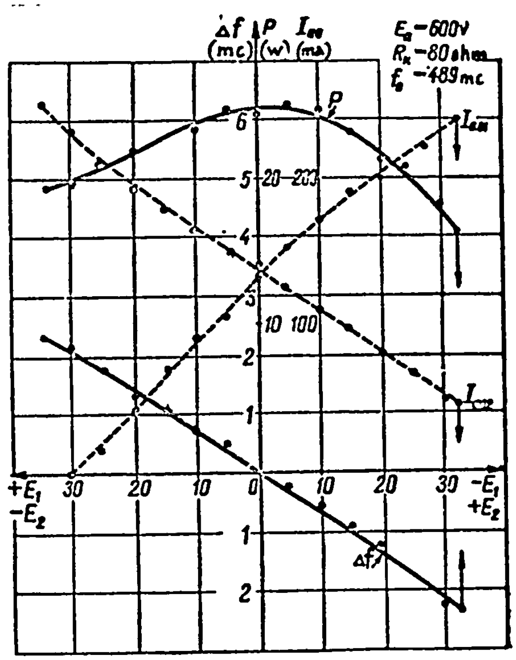

Fig.4

It follows from the above that, from the viewpoint of design as well as from that of the electronic performance, the proposed hookup can be recommended for oscillators which operate in the upper portion of the decimeter wave range.

V.Ivanov and I.Seversky, Engineers, took part in the realization of the given project.

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DESIGNING AN OSCILLATOR OPERATING AT OVERVOLTAGE AND DETUNED LOAD

by

E.P.Khmelnitsky

An account is given of a method for the engineering design of an oscillator operating at overvoltage and with detuned load, in order to improve efficiency.

The general performance characteristic of an oscillator which operates with a detuned load at a considerably overvoltage has been discussed in two papers (Bibl.1,2).

The present article gives an account of the engineering computations for such a performance which is advantageous from the energetics viewpoint.

The current-voltage relations in the plate and grid circuits of an oscillator which operates in the above-mentioned state, form the basis for the method of computation described here. The character of these relations is plotted in Fig.1, Careful consideration of the diagram will show that a given displacement of the gap to the right of its center in the plate current pulse is accompanied by such a phasing of currents and voltages in the oscillator plate circuit that, at the instant corresponding to the amplitude of the positive grid voltage, i.e., at $\omega t = 0$, the negative plate voltage $E_o - U_1 \cos \varphi_{u1}$ is compensated by the voltage of the second and third harmonics.

By virtue of this, the residual plate voltage at $\omega t = 0$

$$E_0 - U_1 \cos \varphi_{u1} + U_2 \cos \varphi_{u2} + U_3 \cos \varphi_{u3}$$

has a positive sign, despite the fact that the plate voltage efficiency is $\xi > 1$.

Under certain conditions, the residual voltage may be higher than the positive grid voltage at that instant

$$E_g + U_g.$$

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As a result, the plate negative voltage region and, consequently, the zone of deep trough in the plate current pulse, are displaced to the left of the origin of coordinates. The displacement of the center of the trough is indicated in Fig.1 by

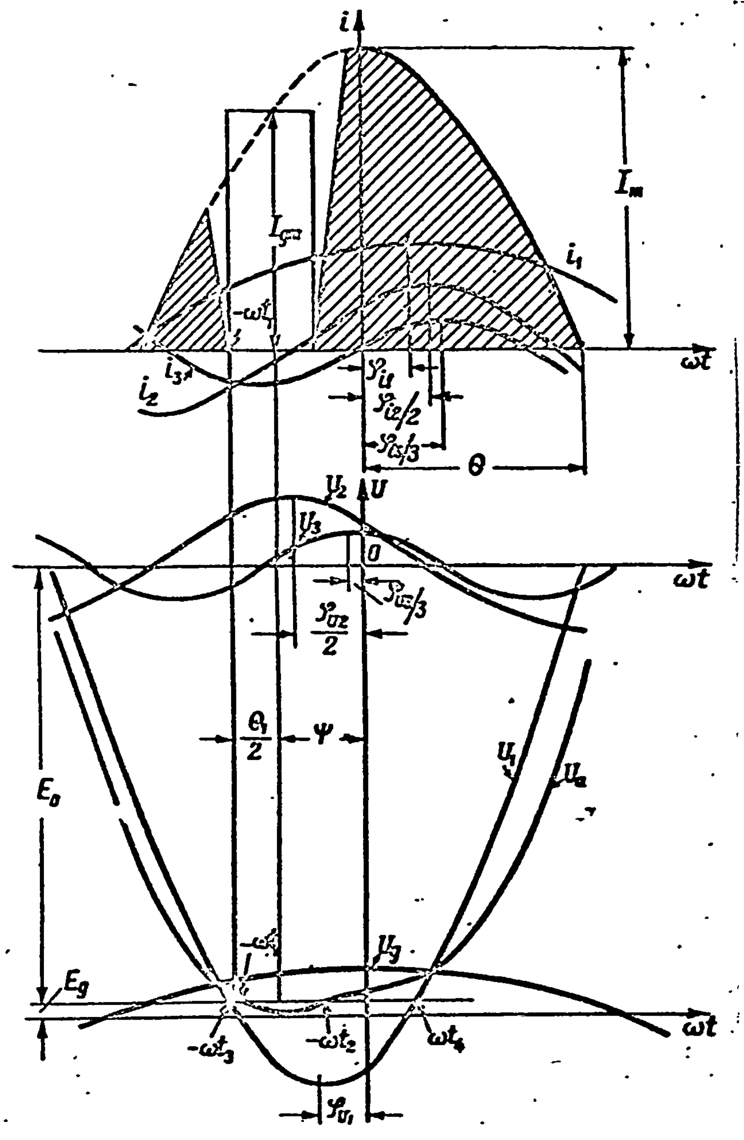

Fig.1

the angle ψ, while the width of the trough, limited by the instants $-\omega t_1$ and $-\omega t_2$, is denoted by the angle θ_1.

The displacement of the pulse trough yields a number of qualitative results. First of all, a relatively rapid increase of the coefficient of the first harmonic

$\alpha_1 = \frac{I_1}{I_m}$ (Fig.2) takes place. Here and further on, I_1 denotes the current amplitude of the fundamental frequency, a current which is a component of the pulse, while I_m denotes the maximum value of the current pulse as indicated in Fig.1. Then, the coefficient of the direct component increases considerably more slowly. As a result, the coefficient of the form $\gamma = \frac{\alpha_1}{\alpha_0}$, when the trough is displaced by an angle of only $\psi \approx 30^\circ$ (Fig.3), attains the value of this coefficient if the pulse is cosinusoidal.

A second characteristic feature of a pulse with an asymmetrical trough, is a certain phase shift equal to the angle φ_{u1} of the voltage amplitude of the first harmonic. This facilitates the task of compensation by means of plate voltage harmonics of negative potential, at the instant $\omega t = 0$.

Finally, the advantageous change in the amplitude and phase relationship for

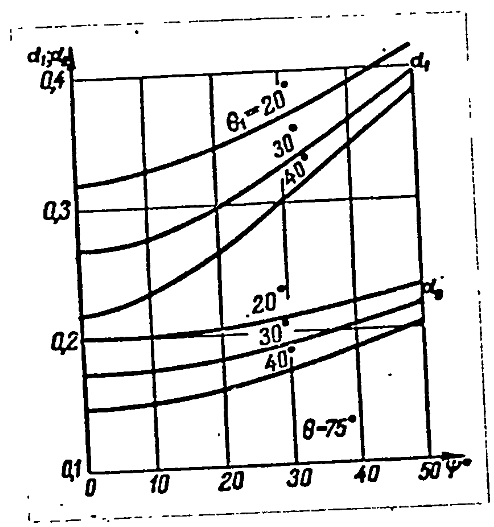

Fig.2

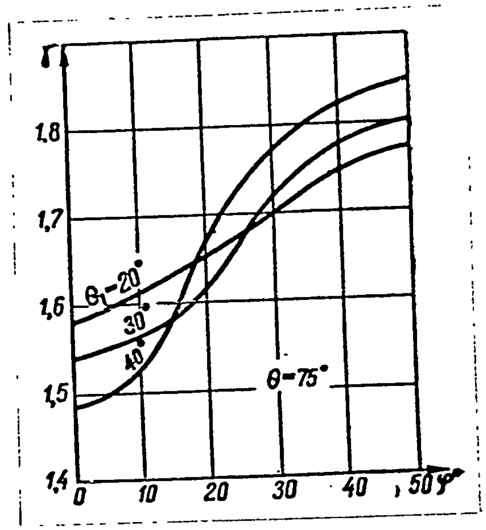

Fig.3

the voltage of the second and third harmonics, from the viewpoint of compensation of the negative plate voltage mentioned above, is an important fact associated with the leftward displacement of the trough. If this were not so, the width of the trough would have been determined by the instants $-\omega t_3$ and ωt_4. The capacitive branch of the circuit serves as the oscillator plate circuit load for these harmonics.

In networks in which the feeding is in parallel, the load for the highest har-

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monics consists of the blocking capacitor C_p and of the capacitance of the circuit C_k, connected in series

$$C=\frac{C_p C_\kappa}{C_p+C_\kappa}.$$

The total voltage of the second and third harmonics may be determined as follows, at $\omega t = 0$:

$$U_2+U_3=\frac{\alpha_2 I_m}{2\omega C}\cos\varphi_{u2}+\frac{\alpha_3 I_m}{3\omega C}\cos\varphi_{u3}=\frac{I_m}{\omega C}\left(\frac{\alpha_2}{2}\cos\varphi_{u2}+\frac{\alpha_3}{3}\cos\varphi_{u3}\right).$$

An investigation of the plate current pulses, at various values of the trough width and shift at various lower cutoff angles of the plate current showed that the expression

$$\frac{\alpha_2}{2}\cos\varphi_{u2}+\frac{\alpha_3}{3}\cos\varphi_{u3}$$

always has a maximum when the trough shifts to the left of the pulse center by an angle of $\psi \approx 30^\circ$, as shown, for instance, in Fig.4 for a cutoff angle of $\theta = 75^\circ$.

It follows from the above that $\psi = 30^\circ$ is the optimum trough shift angle and that all computations for an oscillator of the described duty, should be based on the proposition that $\psi = 30^\circ$.

The computational graphs of the breakdown factors α_1 and α_0, as well as the form factor γ, as a function of the cutoff angle θ, are given in Figs.5 and 6 for trough of width

$$\theta_1 = 20^\circ,\ 30^\circ \text{ and } 40^\circ.$$

Figure 7 gives the relationship of the quantity

$$\frac{\alpha_2}{2}\cos\varphi_{u2}+\frac{\alpha_3}{3}\cos\varphi_{u3}$$

for the same values of θ and θ_1.

All these graphs are based on $\psi = 30^\circ$ and may serve as a basis for the engineering computations of the operating conditions since, if the shape of the pulse, i.e., angles θ and θ_1, are given, the breakdown factor of the pulse can be determined.

Computations of the breakdown factors and of the phase angles were carried out by means of the well-known graphical method with a 5° shift of the ordinates. Besides, the form of the trough was assumed to be based upon θ_1 but to have sides that are not vertical but rather inclined by 5° from the top on both sides (see Fig.1).

The selected plate current pulse shape is obtained, in our case, without determination of the usual factors, such as the magnitudes of the DC plate-grid voltages and of the AC voltage of the first harmonic, by satisfying two additional conditions:

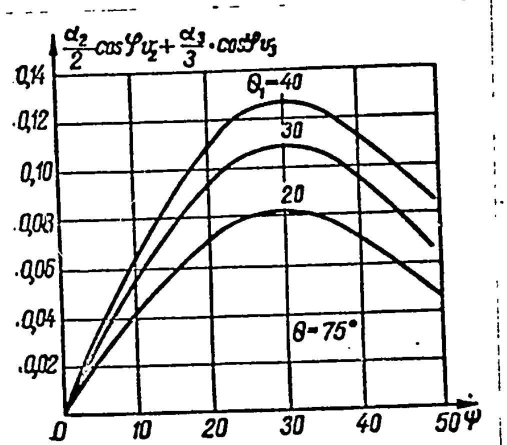

Fig.4

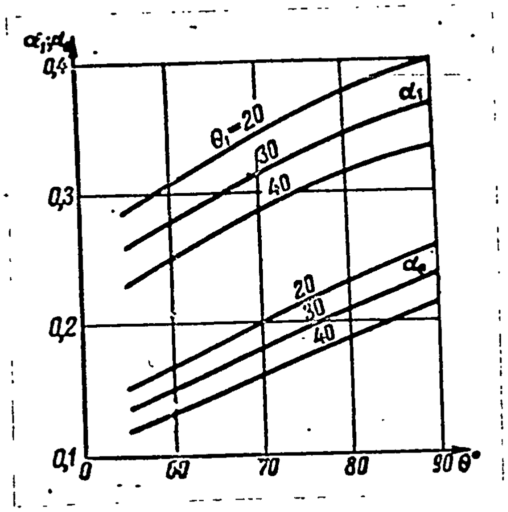

Fig.5

We must consider that the first condition is to obtain the required magnitude of total voltage of the highest harmonics.

The required magnitude of this voltage may be determined from the consideration that, at the instant $\omega t = 0$, the total plate voltage of the AC components must have a magnitude below E_0.

A sufficient condition for undervoltage at that instant is the inequality

$$U_1 \cos \varphi_{u1} - \frac{I_m}{\omega C}\left(\frac{\alpha_2}{2}\cos\varphi_{u2} + \frac{\alpha_3}{3}\cos\varphi_{u3}\right) \leqslant 0{,}8E_0.$$

The required magnitude of the plate circuit capacitance which will satisfy the above-indicated first condition of pulse generation is thus determined.

In this formula, the phase angle of the fundamental frequency φ_{u1} voltage, has always a value close to 15°, as shown by investigations at an optimum angle of trough shift $\psi = 30°$. It may safely be assumed that $\varphi_{u1} = 15°$, without interfering with the accuracy of the computations.

A second condition which will secure the given pulse shape is the choice of the required magnitude of the utilization factor of the plate voltage.

It may be seen from Fig.1 that the left-hand limit of the pulse trough is the instant

$$-\omega t_1 = \psi + \frac{\theta_1}{2}.$$

Since

$$\omega t_1 \approx \omega t_3,$$

it may be considered that

$$E_0 \approx U_1 \cos\left(\psi + \frac{\theta_1}{2} - \varphi_{u1}\right).$$

Hence, a formula

$$\xi = \frac{U_1}{E_0} = \frac{1}{\cos\left(\psi + \frac{\theta_1}{2} - \varphi_{u1}\right)},$$

is obtained, whose final form, after substituting the values assumed earlier into it, reads as follows:

$$\xi = \frac{1}{\cos\left(\frac{\theta_1}{2} + 15°\right)}.$$

The degree of plate circuit detuning required in order to obtain the type of duty described, may be calculated by means of a formula derived from known relationships (3), namely

$$Z_1 = \frac{1}{2C\sqrt{\delta^2 + (\Delta\omega)^2}},$$

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$$\delta = \frac{\Delta\omega}{\operatorname{tg}\varphi_1}.$$

If the expression for δ is substituted from the second equality into the first, we will obtain

$$\Delta\omega = \frac{5 \cdot 10^{11}}{C_k Z_1} \sin\varphi_1.$$

Here C_k is the circuit capacitance in micromicrofarads.

The circuit impedance

$$Z_1 = \frac{U_1}{I_1}$$

is determined in computing the oscillator, while

$$\varphi_1 = \varphi_{k1} - \varphi_{i1},$$

where φ_{i1} is a function of the angles θ and θ_1 and is determined from the graph in Fig.8.

The natural calculated frequency of the circuit is

$$f_k = f_p + \Delta f,$$

where $\Delta f = \frac{\Delta\omega}{2\pi}$,

while f_ρ is the given oscillator frequency.

The design formula for the grid excitation is easily determined from the usual

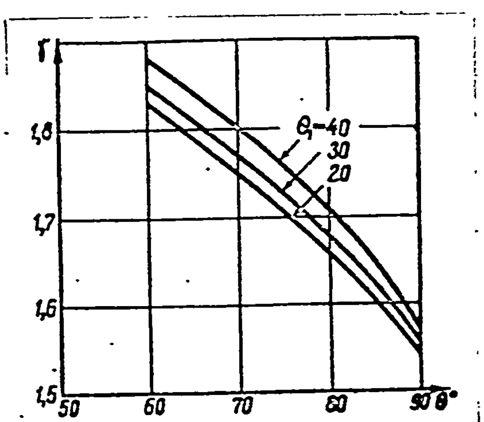

Fig.6

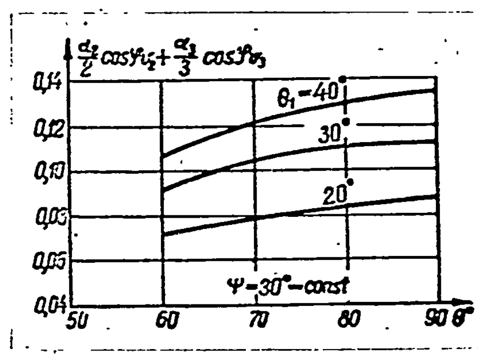

Fig.7

formula

$$U_g = \frac{I_m}{S(1-\cos\theta)} + DU_1$$

by substituting the above-assumed first approximation value of the total voltage of the alternating components for DU_1, which determines the plate voltage reactance. Then,

$$U_g = \frac{I_m}{S(1-\cos\theta)} + 0{,}8DE_0.$$

Accordingly, the grid bias will be

$$E_g = E_{g0} - DE_0 - U_y \cos\theta,$$

where the control voltage is

$$U_y = U_g - 0{,}8DE_0.$$

The following assumptions were made for the purpose of determining the grid current. The grid current pulse was assumed to coincide in time with plate current pulse trough and to have a rectangular shape, as shown in Fig.1. Such an assumption permits computing the grid current by quite simple formulas, with sufficient accuracy. From Fig.1, we have

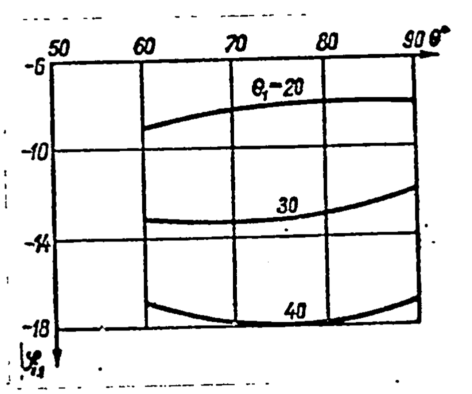

Fig.8

$$I_{g0} = \frac{I_{gm}\theta_1}{2\pi},$$

where

$$I_{gm} = S\{E_g + U_g\cos\psi - E_{g0} + D[E_0 + U_1\cos(\psi - \varphi_{u1}) + U_2\cos(2\psi - \varphi_{u2}) + U_3\cos(3\psi - \varphi_{u3})]\}.$$

If the last term is neglected, due to its small magnitude, we obtain

$$I_{gm}=S(E_g+U_g\cos\psi-E_{g0}).$$

In order to determine the power consumed by the grid circuit, we assume that the approximate value of the amplitude of the first harmonic of the grid current is equal to

$$I_{g1}\approx 1{,}8\, I_{g0}.$$

The power is equal to

$$P_g=\frac{I_{g1}U_g}{2}\cos\psi.$$

In the case of an oscillator with plate modulation, the maximum power consumption in the grid circuit takes place at so-called zero operating conditions, when E_0 and U_1 are equal to zero.

At that instant, the grid-current curve adopts a nearly cosinusoidal slope, and the maximum value of the pulse may be determined from

$$I_{gm}=SU_g(1-\cos\theta_g),$$

where

$$\cos\theta_g=-\frac{E_g-F_{g0}}{U_g}.$$

The coefficient α_{g1} (see Tables for cosinusoidal pulses) is used for calculating

$$I_{g1}=\alpha_{g1}I_{gm},$$
$$P_g=\frac{U_gI_{g1}}{2}.$$

A tentative selection of tubes for the described duty should be carried out on the basis of the emission current

$$I_e\geqslant 6\frac{P_{-m}}{E_a}.$$

It is beyond the scope of this article to discuss the practical applications of the formulas derived here for various cases of oscillator computations. We will therefore limit the discussion to computations for the case when the oscillator output power and the working wave are given.

Given: $P_{\sim} = 150$ kw; $\lambda = 400$m.

A suitable tube with respect to the emission current is the GU-23A; two tubes will be used in parallel in accordance with the single-cycle circuit.

Tube Data

The filament voltage is 12v, the filament current 210 amp, the plate voltage $E_0 = 11$ kv, the permissible plate power leakage $P_a = 60$ kw and for the control grid, $P_g = 2.6$ kw.

The emission current is $I_e = 70$ amp. The steepness of the characteristic is $S = 42$ ma/v.

The permeability is $D = 0.0189$. The starting voltage is $E_{g0} = 120$ v.

If the cutoff angle is given as $\theta = 75^\circ$ and the trough width as $\theta_1 = 40^\circ$, the graphs for the trough shift $\psi = 30^\circ$, in accordance with Figs.5,6,7, and 8, will yield $\alpha_1 = 0.3$, $\alpha_0 = 0.17$, $\gamma = 1.765$, $\left(\frac{\alpha_2}{2}\cos\varphi_{u2} + \frac{\alpha_3}{3}\cos\varphi_{u3}\right) = 0.126$ and $\varphi_{i1} = -18^\circ$.

Computation Sequence

$$\varphi_1 = \varphi_{u1} - \varphi_{i1} = 15^\circ + 18^\circ = 33^\circ,$$

$$\xi = \frac{1}{\cos\left(\frac{\theta_1}{2} + 15^\circ\right)} = 1{,}22,$$

$$\eta = \frac{\xi\gamma}{2}\cos\varphi_1 = 90{,}5\%, \text{ so that } \cos\varphi_1 = 0{,}84,$$

$$U_1 = \xi E_0 = 13{,}4 \text{ kv};$$

$$I_1 = \frac{2P_{\sim}}{U_1 \cos\varphi_1} = 26{,}7 \text{ amp},$$

$$I_m = \frac{I_1}{\alpha_1} = 89 \text{ amp},$$

$$I_0 = \alpha_0 I_m = 15{,}1 \text{ amp},$$

$$P_0 = I_0 E_0 = 166 \text{ kw},$$

$$P_a = P_0 - P_{\sim} = 16 \text{ kw},$$

$$Z_1 = \frac{U_1}{I_1} = 500 \text{ ohm}$$

$$U_g = \frac{I_m}{S(1 - \cos\theta)} + 0{,}8DE_0 = 1600 \text{ v},$$

$$U_y = U_g - 0{,}8DE_0 = 1400 \text{ v},$$

$$E_g = E_{g0} - DE_0 - U_y \cos = -450 \text{ v},$$

$$I_{gm} = S(E_g - E_{g0} + 0{,}85U_g) = 65 \text{ amp},$$

$$I_{g0} = \frac{I_{gm}\theta_1}{2\pi} = 7 \text{ amp}.$$

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The resistance in the grid circuit, when automatic grid bias is used, must be of the magnitude

$$R_g = \frac{E_g}{I_{g0}} = 65\ ohm$$

This resistance will consume the power

$$P_{g0} = I_{g0}^2 R_g = 3{,}2\ \kappa w\ .$$

The total power consumed in the grid circuit will be found by means of

$$I_{g1} = 1{,}8 I_{g0} = 12{,}6\ amp$$

and will be equal to

$$P_g = \frac{I_g U_g}{2} \cos\xi = 6{,}9\ \kappa w\ .$$

The power obtained earlier, which is produced for automatic bias resistance, is a portion of this power; the rest is dissipated on the grid.

The required magnitude of the plate circuit capacitance, being the load for the highest harmonics, must not be higher than

$$C = \frac{C_p C_\kappa}{C_p + C_\kappa} = \frac{I_m\left(\frac{a_2}{2}\cos\varphi_{u2} + \frac{a_3}{3}\cos\varphi_{u3}\right)}{\omega(U_1\cos\varphi_{u1} - 0{,}8E_0)} = 570\ \mu\mu f$$

The magnitude of the blocking capacitance is determined from

$$\frac{1}{\omega C_p} = 0{,}4 Z_1$$

and is equal to

$$C_p = \frac{530\lambda}{0{,}4 Z_1} = 1000\ \mu\mu f;$$

while the circuit capacitance, including the capacitance of the mounting is equal to

$$C_\kappa = \frac{C_p C}{C_p - C} = 1330\ \mu\mu f.$$

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The magnitude of the detuning is determined from

$$\Delta\omega = \frac{5\cdot 10^7}{C_\kappa Z_1}\sin\varphi_1 = 4,1\cdot 10^5,$$

so that

$$\Delta f = \frac{\Delta\omega}{2\pi} = 65\ \kappa c .$$

Consequently, the calculated magnitude of the frequency for determining the circuit inductance is

$$f_\kappa = f_p + \Delta f = 815\ \kappa c .$$

The circuit inductance is calculated by means of the formula

$$L_\kappa = \frac{\lambda^2}{3,55 C_\kappa}\frac{f_p}{f_\kappa} = 31,2\ \mu h .$$

which is convenient for calculation.

The total active resistance of the circuit becomes

$$R_\kappa = \frac{10^6 L_\kappa}{C_\kappa Z_1}\cos\varphi_1 = 39,4\ ohm .$$

The obtained network data, besides being easily realizable, also permit a further improvement in the conditions necessary for the generation of a voltage of the highest harmonics at the expense of a decrease in the circuit capacitance and the corresponding change in the inductance and active resistance.

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USE OF FERRITES IN WAVEGUIDE ENGINEERING

by

A.K.Stolyarov

The basic phenomena taking place in waveguides containing ferrites are considered in this survey.

Introduction

In recent years, investigations have led to the development of a new class of nonmetallic ferromagnetic materials, the ferrites. Ferrites differ from metallic ferromagnetic substances in that they possess a very high resistivity (10^6 to 10^8 ohm-cm) so that they may be grouped with the class of ferromagnetic semiconductors.

Since electromagnetic waves propagate through nonmagnetized ferrites, the same phenomena will take place as in isotropic media. If, in an infinite ferritic medium, a permanent magnetic field acts along the z-axis, the magnetic permeability of such a medium is no longer described by a scaler quantity, but rather by an asymmetric tensor (Bibl.1) of the form

$$[\mu]=\begin{vmatrix} \mu & -ik & 0 \\ ik & \mu & 0 \\ 0 & 0 & \mu_z \end{vmatrix}. \qquad (1)$$

Here μ and k are complex quantities which depend on the constant magnetic field intensity H_z (Bibl.1,2), while k depends also on the direction of H_z. If the losses in the ferrite are neglected, the formulae for μ and for k will be written as

$$\frac{\mu}{\mu_0}=\frac{\mu_0\gamma^2 B_z H_z-\omega^2}{\mu_0^2\gamma^2H_z^2-\omega^2}, \quad B_z=\mu_0(H_z+M_z),$$
$$\frac{k}{\mu_0}=-\frac{\mu_0|\gamma|M_z\omega}{\mu_0^2\gamma^2H_z^2-\omega^2}, \qquad (2)$$

where

$$\mu_0=4\pi\cdot10^{-7}\,\frac{гн}{м},$$

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M_z is the magnetization of the ferrite;

$\gamma = 1.78 \times 10^{11}$ coul/kg, the ratio of the electron charge to its mass.

Infinite Ferritic Medium

If Maxwell's equation is solved for the case of the propagation of electromagnetic waves in the infinite ferritic medium which is magnetized in the direction of the z-axis, it will turn out that (Bibl.2), in the arbitrary direction s (Fig.1), two waves with different propagation constants γ_1 and γ_2 will propagate independently

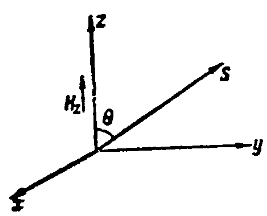

Fig.1

$$\gamma_{1,2}^2 = \frac{k_z^2}{2} \cdot \frac{\left[\frac{\mu^2}{\mu_z^2} - \frac{\mu}{\mu_z} - \frac{k^2}{\mu_z^2}\right]\sin^2\theta + 2\frac{\mu}{\mu_z} \pm \sqrt{\left[\frac{\mu^2}{\mu_z^2} - \frac{\mu}{\mu_z} - \frac{k^2}{\mu_z^2}\right]^2 \sin^4\theta + 4\frac{k^2}{\mu_z^2}\cos^2\theta}}{\left(\frac{\mu}{\mu_z} - 1\right)\sin^2\theta + 1}, \quad (3)$$

where

$$k_z^2 = \omega^2 \varepsilon \mu_z.$$

The case of wave propagation along and perpendicular to a constant magnetic field and is of the greatest practical interest.

Case of the Longitudinal Magnetic Field. If $\theta = 0$ is substituted in eq.(3), we obtain

$$\gamma_1^2 = \omega^2 \varepsilon (\mu + k) = \omega^2 \varepsilon \left(1 - \frac{\mu_0 |\gamma| M_z}{\mu_0 |\gamma| H_z - \omega}\right), \quad (4)$$

$$\gamma_2^2 = \omega^2 \varepsilon (\mu - k) = \omega^2 \varepsilon \left(1 - \frac{\mu_0 |\gamma| M_z}{\mu_0 |\gamma| H_z + \omega}\right). \quad (5)$$

Equations (4) and (5) show that, along the z-axis, two waves with different phase velocities are propagated. A consideration of the field structure of these waves (3) leads to the conclusion that these are two plane waves with circular polarization, the wave with the propagation constant γ_1 being dextropolarized and the wave with the propagation constant γ_2 being levopolarized. Thus, a wave with any

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M_z is the magnetization of the ferrite;

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Infinite Ferritic Medium

If Maxwell's equation is solved for the case of the propagation of electromagnetic waves in the infinite ferritic medium which is magnetized in the direction of the z-axis, it will turn out that (Bibl.2), in the arbitrary direction s (Fig.1), two waves with different propagation constants γ_1 and γ_2 will propagate independently

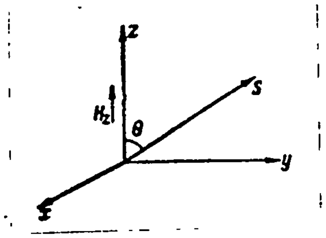

Fig.1

$$\gamma_{1,2}^2 = \frac{k_z^2}{2}\cdot\frac{\left[\frac{\mu^2}{\mu_z^2}-\frac{\mu}{\mu_z}-\frac{k^2}{\mu_z^2}\right]\sin^2\theta+2\frac{\mu}{\mu_z}\pm\sqrt{\left[\frac{\mu^2}{\mu_z^2}-\frac{\mu}{\mu_z}-\frac{k^2}{\mu_z^2}\right]^2\sin^4\theta+4\frac{k^2}{\mu_z^2}\cos^2\theta}}{\left(\frac{\mu}{\mu_z}-1\right)\sin^2\theta+1}, \qquad (3)$$

where

$$k_z^2 = \omega^2\varepsilon\mu_z.$$

The case of wave propagation along and perpendicular to a constant magnetic field and is of the greatest practical interest.

Case of the Longitudinal Magnetic Field. If $\theta = 0$ is substituted in eq.(3), we obtain

$$\gamma_1^2 = \omega^2\varepsilon(\mu + k) = \omega^2\varepsilon\left(1 - \frac{\mu_0|\gamma|M_z}{\mu_0|\gamma|H_z - \omega}\right), \qquad (4)$$

$$\gamma_2^2 = \omega^2\varepsilon(\mu - k) = \omega^2\varepsilon\left(1 - \frac{\mu_0|\gamma|M_z}{\mu_0|\gamma|H_z + \omega}\right). \qquad (5)$$

Equations (4) and (5) show that, along the z-axis, two waves with different phase velocities are propagated. A consideration of the field structure of these waves (3) leads to the conclusion that these are two plane waves with circular polarization, the wave with the propagation constant γ_1 being dextropolarized and the wave with the propagation constant γ_2 being levopolarized. Thus, a wave with any

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polarization which is propagated along the constant magnetic field H_z, breaks up into two waves with circular polarization, which have different phase velocities. If the initial wave is linearly polarized, then as it is propagated in the ferrite, its plane of polarization will be constantly rotating; the direction of this rotation is determined by the direction of the permanent magnetic field H_z and does not depend on the direction of propagation of the wave. In this and similar phenomena of magnetic rotation of the plane of polarization, the reciprocity principle (Faraday's effect) does not apply.

The second magneto-optical effect [see eqs.(4) and (5)] consists in the fact that, for a certain constant field intensity,

$$H_z = \frac{\omega}{\mu_0 |\gamma|} \tag{6}$$

resonant absorption of the dextropolarized wave (γ_1) is observed, whereas the absorption of the levopolarized wave (γ_2) is almost independent of the quantity H_z (Bibl.4).

The indicated magneto-optical phenomena have been investigated in various papers (Bibl.2,5,7). These are used as basis in making waveguide valves, switches, directional phase shifters in circular waveguides.

Figures 2A and 2B give the schematic diagram and the operating principle of a waveguide system which makes use of Faraday's effect. Rectangular waveguides placed on both sides of the circular waveguide, which contains a "shifting ferritic element", form an angle of 45°.

The wave from the rectangular waveguide (1) enters the spherical waveguide (Fig.2B,a,b). The rotating element rotates the plane of polarization through an angle of 45° (Fig.2B c) and the wave passes into the turned rectangular waveguide(2) (Fig.2B d). When propagation in the reverse direction takes place, as can be seen from Fig.2B d-e, the wave will arrive with such a polarization that it will be unable to propagate in the waveguide (1) and may be deflected into the waveguide (3).

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Thus, a system is obtained in which energy is transmitted from channel (1) to channel (2), and from (2) to (3).

As the direction of the permanent field H_z changes, energy transfer will take

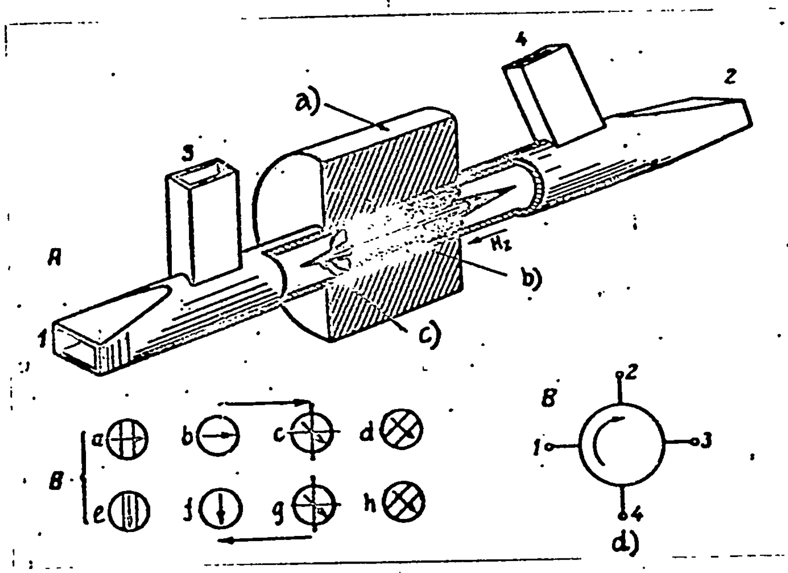

Fig.2

a) Permanent magnet; b) Phenoplast; c) Ferrite rod; d) Described circulator

place from (1) to (4), from (4) to (3), from (3) to (2).

Such a system, which accomplishes the above-indicated channel commutation, is called a circulator. A schematic diagram of a circulator is shown in Fig.2C. By means of shifting the direction of the external field H_z, energy can be transmitted from channel (1) alternately into channel (2) or (4), i.e., the system maybe used as a switch

At present, the above circuit is used as basis for designing rapid-action switches. The basic difficulties in the development are the creation of a broad-band ferritic element which would operate at a very low permanent field intensity H_z, as well as the creation of a circular waveguide which would allow a rapid change-over of the direction of the field H_z.

Figure 3 shows the schematic diagram of a valve which makes use of resonant absorption of the dextropolarized wave by the ferrite. The wave in the cylindrical waveguide (1) (Fig.3a) is transformed into a levopolarized wave and passes through the ferrite with little damping. The wave, which propagates from the waveguide (2) to the waveguide (1), is transformed into a dextropolarized wave and is then strongly absorbed by the ferrite.

Valve systems with very low coefficients of reflection (TWR > 0.95) are of greatest interest for radio-relay lines.

Tuning the systems in question to TWR > 0.95 is quite difficult. Therefore, phenomena which take place in a rectangular waveguide with a magnetized ferrite will be considered in detail below, since their use permits designing simpler valve systems which would also have better qualitative indexes.

Case of the Transverse Magnetic Field. If $\theta = 90^\circ$ is substituted into eq.(3), the following will be obtained for the propagation constants of two independent waves:

$$\gamma_1^2 = k_\perp^2 = \omega^2\varepsilon\frac{\mu^2 - k^2}{\mu} = \omega^2\varepsilon\mu_\perp, \qquad (7)$$

$$\gamma_2^2 = k_z^2 = \omega^2\varepsilon\mu_z, \qquad (8)$$

where

$$\mu_\perp = \frac{\mu^2 - k^2}{\mu}.$$

Analysis of the field structure of these waves (Bibl.3) shows that the first one (extraordinary wave) has a longitudinal magnetic field component; the wave propagation constant depends on the permanent field intensity H_z.

The second wave (ordinary wave) is a transverse wave and its propagation constant is almost independent of the external field H_z.

Thus, in the case of a transverse field, a wave of any polarization breaks up into two linearly polarized waves with different propagation constants.

The second magneto-optical phenomenon consists in a resonant absorption of the

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extraordinary wave at a definite value of the external field H_z.

Actually, if μ and k in eq.(7) for $\mu_\perp$ are replaced by their values from eq.(2), it will be found that $\mu_\perp \to \infty$ at

$$\omega_\perp = \omega_0 \sqrt{1 + \frac{M_z}{H_z}}, \qquad (9)$$

where $\omega_0 = \mu_0 |\gamma| H_z$ is the longitudinal gyromagnetic resonant frequency; $\omega_\perp$ is the transverse gyromagnetic resonant frequency.

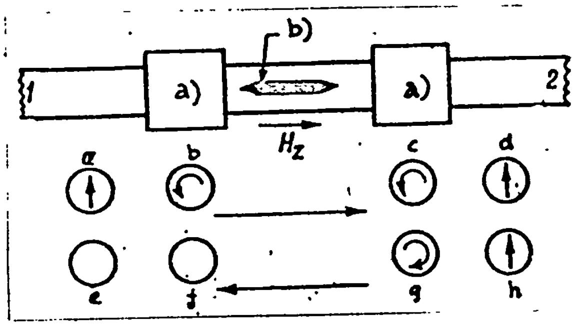

Fig.3

a) Polarizer; b) Ferrite

Magneto-optical Phenomena in Rectangular Waveguides which Contain a Transversely Magnetized Ferrite.

The problem of electromagnetic wave propagation in a waveguide with a magnetized ferrite was solved by the partial wave method (Bibl.3,6). The equation which determines the wave propagation constants along the y-axis (Fig.4) and the expression for the field component, in the case of type H_{n0} wave propagation, will have the form

$$\gamma_y \frac{k}{\mu}(\mathrm{tg}\, k_{0x}t_3 - \mathrm{tg}\, k_{0x}t_2)\,\mathrm{tg}\, k_{\perp x}t_1 + (\mathrm{tg}\, k_{0x}t_2 + \mathrm{tg}\, k_{0x}t_3)\, k_{\perp x} + k_{0x}\,\mathrm{tg}\, k_{\perp x}t_1 \left[\frac{\mu_\perp}{\mu_0} - \frac{\mu_0}{\mu_\perp k_{0x}^2}\left(k_{\perp x}^2 + \frac{k^2}{\mu^2}\gamma_y^2\right)\mathrm{tg}\, k_{0x}t_2\,\mathrm{tg}\, k_{0x}t_3\right] = 0, \qquad (10)$$

where

$$k_{\perp x}^2 = k_\perp^2 - \gamma_y^2; \quad k_{0x}^2 = k_0^2 - \gamma_y^2; \quad k_0^2 = \omega^2 \varepsilon_0 \mu_0.$$

In the region I (Fig.4):

$$E_{zI} = -iA\left[\sin k_{0x}t_2 \cos k_{\perp x}x + \left(\frac{k_{0x}}{k_{\perp x}}\frac{\mu_\perp}{\mu_0}\cos k_{0x}t_2 - \frac{k}{\mu}\frac{\gamma_y}{k_{\perp x}}\sin k_{0x}t_2\right)\sin k_{\perp x}x\right] e^{-i\gamma_y y};$$

$$h_{yI} = A\frac{\omega\varepsilon}{k_\perp^2}\left[\left(\frac{k^2}{\mu^2}\frac{\gamma_y^2}{k_{\perp x}^2} + 1\right)k_{\perp x}\sin k_{0x}t_2 \sin k_{\perp x}x - \right.$$

$$-k_{0x}\frac{\mu_\perp}{\mu_0}\cos k_{0x}t_2\left(\frac{k}{\mu}\frac{\gamma_y}{k_{\perp x}}\sin k_{\perp x}x+\cos k_{\perp x}x\right)\Big]\,e^{-i\gamma_y y}.$$

$$h_{xI}=-iA\frac{\omega\varepsilon}{k_\perp^2}\Big[\left(\gamma_y\cos k_{\perp x}x-\frac{k}{\mu}k_{\perp x}\sin k_{\perp x}x\right)\sin k_{0x}t_2+$$
$$+\left(\frac{k_{0x}}{k_{\perp x}}\frac{\mu_\perp}{\mu_0}\cos k_{0x}t_2-\frac{k}{\mu}\frac{\gamma_y}{k_{\perp x}}\sin k_{0x}t_2\right)\left(\gamma_y\sin k_{\perp x}x+\frac{k}{\mu}k_{\perp x}\cos k_{\perp x}x\right)\Big]e^{-i\gamma_y y}. \quad (11)$$

In the region II:

$$E_{zII}=-iA\sin k_{0x}(x+t_2)\,e^{-i\gamma_y y};$$
$$h_{yII}=-A\frac{\omega\varepsilon_0}{k_0^2}k_{0x}\cos k_{0x}(x+t_2)\,e^{-i\gamma_y y};$$
$$h_{xII}=-iA\frac{\omega\varepsilon_0}{k_0^2}\gamma_y\sin k_{0x}(x+t_2)\,e^{-i\gamma_y y}. \quad (12)$$

In the region III:

$$E_{zIII}=iB\sin k_{0x}(x-t_4)\,e^{-i\gamma_y y};$$
$$h_{yIII}=B\frac{\omega\varepsilon_0}{k_0^2}k_{0x}\cos k_{0x}(x-t_4)e^{-i\gamma_y y};$$
$$h_{xIII}=iB\frac{\omega\varepsilon_0}{k_0^2}\gamma_y\sin k_{0x}(x-t_4)\,e^{-i\gamma_y y}; \quad (13)$$

where

$$B=A\Big[\sin k_{0x}t_2\cos k_{\perp x}t_1+$$
$$+\left(\frac{\mu_\perp}{\mu_0}\frac{k_{0x}}{k_{\perp x}}\cos k_{0x}t_2-\frac{k}{\mu}\frac{\gamma_y}{k_{\perp x}}\sin k_{0x}t_2\right)\sin k_{\perp x}t_1\Big]\operatorname{cosec}k_{0x}t_3.$$

It follows from eq.(10) that a change in direction of the permanent field H_z or in direction of the wave propagation will cause a change in the magnitude of the propagation constant γ_y, i.e., the phase velocities as well as the damping of waves propagating in opposite directions differ. This, in turn, means [eqs.(11-13)] that, as the direction of propagation changes, the field structure of the

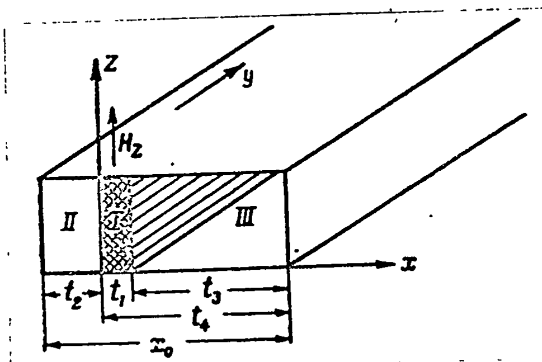

Fig.4

wave propagated in the waveguide also changes.

Thus, the conducting waveguide planes do not simply distort the phenomena taking place in an unlimited ferritic medium, but produce entirely new phenomena which permit designing waveguide systems which do not satisfy the reciprocity principle.

Let us consider waveguide systems which make use of the indicated magneto-optical effects in rectangular waveguides.

Waveguide Systems Based on Nonreciprocal Phase Shifts

As indicated above, the phase velocity of a wave propagated in a rectangular waveguide, which contains a magnetized ferrite (Fig.4), depends on the direction of propagation. This means that the section of the rectangular waveguide containing the ferrite is a nonreciprocal phase shifter and may serve as the basic element of waveguide valve systems.

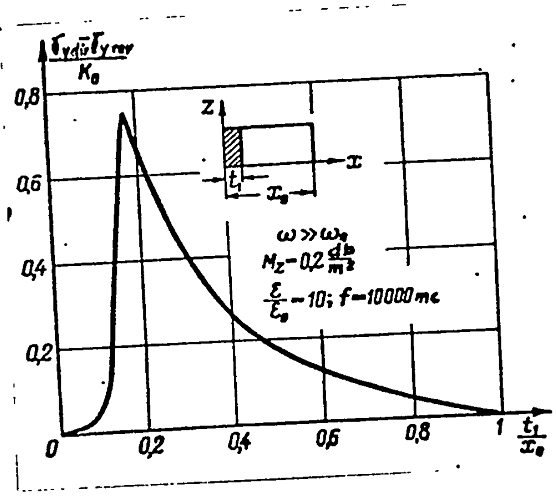

Fig.5

Therefore, the determination of the conditions, under which the ferritic element in the rectangular waveguide produces maximum nonreciprocal phase shifts, is of interest. It follows from eq.(10) that, if the waveguide is completely filled with the ferrite or if the ferritic plate is placed symmetrically in the center of the waveguide, no nonreciprocal phase shifts take place. An investigation of eq.(10) indicates that the use of ferrites is advantageous in relatively weak fields where $\omega \gg \omega_0$.

Here, ω is the frequency of the propagating oscillations and ω_0 the frequency at longitudinal gyromagnetic resonance.

The curve in Fig.5 shows the dependence of the difference between phases of the direct and reverse H_{10} type wave (nonreciprocal phase shift) on the extent to which ferrite fills the waveguide; it follows from the diagram that the maximum ISTAT difference is obtained in a sector whose width comprises 17% of the waveguide size. Such an optimum degree of filling depends on the parameters of the ferrite and the frequency. Figure 6 shows the difference between the phases of the direct and reverse waves, as a function of the position of the ferritic plate in the waveguide.

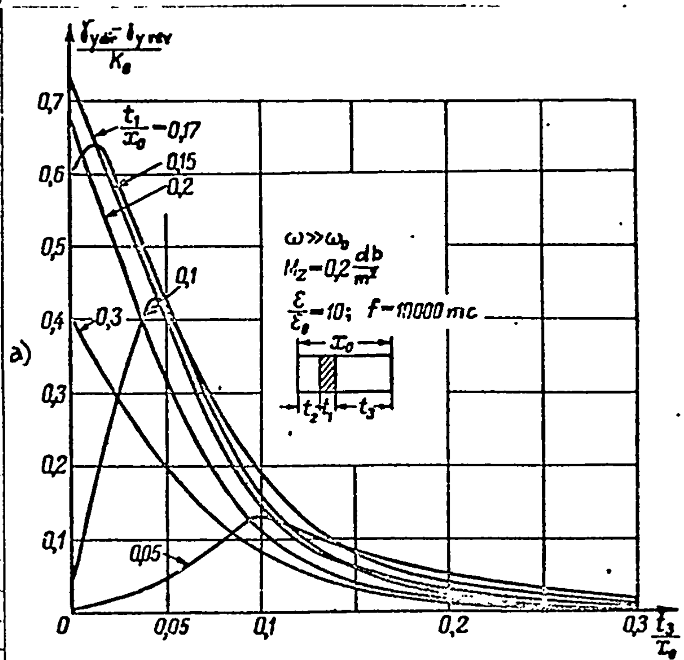

Fig.6

a) Nonreciprocal phase shift in relative units

It follows from consideration of these curves that, first of all, for every width of the ferritic plate, there is an optimum position of the plate in the waveguide $\left(\frac{t_3}{x_0}\right)_{opt}$, at which the difference between the phases of the direct and reverse waves reaches its maximum value; secondly, that, starting from a certain value of the plate width (in our case, equal to 0.17), the phase difference of the direct and reverse waves begins to decrease with increasing plate width.

The curves in Figs.7 and 8 illustrate the effect of the ferrite permittivity (ε) and magnetization (M_z) on the phase difference of the direct and reverse wave on the optimum position of the plate in the waveguide.

The waveguide-valve system, based on nonreciprocal phase shifts, is shown in Fig.9. The system contains two directive couplers or hybrid junctions between which is located the ferritic element which maintains the phase shift specified in Fig.9.

Any coupler with a transient damping of 3 db can be used as a directive coupler. The most compact coupler is the hybrid junction with a wide slot in the narrow waveguide wall, described by Riblet (Bibl.8).

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If a wave of type H_{10} is propagated in the waveguide (3), the wave energy will be divided equally between the two opposite arms of the directive junction, the waves

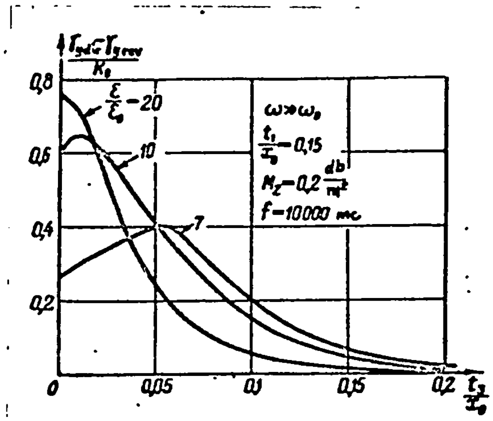

Fig.7

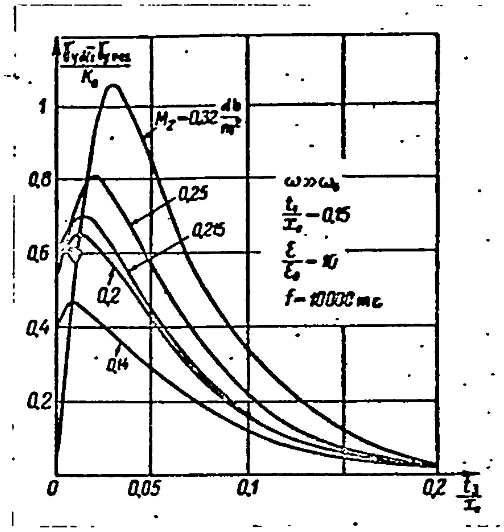

Fig.8

in the arms being 90° out of phase. If the two directive couplers are connected in series, a system is obtained wherein the waves pass from channel (3) to channel (1) and from (1) to (3). If a waveguide section with a ferrite, which gives a nonreciprocal 180° phase shift to the waves propagating in opposite directions, is connected between the directive coupler, then the energy will be transmitted from arm (3) to arm (2), and from arm (2) to arm (4), i.e., the system will be a circulator. If the direction of the external magnetic field is shifted, corresponding to a change in the sign of the phase difference of the direct and reverse waves, energy may be transmitted successively from channel (3) into channel (1) or (2). By changing the external magnetic field intensity, the power distribution between channels (1) and (2) can be regulated, i.e., the system may be used as an amplitude modulator.

The described system is essentially used as a waveguide valve of high power transmittance.

Such a valve (5), designed for operation at a frequency of 9000 mc, introduced a damping of 0.1 db in the forward direction, and of 30 db in the reverse direction. In the 10-percent frequency band, damping of the reverse wave does not drop below

20 db, the direct wave being attenuated by not more than 0.5 db. The basic difficulty in designing such a valve for radio-relay lines is the development of a directive coupler with a very low reflection coefficient (TWR > .95) in a wide frequency band.

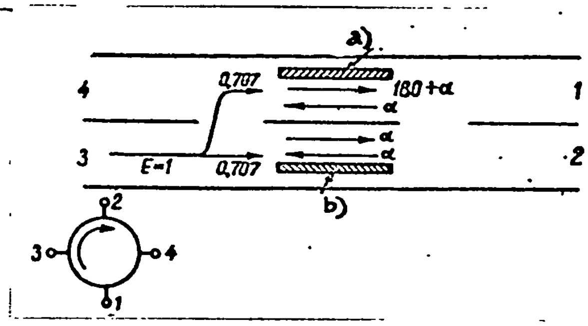

Fig.9

a) Ferrite plate; b) Dielectric plate

Valve Based on Transverse Ferromagnetic Resonance

For a certain permanent magnetic field intensity, determined by eq.(9), a resonant absorption of the wave, propagated through the ferrite, is observed. On the other hand, it follows from eq.(10) that the attenuations of the waves, propagated in opposite directions in the rectangular waveguide with a ferrite, are different. If conditions are selected under which a resonant absorption of the wave, propagated in the forward direction, would be less

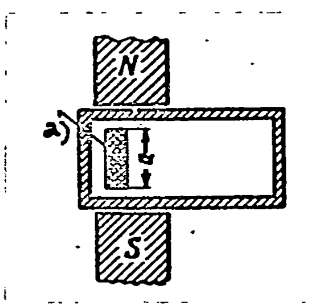

Fig.10

a) Ferrite plate

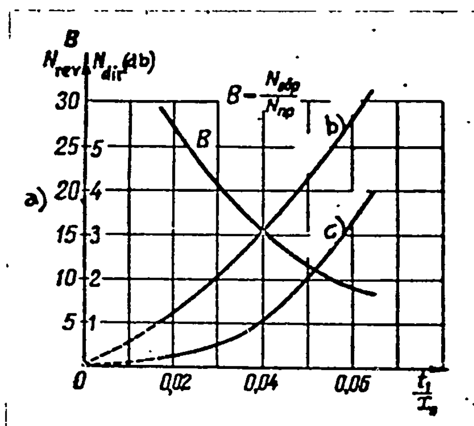

Fig.11

a) Damping in db; b) Reverse; c) Direct

than the resonant absorption of the wave propagated in the reverse direction, then a system could be used as a valve where the return wave would be absorbed by the ferrite itself (Fig.10). Investigation of such a system by solving eq.(10) in the resonant region meets with considerable difficulties and all authors made such investi gations by experimental means (Bibl.9,5,10).

The basic goal of the investigations consisted in determining the effect of such factors as the size and waveguide position of the ferrites, intensity of the external magnetic field, frequency of the propagating oscillations etc., on the resonant absorption. The following conclusions may be drawn from the obtained results:

1) There exists such an optimum plate position in the waveguide, at which its valve properties V, i.e., the ratio of reverse wave losses to direct wave losses, reaches a maximum. Then, the thinner the plate, the closer to the wall will be its optimum position.

2) The curves in Fig.11 show that the valve properties of the system increase with decreasing width of the plate.

3) The valve properties of the system improve with decreasing height a of the plate, to a certain definite magnitude which is independent of the width of the ferrite and is equal to approximately 0.6 - 0.7 of the narrow wall height of the waveguide. Since the decrease in height a of the plate decreases the bias currents in the ferrite, the improvement in the valve properties is apparently associated with a decrease in the ferrite dielectric losses.

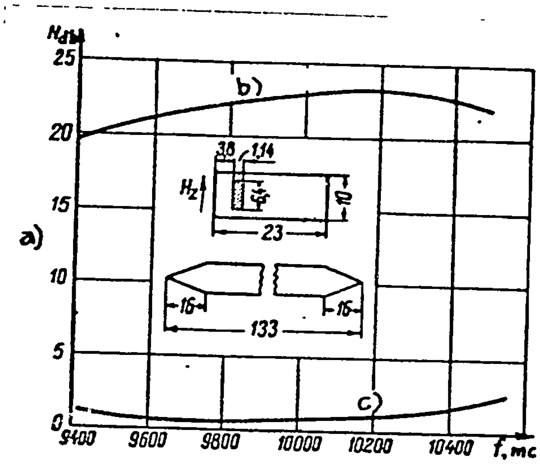

Fig.12

a) Damping, in db; b) Reverse; c) Direct

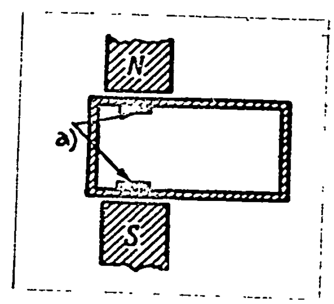

Fig.13

a) Ferrite plate

4) The valve properties of the system do not depend on the length of the fe

plate.

Figure 12 shows the frequency characteristic of the resonant valve (5), which is designed to operate in the 3-cm wave range.

In another paper (Bibl.10), a somewhat modified resonant valve circuit is considered (Fig.13), wherein ferrites in the form of thin plates are fastened to the wide walls of the waveguide. Such a position of the ferrites permits a considerable decrease in the bias currents, i.e., a decrease in dielectric losses.

The authors point out that such a valve, although not a resonant valve, possesses good band properties. At a frequency of 10,000 mc, a valve was designed with a ratio of reverse losses to direct losses not worse than 20:1 in the 12% band. No theoretical investigation of such a system was ever made.

A considerably more detailed investigation of resonant waveguide valves containing ferrites attached to the wide walls of the waveguide, was published (Bibl.17). The curve in Fig.14 shows the dependence of the direct losses and of the reverse losses on the permanent field intensity in such a valve. For comparison purposes, Fig.14 shows an analogous loss curve for a valve design on the old circuit. A comparison of the curves shows that the valve with ferrites attached to the wide walls of the waveguide has a considerably better ratio of reverse losses to direct losses (75:1 versus 15:1). In addition, this valve is more convenient in design and, other conditions being equal, can transmit a higher power.

However, the disadvantages of the valve include a low attenuation per unit length of the reverse wave ($2.56 \frac{db}{cm}$ versus $11.3 \frac{db}{cm}$) and a higher permanent field intensity (4800 oersteds versus 3000 oersteds), which is due to the considerable demagnetizing factors of the ferritic plate.

Further improvement in the characteristics of the resonant valve with ferrites attached to the wide walls of the waveguide is accomplished by placing dielectric plates with a sufficiently high permittivity next to the ferrite (Fig.15).

The dielectric plates concentrate the wave energy in the region where the ferrite is installed, as a result of which the attenuation per unit length of the reverse wave increases up to 9 $\frac{db}{cm}$. The fact that the attenuation of the direct wave did not increase and that, as a result, the ratio of the reverse losses to the direct losses increased up to 150:1 is of interest.

The frequency characteristic of a valve designed on the basis of such a circuit is shown in Fig.15.

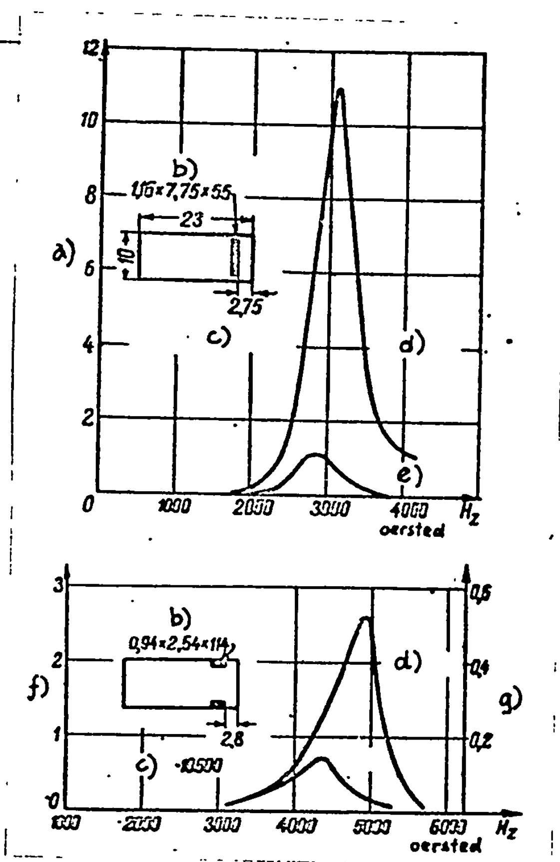

Fig.14

a) Damping, in db/cm; b) Ferrite; c) Frequency: 10,500 mc; d) Reverse; e) Direct; f)Reverse losses, in db/cm; g) Direct losses, in db/cm

Systems Making Use of Nonreciprocal Field "Distortions" in the Waveguide

The electromagnetic field in a waveguide containing a magnetized ferritic plate, is determined from eqs.(11) to (13). The expressions readily show that the field structure of the waves, propagated in the forward direction (along the positive direction of the y-axis), and that of those propagated in the reverse direction, differ. Theoretical investigations of the field structure of electromagnetic waves in a waveguide containing one or two magnetized ferrite plates, were carried out by various authors (Bibl. 11, 12, 13, 14). Figure 16 gives design curves which show the positions of the electric component E_z and of the magnetic component h_y of the field in the cross-section of a waveguide containing one magnetized ferrite plate.

For comparison purposes, on the same drawings are plotted curves which characterize the field structure in an empty waveguide. The change in the electric component E_z of the field, in the cross section of a waveguide containing two ferrite plates, is shown in Fig.17. A study of the curves readily shows that the waveguide with a ferrite has cross sections in which the electric field intensity and, consequently, the intensity of the energy flux are different for the direct and for the reverse waves.

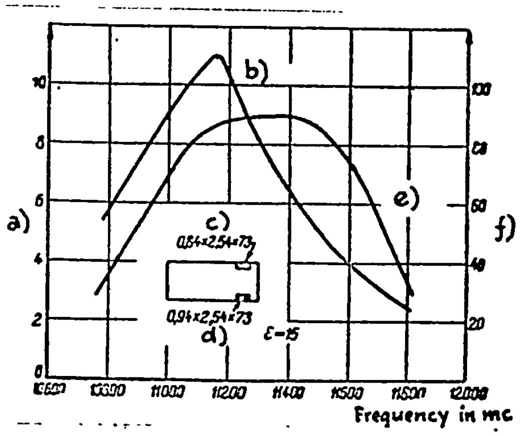

Fig.15

a) Reverse losses, in db/cm; b) Relative; c) Ferrite; d) Dielectric ε=15; e) Reverse losses; f) Relative reverse and direct losses; g) Frequency in mc

If an absorbing plate is placed in the indicated cross section, the damping of the waves propagated in the forward and reverse directions, will differ since such a system will possess valve properties. There are no literature references on an analysis of the conditions under which the valve properties of the system will be maximum.

Other papers (Bibl.5, 15) give characteristics for valves with an absorbing plate designed to operate in the range of 6000, 11,000, and 24,000 mc.

The 6000-mc valve was developed for a waveguide of 15.8×34.84 mm (Fig.18) and en-

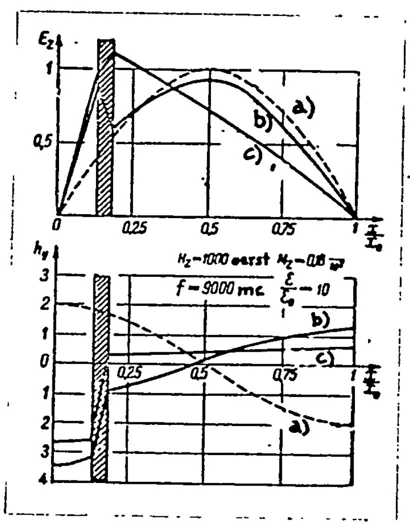

Fig.16

a) Empty Waveguide; b) Direct; c) Reverse

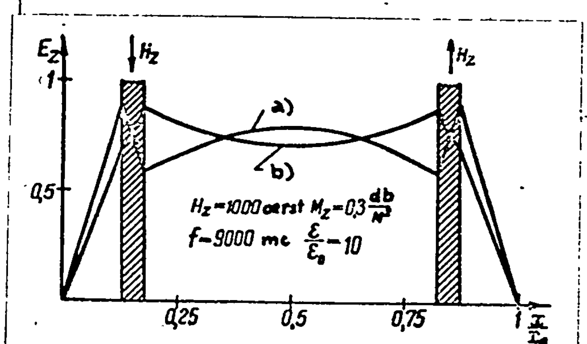

Fig.17

a) Direct; b) Reverse

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sures a ratio of reverse losses to forward losses not worse than 20:1 in the 10% band. The schematic diagram and the characteristics of a valve in the 11,000-mc range are presented in Fig.19.

Valves of such a type are used abroad in radio-relay lines (Bibl.16).

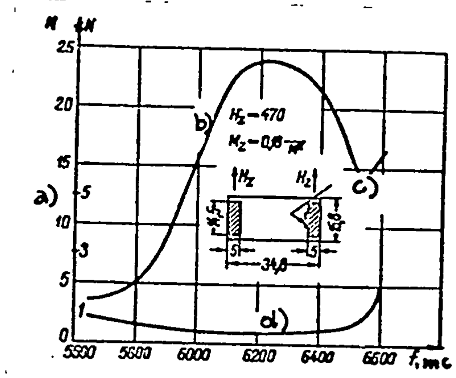

Fig.18

a) Damping, in db; b) Reverse; c) Absorbing plate; d) Direct

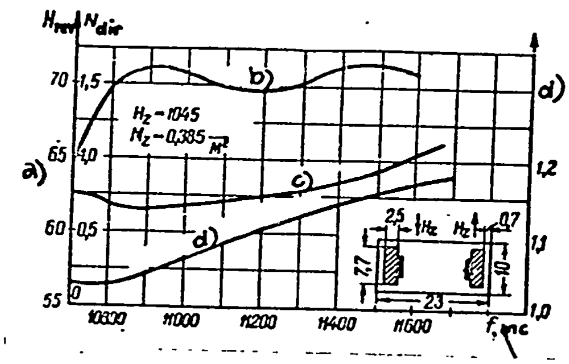

Fig.19

a) Damping, in db; b) Reverse; c) Direct; d) Standing wave ratio

On the basis of utilization of the nonreciprocal field "distortion" in a waveguide containing a ferrite, waveguide circulators, switches, modulators, and other systems, may be designed as well. Due to the shortness of this survey, let us consider the operating principle of one of the possible circulator circuits only (Fig.20).

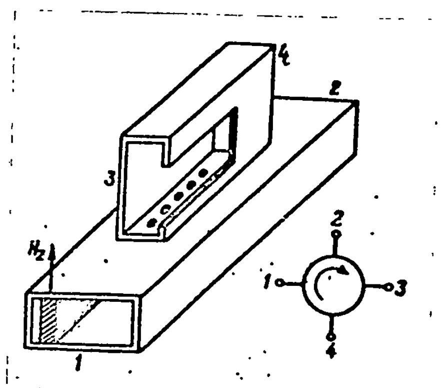

Fig.20

It is known that if a slot is cut in the wide wall of the waveguide, along the line where the longitudinal component h_y of the magnetic field is equal to zero, such a slot will not emit.

The slot in the waveguide of Fig.20 corresponds to the node of the longitudinal component h_y of the magnetic field for one of the directions of wave propagation; therefore, the slot does not emit and damping is negligible in such a direction.

Since the wave is propagated in the reverse direction, the longitudinal component h_y of the magnetic field is quite large near the slot (Fig.16), which results in an energy emission from the waveguide 1-2 into the waveguide 3-4. If the dimensions of the slot (or of the system of slots) are so selected as to make the transmission of energy from (2) to (3) and from (4) to (1) complete, then the system becomes a circulator. The direction of the commutation channels is indicated in Fig.20.

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DETERMINATION OF STEADY-STATE ERROR IN PULSED SYSTEMS WITH LINEAR SYNCHRONISM

by

L.N.Shchelovanov

A formula is derived for steady-state error arising with a divergence in the respective frequencies of the phasing and phased generators. An example is given for calculation of the steady-state error, and ways of reducing it are shown.

Introduction

Systems designed to effect phase correction of generator oscillations find widespread use in synchronous telegraph equipment. Similar synchronization systems can also be employed in facsimile and television equipment.

In response to oscillations of the prongs of the tuning fork at the output of a wave-shaping circuit FK, sawtooth pulses are formed, which are impressed across the

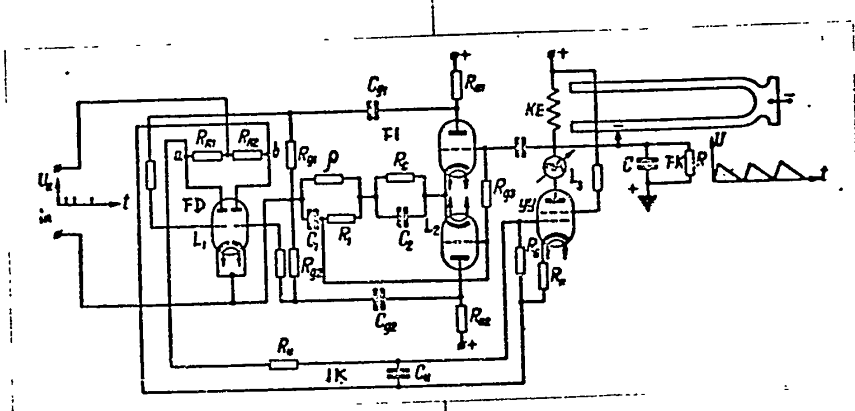

Fig.1

input of the phase inverter FI, and then, in opposition, the sawtooth pulses arrive at the control grids of the phase discriminator FD. Pulses taken from the line are applied to the plates of the phase discriminator. It will be seen from the circ arrangement of the phase discriminator that, at the instant of arrival of the pulses across the load resistors R_{A1},R_{A2}, currents will be flowing in opposite directions, while the current flowing through a corresponding resistor will be dependent on the value of the voltage applied across the control grid. The output voltage is taken off the points a--b of the phase discriminator. The output voltage is equal to zero when the received pulses coincide with the centers of the sawtooth oscillations (Fig.2); when the differentiated pulses are shifted to the left, positive narrow pulses are obtained at the output (Fig.3); when the differentiated pulses are shifted to the right, the output pulses change in sign (Fig.4) (it is assumed that the integrating circuit is inoperative).

The duration of the output pulses remains constant and equal to the duration of the differentiated pulses, while the amplitude varies in proportion to the change in the phase-shift angle. Pulses from the phase discriminator output are fed to the integrating circuit and from here to the amplifier-tube input, which has a compensating electromagnet inserted in its plate circuit to control the tuning-fork frequency. On the basis of its operating principle, the synchronization system under consideration is classified with pulse control systems of type I (Bibl.1).

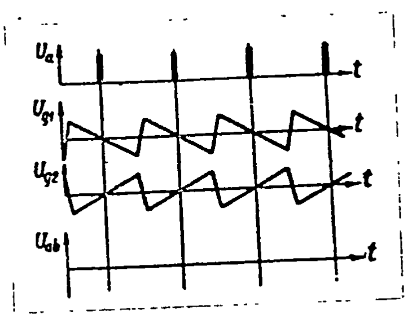

Fig.2

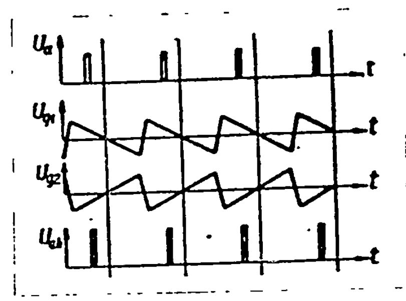

Fig.3

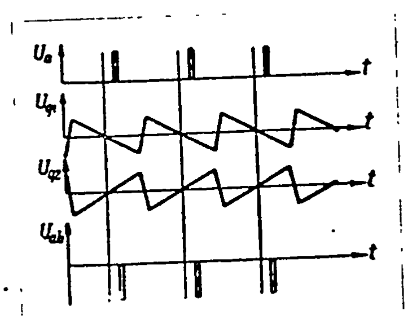

Fig.4

The question of the quality of control systems is a very important one. It must be kept in mind that the margin of synchronous telegraph equipment depends on the accuracy of phasing. One of the criteria governing the quality of the operation of phasing systems is the magnitude of the steady-state error for a disturbance characteristic of these systems - changes in the frequencies of the oscillations of the phasing and phased generators.

The term steady-state error, in automatic control theory, is taken to mean the error occuring under steady-state conditions in response to a characteristic effect exerted on the system. Numerically, the steady-state error is equal to the difference between the internal characteristic effect and the output variable after termination of the transient process, i.e., the constant component of the input variable. If for any reason the frequency of the tuning-fork oscillations diminishes, positive pulses will be formed, with the control circuit closed, at the output of the FD measuring device; these pulses will be smoothed by the integrating circuit, and a positive voltage will appear across the input of the control device. This will lead to an increase in current flowing in the compensating electromagnet, and the difference in frequencies will be eliminated. But the formation of positive pulses at the output of the phase discriminator can take place only because of the constant shift in phase of the oscillations associated with the object of control (the tuning fork) and the pulses taken off the line. The phase shift angle formed in response to the divergence of frequencies of the phasing and phased oscillators is itself the steady-state error ε_∞.

Let us attempt to clarify, with the aid of oscillograms taken on a working model of a linear synchronism system (Fig.5), how steady-state errors are formed in response to changes in the frequency of tuning-fork oscillations. On the oscillograms shown, curve 1 is the voltage change at the output of the phase discriminator; curves 2 and 3 are the voltages applied from the tuning-fork to the control grids of the phase discriminator (Fig.1); curve 4 represents the differentiated pulses formed

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of the oscillations taken from the opposite station. The upper oscillogram, Fig.5a, was taken in the absence of any frequency deviation. After the frequency of the fork was detuned, the differentiated pulses were displaced from their original position (Fig.5b). The difference between the newly assumed position of the differentiated pulses and their position when there was no frequency deviation is equal to the steady-state error ε_∞.

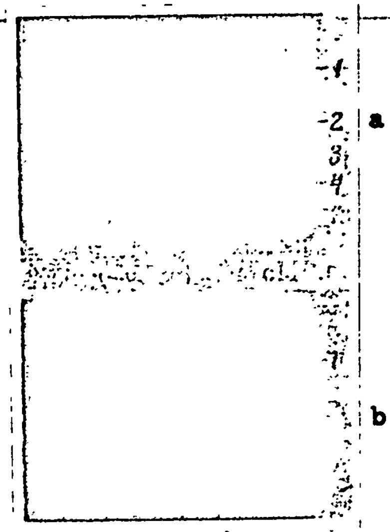

Fig.5

In order to determine the steady-state error, we will define the input variable of a closed control circuit subject to a linear perturbation. The equation of a pulse control system (Bibl.1) takes the form

$$\theta^*_{in}(q) = \theta^*_0(q)\,\frac{1}{1+W^*_{\sim}(q)}, \qquad (1)$$

where θ^*_0 (q) is the transform of the external disturbance;

$W^*_{\sim}$ (q) is the transfer function of an open circuit of a linear synchronism pulse control system (Bibl.3).

$$W^*_{\sim}(q) = ak'_0\beta_0\,\frac{e^q}{(e^q-b)(e^q-1)}. \qquad (2)$$

Here a, k'_0, β_0, b represent constant coefficients depending on the circuit parameters.

In response to a change in frequency, the phase will vary linearly, i.e., the perturbing action will be expressed by a linear function

$$\theta_0[n] = \Delta\varphi n, \qquad (3)$$

where $\Delta\varphi$ represents the phase increment over one period, due to the change in frequency.

Applying a discrete Laplace transformation to eq.(3), we obtain

$$\theta_0^*(q) = \Delta\varphi \frac{e^q}{(e^q - 1)^2}. \quad (4)$$

Substituting eqs.(2) and (4) in eq.(1), we find

$$\theta_{in}^*(q) = \Delta\varphi \frac{e^q}{(e^q - 1)^2} \frac{1}{1 + ak_0'\beta_0 \frac{e^q}{(e^q - b)(e^q - 1)}}. \quad (5)$$

After a series of algebraic transformations, eq.(5) assumes the form

$$\theta_{in}^*(q) = \Delta\varphi \frac{e^q(e^q - b)}{[e^q - (g + \rho)][e^q - (g - \rho)](e^q - 1)}, \quad (6)$$

where

$$g = \frac{1 + b - ak_0'\beta_0}{2}; \quad \rho^2 = g^2 - b. \quad (7)$$

Proceeding from the transform to the original and eliminating the direct component, we find the expression for the steady-state error ε_∞. The original function $\theta_{in}[n]$ may be found simplest by using the inverse formula for a discrete Laplace transform (Bibl.1)

$$\varepsilon_\infty = \Delta\varphi \frac{1 - b}{(1 - g)^2 - \rho^2}.$$

Taking eq.(7) into account, let us evaluate the denominator ε_∞

$$(1 - g)^2 - \rho^2 = ak_0'\beta_0. \quad (8)$$

Then, we have

$$\varepsilon_\infty = \Delta\varphi \frac{1 - b}{ak_0'\beta_0}, \quad (9)$$

where

$$b = e^{-\beta_1}; \quad ak_0'\beta_0 \approx k_0\beta_0\tau\beta_1.$$

Substituting these last equations into eq.(9), we get

$$\varepsilon_{\infty} = \Delta\varphi \frac{1 - e^{-\beta_2}}{k_0 \beta_0 \gamma \beta_1}. \tag{10}$$

Since β_2, for all practical purposes, is considerably less than unity ($\beta_2 \ll 1$), then $1 - e^{-\beta_2} \approx \beta_2$, and the formula for steady-state error will be

$$\varepsilon_{\infty} = \frac{\Delta\varphi \beta_2}{k_0 \beta_0 \gamma \beta_1} \tag{11}$$

or

$$\varepsilon_{\infty} = \frac{\Delta\varphi T_1}{k_0 \beta_0 \gamma T_2}. \tag{12}$$

Calculation Sample for Steady-State Error

We may determine the absolute value of the steady-state error in radians, using eq.(12).

First we will find the value for the individual coefficients.

$\Delta\varphi$ is the phase increment over one period, due to change in frequency. It is clear from Fig.6 that $\Delta\varphi = \frac{2\pi\Delta T}{T_0}$ or, expressing ΔT and T_0 in terms of frequency, we get

$$\Delta\varphi = \frac{2\pi\Delta f}{f_1} \approx \frac{2\pi\Delta f}{f_0}. \tag{13}$$

Let us now assume that, due to a change in the supply voltage, a relative change has taken place in the frequency of the tuning-fork oscillations, amounting to 0.2%, i.e.,

$$\frac{\Delta f}{f_0} 100 = 0{,}2\%.$$

Accordingly, $\Delta\varphi = 2\pi \cdot 0.2 \times 10^{-2} = 1.25 \times 10^{-2}$ rad.

The expressions for the remaining coefficients included in eq.(12) were derived previously (Bibl.2 and 3):

$$T_1 = C_u \frac{R_u (R_i + R_A) + 2R_A R_i}{R_i + R_A}, \tag{14}$$

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$$T_2 = C_u(2R_A + R_u). \qquad (15)$$

Further, $k_o = \frac{2\mu R_A}{R_i + R_A} k_2 k_4 k_5$ is the net gain coefficient of the system;

$k_2 = \frac{R_i S}{R_i + Re}$ is the gain of the regulating device;

where S is the transconductance of the tube characteristic;

R_e is the resistance of the compensating electromagnet.

$k_4 = \frac{U_k T}{2\pi RC}$ is the slope of sawtooth oscillations at the FK output.

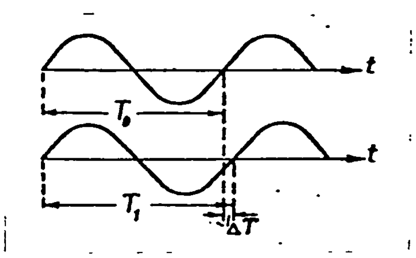

Fig.6

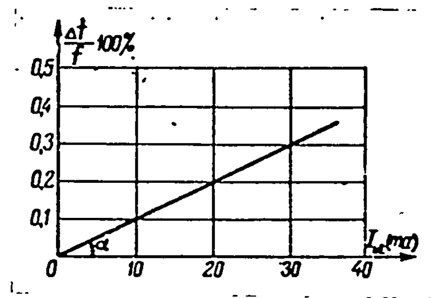

Fig.7

Here U_k denotes the voltage applied by the tuning fork to the wave-shaping circuit; T refers to the period of oscillation of the tuning fork; R and C are the parameters of the wave-shaping circuit.

$$k_5 = \frac{1}{2}\frac{\mu R_{ag}}{R_i + R_{ag}}$$ is the phase inverter gain,

where

$$R_{ag} = \frac{R_a R_g}{R_a + R_g} \text{ (see Fig. 1).}$$

$$\beta_0 = \frac{T}{T_0}, \text{ where } T_0 = \frac{1}{N_\infty}, \quad N = \operatorname{tg} \alpha.$$

We find N from the graph showing the dependence of the relative change in tuning-fork frequency on current fluctuations in the KE (Fig.7).

Let the circuit parameters (Fig.1) be such that $T_1 = 63$ msec, $T_2 = 150$ msec,

$k_o = 0.73$, $\beta_o = 0.628$, $\gamma = 0.1$. Then, the steady-state error will be equal to

$$\varepsilon_\infty = \frac{1,25\cdot 10^{-2}\cdot 63}{0,73\cdot 0,628\cdot 0,1\cdot 150} = 0,114 \ rad.$$

From the time curves (Fig.2-4), we can see that the magnitude of the phase shift of the differentiated pulses relative to the in-phase position is limited by the duration of the leading edge of the sawtooth oscillations. If the duration of the leading edge is denoted in terms of radians as θ_n, then the allowable phase shift for the differentiated pulses relative to their in-phase position will be equal to $\pm \frac{\theta_n}{2}$. It is appropriate to evaluate the value of the steady-state error relative to a number equal to the ratio of the absolute value of the steady-state error ε_∞ and the allowable phase shift of the differentiated pulses

$$\varepsilon_{rel} = \frac{2\varepsilon_\infty}{\theta_n}, \qquad (16)$$

or, percentagewise,

$$\varepsilon_{rel}\ \% = \frac{2\varepsilon_\infty}{\theta_n} 100\%. \qquad (17)$$

In the example considered, we have

$$\varepsilon_{rel}\ \% = \frac{2\cdot 0,114}{0,8\cdot 2\pi} 100 = 5,6\%.$$

In conclusion, let us analyze the effect of the integrating circuit parameters (C_u and R_u) on the magnitude of the steady-state error. To this end, let us consider the variation in the ratio $\frac{T_1}{T_2}$ in eq.(12) in response to changes in C_u and R_u.

Since C_u enters eqs.(14) and (15) with a common multiplying factor, the ratio $\frac{T_1}{T_2}$ will not depend on C_u and, accordingly, the magnitude of the steady-state error will not depend on the capacitance C_u of the integrating circuit.

In order to clarify the effect of R_u on the magnitude of the steady-state error, let us find the ratio $\frac{T_1}{T_2}$

$$\frac{T_1}{T_2} = \frac{R_u(R_i + R_A) + 2R_A R_i}{(R_i + R_A)(2R_A + R_u)} = \frac{R_u + \frac{2R_A R_i}{R_i + R_A}}{R_u + 2R_A}. \qquad (18)$$

In eq.(18), $\frac{2R_A R_i}{R_i + R_A} < 2R_A$.

Therefore, eq.(18) represents a proper fraction. The fraction (18) tends toward unity as R increases, while the ratio $\frac{T_1}{T_2}$ decreases in direct proportion to R_u. Consequently, an increase in the resistance R presented by the integrating circuit leads to an increase in the value of the steady-state error and, conversely, a decrease in the steady-state error results from a decrease in R_u.

Accordingly, the steady-state error ε_∞ can be reduced by: 1) enhancing the stability of the phasing and phased generators [eq.(12)] ; 2) enlarging the net gain; in that case, it is necessary its value must not exceed the cutoff gain coefficient (Bibl.3); 3) decreasing the resistance associated with the integrating circuit.

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NETWORK ANALYSIS OF DC TELEGRAPH APPARATUS

by

Kh.I.Cherne

The laws of current change affecting the currents flowing through the electromagnets of telegraph equipment connected to short aerial lines, using a DC circuit, are determined.

Introduction

The transients arising during operation of telegraph equipment through uniform lines have already been treated in the literature (Bibl.1,2,3).

In the present paper, the laws of current change are determined in relation to the currents flowing through the electromagnets of telegraph apparatus hooked up to a DC circuit (Fig.1) through a short aerial uniform line, both in the operation of the transmitter at a station, using a battery, and in the operation of a station without battery. The cases of opening and closing of the key are considered, taking into account non-zero initial conditions.

The complexity involved in the solution of transient problems in the operation

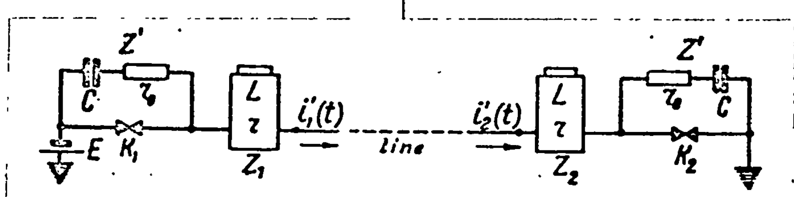

Fig.1

of telegraph equipment through uniform lines requires a few assumptions.

All the circuit elements in Fig.1, shown in this paper, are assumed as linear. In addition, it is presumed that the effect of line inductance and line capacitance on the transient processes may be safely neglected and that the remaining two primary parameters are constant in time and independent of the frequency, within the frequency range in use for DC telegraph working. The presence of spark-quenching circuits (Fig.1) gives us reason to assume that no arcing takes place between the key contacts during operation of the transmitters.

Derivation of the General Parameter Relations

Let us now determine the currents $\dot{I}_1'$ and $\dot{I}_2'$, which will flow through the input and output terminals of the linear passive quadrupole network (Fig.2a) when the contacts k open, taking as known quantities the parameter-coefficients A_{11}, A_{12}, A_{21}, A_{22}, the load impedances Z_1 and Z_2, the emf E, and the impedance Z' by which the load impedance Z_1 is increased, when the contacts k open.

Before opening of the contacts referred to, the currents $\dot{I}_1$ and $\dot{I}_2$ whose values can be determined on the basis of known formulas (Bibl.4, p.223) flow through the loads Z_1 and Z_2 of the four-terminal network:

$$\dot{I}_1=\frac{\dot{E}\,(A_{21}Z_2+A_{22})}{H}, \qquad (1)$$

$$\dot{I}_2=\frac{\dot{E}}{H}, \qquad (2)$$

where

$$H=A_{11}Z_2+A_{12}+A_{21}Z_1Z_2+A_{22}Z_1. \qquad (3)$$

After opening of the contacts k, the impedance Z' is inserted and the currents $\dot{I}_1$ and $\dot{I}_2$ are given the increments $\Delta\dot{I}_1$ and $\Delta\dot{I}_2$, respectively. We will use the compensation theorem for determining these increments.

According to this theorem, the indicated current increments $\Delta\dot{I}_1$ and $\Delta\dot{I}_2$ are equal to the currents which would be generated in the circuit by a source with a driving voltage $\dot{E}'$ equal to $\dot{I}_1 Z'$ provided it was connected in series with the impedance Z', opposing the current $\dot{I}_1$, as shown in Fig.2b.

A comparison of Figs.2b and 2a shows that, in order to determine the current increments $\Delta\dot{I}_1$ and $\Delta\dot{I}_2$, we can use eqs.(1) and (2), in which, for this purpose, the

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quantities $\dot{I}_1$ and $\dot{I}_2$, $\dot{E}$ and H, must be substituted by $\Delta\dot{I}_1$, $\Delta\dot{I}_2$, $\dot{E}'$ and H', respectively. We obtain the expression for H' from eq.(3) by substituting the value of $Z_1 + Z'$ for Z_1. After these substitutions, we obtain:

$$\Delta \dot{I}_1 = \frac{\dot{I}_1 Z' (A_{21} Z_2 + A_{22})}{H'}, \tag{4}$$

$$\Delta \dot{I}_2 = \frac{\dot{I}_1 Z'}{H'}, \tag{5}$$

where

$$H' = A_{11} Z_2 + A_{12} + A_{21} (Z_1 + Z') Z_2 + A_{22} (Z_1 + Z'). \tag{6}$$

With $\Delta\dot{I}_1$ and $\Delta\dot{I}_2$ known, we can determine the unknown currents $\dot{I}'_1$, and $\dot{I}'_2$:

$$\dot{I}'_1 = \dot{I}_1 - \Delta \dot{I}_1, \tag{7}$$

$$\dot{I}'_2 = \dot{I}_2 - \Delta \dot{I}_2. \tag{8}$$

Operation of Transmitter of a Station with Battery

Opening of Key k_1. Before opening of the key k_1 in the circuit shown in Fig.1, direct currents, equal to I_{o1} at the circuit input and I_{o2} at the output, are flowing.

A comparison of Figs.1 and 2a shows that these currents can be determined on the basis of eqs.(1) and (2). It is only necessary to keep in mind here that, when direct current is used, $Z_1 = Z_2 = r$ and that the parameter-coefficients of a uniform line, under the assumptions made, are equal to

$$A_{11} = A_{22} = \operatorname{ch} \beta_0 l, \quad A_{12} = \rho_0 \operatorname{sh} \beta_0 l, \quad A_{21} = \frac{\operatorname{sh} \beta_0 l}{\rho_0}, \tag{9}$$

where

$$\rho_0 = \sqrt{\frac{R}{G}}, \quad \beta_0 = \sqrt{RG}, \tag{10}$$

Here, R and G are the primary line parameters expressed in ohm/km and mho/km, respectively, while a l is the length of the line in kilometers.

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Substituting these values into eqs.(1) and (2), we obtain

$$I_{01}=\frac{Em_0}{H_0}, \qquad (11)$$

$$m_0=\frac{\rho_0 \operatorname{ch}\beta_0 l + r \operatorname{sh}\beta_0 l}{\rho_0}, \qquad (12)$$

where

$$I_{02}=\frac{E}{H_0}, \qquad (13)$$

$$H_0=2r \operatorname{ch}\beta_0 l+\left(\rho_0+\frac{r^2}{\rho_0}\right)\operatorname{sh}\beta_0 l. \qquad (14)$$

When the key k_1 opens (Fig.3a), the load impedance at the west end of the line is increased by Z', which gives rise to a transient response. We will determine the transforms of the current increments through each electromagnet, brought about by the inclusion of the impedance Z', according to the compensation theorem (Fig.3b) on the basis of eqs.(4)-(6) on substitution of all the values of the transforms entering into the formulas.

Since, in the case under consideration, the transforms of the line parameters

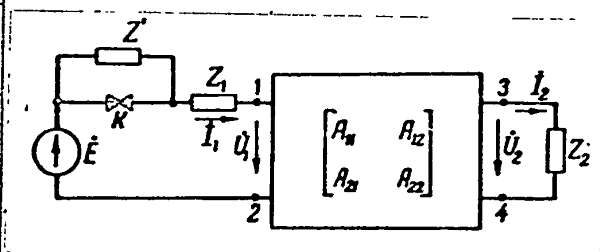

Fig.2 a)

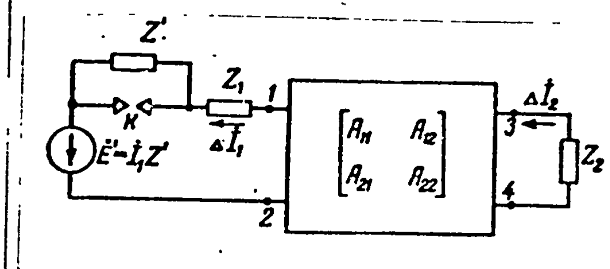

Fig.2 b)

coincide with the parameters themselves, whose expressions are given in eqs.(9) and (10), and since

$$I_1(p)=\frac{I_{01}}{p}, \qquad (15)$$

$$Z'(p)=\frac{1+pCr_0}{pC}, \qquad (16)$$

$$Z_1(p)=Z_2(p)=r+pL, \qquad (17)$$

then, after the substitution indicated, we obtain the expressions for the current

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increment transforms in the following form:

$$\Delta i_1(p) = \frac{I_{01} r_0 (p^2 + p b_1 + b_2)}{pL(p - p_1)(p - p_2)(p - p_3)}, \qquad (18)$$

$$\Delta i_2(p) = \frac{I_{01} a_3 (1 + pCr_0)}{pm_0 (p - p_1)(p - p_2)(p - p_3)}. \qquad (19)$$

Here

$$b_1 = \frac{r_0 \rho_0 m_0 C + L \operatorname{sh} \beta_0 l}{LCr_0 \operatorname{sh} \beta_0 l}, \qquad (20)$$

$$b_2 = \frac{\rho_0 m_0}{LCr_0 \operatorname{sh} \beta_0 l}, \qquad (21)$$

p_1, p_2, p_3 are the roots of the equation

$$p^3 + p^2 a_1 + p a_2 + a_3 = 0, \qquad (22)$$

where

$$a_1 = \frac{r_0 + 2(r + \rho_0 \operatorname{cth} \beta_0 l)}{L}, \qquad (23)$$

$$a_2 = \frac{[L + C(\rho_0^2 + r^2 + rr_0)] \operatorname{sh} \beta_0 l + \rho_0 (r_0 + 2r) \operatorname{ch} \beta_0 l}{L^2 C \operatorname{sh} \beta_0 l}, \qquad (24)$$

$$a_3 = \frac{\rho_0 m_0}{L^2 C \operatorname{sh} \beta_0 l}. \qquad (25)$$

With the transforms $\Delta i_1(p)$ and $\Delta i_2(p)$ known, we find the current increments $\Delta i_1(t)$ and $\Delta i_2(t)$.

Applying Heaviside's expansion formula [Bibl.5, p.23, formula(3)] , we get

$$\Delta i_1(t) = I_{01}\left[1 + \frac{r_0}{L}(K_1 e^{p_1 t} + K_2 e^{p_2 t} + K_3 e^{p_3 t})\right], \qquad (26)$$

$$\Delta i_2(t) = \frac{I_{01}}{m_0}\left[1 + a_3 (N_1 e^{p_1 t} + N_2 e^{p_2 t} + N_3 e^{p_3 t})\right]. \qquad (27)$$

$$K_i = \frac{p_i^2 + p_i b_1 + b_2}{q_i}, \quad (i = 1, 2, 3) \qquad (28)$$

$$N_i = \frac{1 + p_i C r_0}{q_i}, \quad (i = 1, 2, 3) \qquad (29)$$

$$q_1 = p_1 (p_1 - p_2)(p_1 - p_3), \qquad (30)$$

$$q_2 = p_2 (p_2 - p_1)(p_2 - p_3), \qquad (31)$$

$$q_3 = p_3 (p_3 - p_1)(p_3 - p_2). \qquad (32)$$

The currents $i_1(t)$ at the line input and $i_2(t)$ at the output, after opening of the key k_1, may be computed in accordance with eqs.(7) and (8) from the expressions:

$$i_1(t) = I_{01} - \Delta i_1(t), \quad (33)$$

$$i_2(t) = I_{02} - \Delta i_2(t). \quad (34)$$

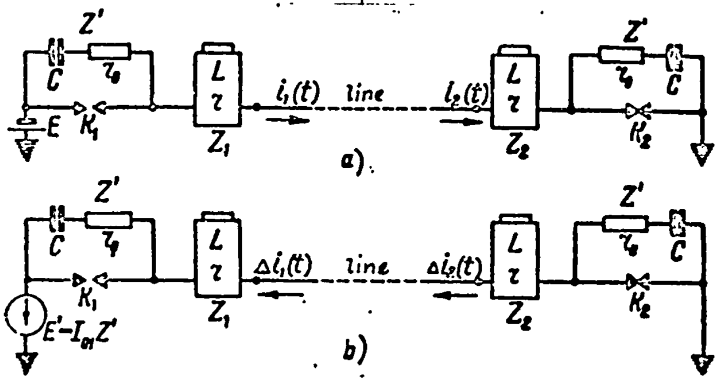

Fig.3

Substituting into these equations the expressions (26) and (27), we obtain the unknown currents in the following form:

$$i_1(t) = -I_{01} \frac{r_n}{L} (K_1 e^{p_1 t} + K_2 e^{p_2 t} + K_3 e^{p_3 t}), \quad (35)$$

$$i_2(t) = -I_{02} a_3 (N_1 e^{p_1 t} + N_2 e^{p_2 t} + N_3 e^{p_3 t}). \quad (36)$$

Depending on the values of the roots of eq.(22), the following three cases are possible:

a) All the roots of eq.(22) are real and there are no multiple roots.

In that case, a nonrecurrent process takes place in the circuit and the currents $i_1(t)$ and $i_2(t)$ are determined on the basis of eqs.(35) and (36). The law of change obeyed by each of these currents as a function of time, is shown in Fig.4a.

b) All the roots of eq.(22) are real, and two of them are multiple roots.

Let the roots p_1, p_2 and p_3 of eq.(22) be real, with $p_2 = p_3$.

After substituting eq.(28) into eq.(35) and eq.(29) into eq.(36) and evaluating the indeterminate forms at $p_3 = p_2$, we obtain, for the following expressions for the currents $i_1(t)$ and $i_2(t)$:

$$i_1(t) = -I_{01}\frac{r_0}{L}\left[K_1' e^{p_1 t} - \frac{e^{p_2 t}}{p_2^2 m_1^2}(m_1 m_2 t + m_3)\right], \quad (37)$$

$$i_2(t) = -I_{02} a_3 \left[N_1' e^{p_1 t} - \frac{e^{p_2 t}}{p_2^2 m_1^2}(m_5 t + m_4)\right], \quad (38)$$

where

$$m_1 = p_1 - p_2, \quad (39)$$

$$m_2 = p_2(p_2^2 + p_2 b_1 + b_2), \quad (40)$$

$$m_3 = m_2 + m_1(p_2^2 - b_2), \quad (41)$$

$$m_4 = m_6 - m_1, \quad (42)$$

$$m_5 = m_1 m_6, \quad (43)$$

$$m_6 = p_2(1 + p_2 C r_0), \quad (44)$$

$$K_1' = \frac{p_1^2 + p_1 b_1 + b_2}{p_1 m_1^2}, \quad (45)$$

$$N_1' = \frac{1 + p_1 C r_0}{p_1 m_1^2}. \quad (46)$$

In that case, a critical nonrecurrent process takes place in the circuit.

The law of current change for each of the currents $i_1(t)$ and $i_2(t)$ is analogous to the one shown in Fig.4a below.

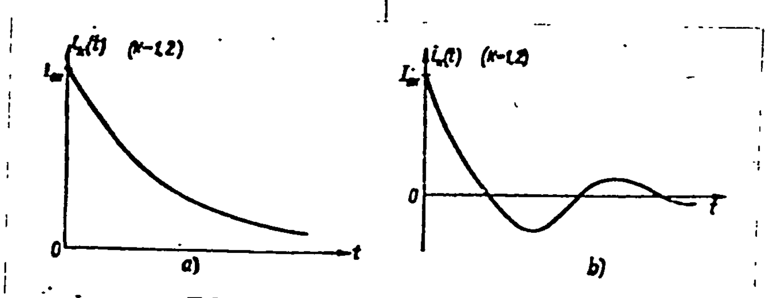

Fig.4

c) One root of eq.(22) is real and two roots are complex conjugates.

Let the root p_1 of eq.(22) be real, and roots p_2 and p_3 be complex conjugates, as follows

$$p_2 = \alpha + i\omega, \quad (47)$$

$$p_3 = \alpha - i\omega. \quad (48)$$

Substituting the values of the roots into eqs.(26)-(32), we get

$$i_1(t) = -I_{01}\frac{r_0}{L}\left[K_1'' e^{p_1 t} - \frac{M_1 e^{\alpha t}}{\omega m_7 m_8}\sin(\omega t + \psi_1)\right], \quad (49)$$

$$i_2(t) = -I_{02}a_3\left[N_1''e^{p_1t} - \frac{M_2e^{\alpha t}}{\omega m_7 m_9}\sin(\omega t+\psi_2)\right], \qquad (50)$$

where

$$M_1 = \frac{m_{10}}{\cos\psi_1}, \qquad (51)$$

$$\mathrm{tg}\,\psi_1 = \frac{\omega m_{11}}{m_{10}}, \qquad (52)$$

$$K_1'' = \frac{p_1^2 + p_1b_1 + b_2}{p_1 m_9}, \qquad (53)$$

$$M_2 = \frac{m_{16}}{\cos\psi_2}, \qquad (54)$$

$$\mathrm{tg}\,\psi_2 = \frac{\omega m_{15}}{m_{16}}, \qquad (55)$$

$$N_1'' = \frac{1 + p_1Cr_0}{p_1 m_9}, \qquad (56)$$

$$m_7 = \alpha^2 + \omega^2, \qquad (57)$$

$$m_8 = p_1 - \alpha, \qquad (58)$$

$$m_9 = m_8^2 + \omega^2, \qquad (59)$$

$$m_{10} = m_8m_{12} - \omega^2 m_{13}, \qquad (60)$$

$$m_{11} = m_{12} + m_8m_{13}, \qquad (61)$$

$$m_{12} = \alpha(m_7 + b_2) + m_7b_1, \qquad (62)$$

$$m_{13} = m_7 - b_2, \qquad (63)$$

$$m_{14} = \alpha + m_7Cr_0, \qquad (64)$$

$$m_{15} = m_{14} - m_8, \qquad (65)$$

$$m_{16} = m_8m_{14} + \omega^2. \qquad (66)$$

In this case, a recurrent process takes place in the circuit. The law of current change governing each of the currents $i_1(t)$ and $i_2(t)$ is shown in Fig.4b.

Closing of Contacts k_1. At the instant when the contacts k_1 close, let the currents $i_1(t)$ and $i_2(t)$, set up by the previous opening of those contacts, diminish to the point where they can be neglected in first approximation. Then, the voltage across the capacitor of the spark-quenching circuit of the west transmitter will be equal to the emf of the battery E.

After the contacts k_1 close (Fig.5), the discharge current of this capacitor will be cut off across the spark-quenching circuit and the currents flowing through the electromagnets at both ends of the transmission line will not depend on the voltage across the capacitor.

Comparing Figs.5 and Fig.2a, we can see that the current transforms $i_1'(t)$ and $i_2'(t)$, due to the closing of the contacts k_1, can be determined on the basis of eqs.(1)-(3), into which the transforms of the values included in the formulas must be substituted.

Since

$$E(p)=\frac{E}{p}, \qquad (67)$$

then, making use of eq.(17) and taking into account the reference made above to the

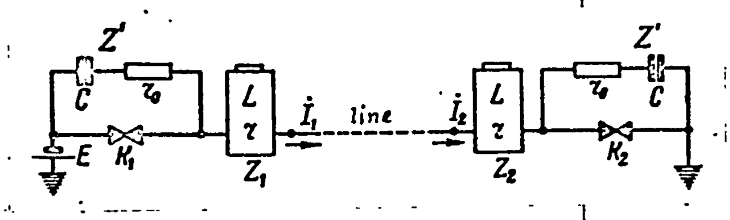

Fig.5

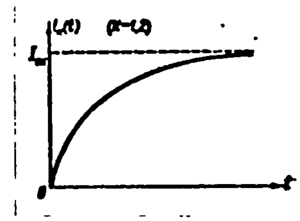

Fig.6

values of the transforms of the uniform transmission-line parameters in the case considered, we obtain the transforms $i_1'(t)$ and $i_2'(t)$ of the currents in the following form:

$$i_1(t)=\frac{E(m_0\beta+p)}{pL(p-p_1')(p-p_2')}, \qquad (68)$$

$$i_2(t)=\frac{E\beta}{pL(p-p_1')(p-p_2')}, \qquad (69)$$

where

$$p_1'=\alpha'+\beta, \qquad (70)$$

$$p_2'=\alpha'-\beta, \qquad (71)$$

$$\alpha'=-m_0\beta, \qquad (72)$$

$$\beta=\frac{\rho_0}{L\,\mathrm{sh}\,\beta_0 l}. \qquad (73)$$

With the transforms $i_1'(p)$ and $i_2'(p)$ known, let us find the currents $i_1'(t)$ and $i_2'(t)$. Applying Heaviside's expansion theorem [Bibl.5, p.23, eq.(3)], we obtain

$$i_1'=I_{01}\left[1-e^{\alpha' t}\left(\mathrm{ch}\,\beta t+\frac{\mathrm{sh}\,\beta t}{m_0}\right)\right], \qquad (74)$$

$$i_2'=I_{02}\left[1-e^{\alpha' t}(\mathrm{ch}\,\beta t+m_0\,\mathrm{sh}\,\beta t)\right]. \qquad (75)$$

The law of current change for each of these currents, as a function of time, is shown in Fig.6.

Operation of Transmitter of a Station Without a Battery

Opening of key k_2. Before opening of the key k_2, DC currents I_{01} and I_{02} whose values are determined above, flow through the electromagnets of the west and east-end equipment.

After the key k_2 opens (Fig.7a), the load impedance at the east end of the transmission line increases by Z'. The current increments through each electromagnet, brought about by insertion of the impedance Z', can be determined from the circuit diagram in Fig.7b, obtained on the basis of the compensation theorem.

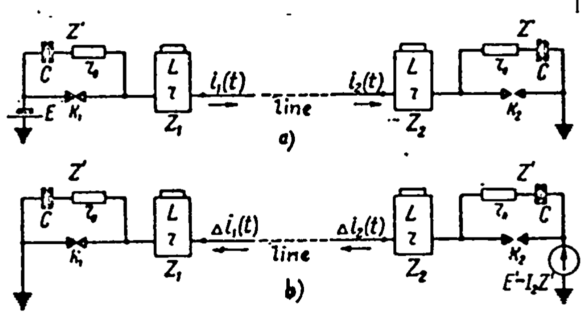

Fig.7

A comparison of the circuit depicted in Fig.7b with that in Fig.4a shows that, in order to determine the current increments occurring in response to the opening of the key k_2, it is possible to use eqs.(26) and (27), obtained for the circuit in Fig.4a, provided I_{01} is replaced by I_{02}, $\Delta i_1(t)$ by $\Delta i_2(t)$, and $\Delta i_2(t)$ by $\Delta i_1(t)$.

Performing the indicated substitution and making use of eqs.(33), (34), (11), and (12), we obtain the following expressions for the currents $i_1(t)$ and $i_2(t)$ in the circuit shown in Fig.7a:

$$i_1(t) = I_{01}\left\{1 - \frac{1}{m_0^2}\left[1 + a_3\left(N_1 e^{p_1 t} + N_2 e^{p_2 t} + N_3 e^{p_3 t}\right)\right]\right\}, \qquad (76)$$

$$i_2(t) = -I_{02}\frac{r}{L}\left(K_1 e^{p_1 t} + K_2 e^{p_2 t} + K_3 e^{p_3 t}\right). \qquad (77)$$

The values of the terms used in these expressions have been given above [eqs. (11)-(13), (22), (25), (28), (29)].

Depending on the values of the roots of eq.(22), the following three cases are possible.

a) All roots of eq.(22) are real and there are no repeated roots.

In this case, the currents $i_1(t)$ and $i_2(t)$ are determined on the basis of eqs.(76) and (77). A nonrecurrent process takes place in the circuit. The current $i_1(t)$ varies as a function of time, as shown in Fig.8a, and the change in the current $i_2(t)$ corresponds to that shown in Fig.4a. At $t \to \infty$, the current $i_1(t)$ tends toward a constant value I_{1xx}, equal to

$$I_{1xx} = \frac{E}{m_{17}}, \qquad (78)$$

where

$$m_{17} = r + \rho_0 \operatorname{cth} \beta_0 l. \qquad (79)$$

This expression is easily obtained from eq.(76) at $t \to \infty$, if use is made of eqs.(11), (13), and (14).

b) All roots of eq.(22) are real, two of them being repeated roots.

Let the roots p_1, p_2, and p_3 of eq.(22) be real, at $p_2 = p_3$.

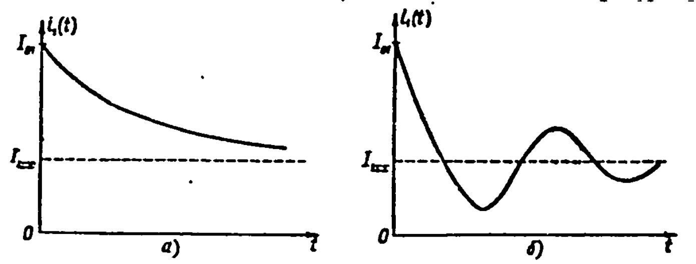

Fig.8

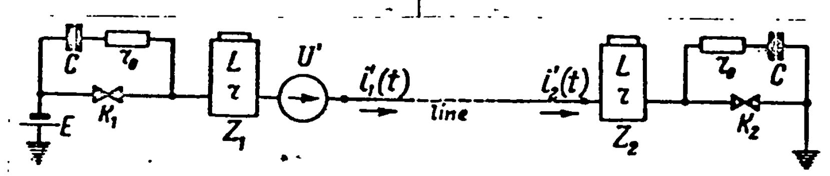

Fig.9

Substituting eq.(29) into eq.(76) and eq.(28) into eq.(77) and evaluating the indeterminate forms at $p_3 = p_2$, we obtain

$$i_1(t) = I_{01}\left[1 - \frac{1}{m_0^2}\left\{1 + a_3\left[N_1' e^{p_1 t} - \frac{e^{p_2 t}}{p_2^2 m_1^2}(m_5 t + m_4)\right]\right\}\right], \qquad (80)$$

$$i_2(t) = -I_{02}\frac{r_0}{L}\left[K_1' e^{p_1 t} - \frac{e^{p_2 t}}{p_2^2 m_1^2}(m_1 m_2 t + m_3)\right]. \qquad (81)$$

The values for the terms used in these equations are given above [in eqs.(11)-(13), (25), (39)-(46)].

A critical nonrecurrent process takes place in the circuit. The law of current change, affecting the current $i_1(t)$, is analogous to the one shown graphically in Fig.8a, and that affecting the current $i_2(t)$ to the one shown in Fig.4a.

c) One root of eq.(22) is real, while the roots p_2 and p_3 are complex conjugates.

Let the roots p_1 of eq.(22) be real, while the roots p_2 and p_3 are complex conjugates [cf. eqs.(47) and (48)]. Substituting the values for the roots into eqs.(76) and (77), we obtain

$$i_1(t)=I_{01}\left[1-\frac{1}{m_0^2}\left\{1+a_3\left[N_1'' e^{p_1 t}-\frac{M_2 e^{\alpha t}}{\omega m_7 m_9}\sin(\omega t+\psi_2)\right]\right\}\right], \qquad (82)$$

$$i_2(t)=-I_{02}\frac{r_0}{L}\left[K_1' e^{p_1 t}-\frac{M_1 e^{\alpha t}}{\omega m_7 m_9}\sin(\omega t+\psi_1)\right]. \qquad (83)$$

The values for the terms used in these expressions are given above [eqs.(11)-(13), (25), (51)-(59)].

A recurrent process takes place in the circuit. The current $i_1(t)$ varies as a function of time, as shown in Fig.8b, while the change in the current $i_2(t)$ corresponds to Fig.4b.

<u>Closing of the key k_2</u>. At the instant that the key contacts k_2 close, the transient process set up by the previous opening of the keys ceased for all practical purposes, so that it may be assumed that, in the circuit shown in Fig.7a, $i_1(t) = I_{1xx}$, while $i_2(t) = 0$.

The effect of the current I_{1xx} on the transient response after closing of the key k_2 may be taken into account by introducing a fictitious source with a voltage equal to

$$U^t = pLI_{1xx}, \qquad (84)$$

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as shown* in Fig.9.

Comparing Figs.9 and 2a, we see that the transforms of the currents $i_1(t)$ and $i_2(t)$ due to closing of the key k_2, can be determined on the basis of eqs.(1)-(3) in which, for this purpose, the transforms of the values included in the formulas must be substituted.

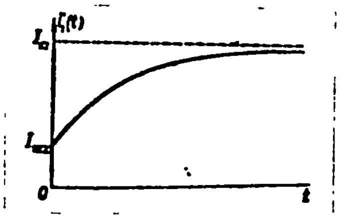

Fig.10

Since

$$E(p)=\frac{E+U'}{p}, \qquad (85)$$

then, using eqs.(17) and (84) and taking into account the reference above as to the values of the transforms of the uniform transmission-line parameters in the case under consideration, we obtain the transforms for the currents $i_1'(t)$ and $i_2'(t)$ in the following form:

$$i_1'(p)=\frac{E(m_{17}+pL)^2}{pL^2m_{17}(p-p_1')(p-p_2')}, \qquad (86)$$

$$i_2'(p)=\frac{E\beta(m_{17}+pL)}{pLm_{17}(p-p_1')(p-p_2')}. \qquad (87)$$

The values for the terms used in these expressions are given above [eqs.(70)-(73) and (79)].

With the transforms $i_1'(p)$ and $i_2'(p)$ known, let us find the currents $i_1'(t)$ and $i_2'(t)$. Using Heaviside's expansion theorem [(Bibl.5, p.23, eq.(3)], we obtain

$$i_1'(t)=I_{01}\left[1-\frac{e^{\alpha' t}}{m_0}\left(\operatorname{sh}\beta t+\frac{\operatorname{ch}\beta t}{m_0}\right)\right], \qquad (88)$$

$$i_2'(t)=I_{02}\left[1-e^{\alpha' t}\left(\frac{\operatorname{sh}\beta t}{m_0}+\operatorname{ch}\beta t\right)\right]. \qquad (89)$$

* We resort to one of the methods of taking the nonzero initial conditions into account, developed by Prof. E.V.Zelyakh in his report "Contribution to the Analysis of Transients in Electric Circuits with Nonzero Initial Conditions", read at a scientific engineering conference of the LEIS on 25 May 1950.

The values for the terms used here are given above [eqs.(11), (12), (13), (72), and (73)].

The current $i_1'(t)$ varies as a function of time, as shown in Fig.10, while the change in the current $i_2'(t)$ corresponds to that shown in Fig.6.

A comparison of the figures showing the laws of current change governing the currents flowing through each electromagnet, during transmission by a station using a battery and the corresponding currents during transmission by a station without a battery, show that the current changes in each electromagnet depend on which transmitter is performing the transmission.

This, as is known, is explained by the presence of transmission-line leakance and is one of the causes of distortion in transmission (Bibl.6).

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OSCILLOGRAPH TECHNIQUE FOR MEASURING THE "HUNT" OF THE SCANNING AND TRANSMITTING PART OF FACSIMILE APPARATUS

by

M.A.Kudryashov, P.N.Ivanov

Problems involved in measuring irregularity of motion by facsimile scanning mechanisms are considered. A technique for measuring the hunting of the scanning instrument of a transmitting facsimile apparatus is described.

One of the main parameters determining the quality of the operation of a facsimile setup is the phenomenon known as "hunt" on the part of the scanning and recording systems.

When the scanning and recording systems are operating with insufficient stability, with irregular or oscillatory motion of the scanning and recording elements along the scanning line (i.e., with "hunting"), this manifests itself in a distortion of the outline of letters, figures, and other details in the recorded image. The contours of detail in the copy become blurred and take on ragged edges. These distortions are due to the fact that the picture elements (elementary dot areas) of the scanner and recorder do not coincide on the copy.

In order to obtain good definition in the reconstructed image, the amount of hunt in the recorded copy should not exceed + 0.05 mm in the case of photorecording. To fulfill this condition, very precise working of all sections of the transmitting and receiving system is required; the basic details of the transmitting and receiving mechanisms should be made to very close tolerance.

Since the transmitting and receiving portions of the apparatus are, from the standpoint of design, independent mechanisms, the tuning, adjustment, and quality control of their operation should be performed independently. The tolerance for the magnitude of "hunting" in recorded copy is a complex magnitude, consisting of the tolerances for oscillatory motion of the scanning devices in the receiving and transmitting sections of the apparatus. It is therefore preferable to specify the measurement technique as a separate, independent measurement of the hunting of the receiving and transmitting sections of the apparatus.

At present, the receiving section of the apparatus is tested for hunt in the following manner: Alternating current is fed from a single generator into the electromechanical drive and the recording stage of the apparatus (a half-wave rectified detector stage). On blank copy, in the receiving section of the setup, the frequency of the periods of the current delivered is recorded in the form of picture-element areas. In the ideal case, the picture-elements should be arranged in each scanned line at strictly uniform spacing, determined by the dimensions of the scanning element and the frequency of the current supplied. In this way, the recording of the "hunting" assumes the form of parallel lines composed of picture elements; these lines show up on the blank in the direction of feed, i.e., perpendicular to the scanning line. Deviations from the ideal arrangement of the picture elements determine the magnitude of hunting by the receiving section of the apparatus. The tolerance for the amount of hunting is given as the allowable linear deviation of the boundaries of the picture elements, both in the positive and negative directions. The magnitude of hunting is measured with an instrument microscope or a magnifying glass with divisions to hundredths of a millimeter.

The transmitting section of the facsimile setup is tested in the following manner:

1) Similarly to the receiving section, where there is a possibility of realizing an "inverse" electro-optical scanning system, i.e., instead of an illuminating lamp or photoelectronic multiplier, a point-source gas-discharge lamp (light modulator) is installed, and in place of the copy to be scanned, photosensitized paper is used, the necessary measures being taken to safeguard the paper from being exposed to stray light during the recording;

2) In the receiving section setup, with the aid of recording "test cards" (special graphic charts), with a preliminary testing of the "natural hunting" of the receiving apparatus.

In addition to the techniques described above, applicable for testing the hunt, similar measurements can be performed with a cathode-ray oscillograph. This technique has not been in use up to the present, although it is suitable for testing the magnitude of hunting by the transmitting section of facsimile equipment.

Measuring the amount of hunting through the intermediary of the cathode-ray oscillograph can be carried out in the following manner:

1) The input of the horizontal sweep amplifier of a CR oscillograph, (an EO-5 or an EO-7, for instance) is connected to the output of a tuning-fork oscillator or the output of any other fixed-frequency oscillator with high stability. The horizontal sweep is synchronized with the oscillator frequency, one period of oscillation of the generator being traced on the oscillograph screen in a sweep of maximum possible length in the horizontal direction.

2) The input of the vertical sweep amplifier of the oscillograph is connected to the output of the electric scanning system to be tested.

3) The electromechanical setup (electromechanical drive) of the system under study is excited by the same oscillator used for actuating the horizontal sweep of the oscillograph. It must be kept in mind that it is not obligatory to fulfill this particular condition: It is feasible to energize the electromechanical section by using some other high-stability oscillator, a multiple of the first; however, in this case, the accuracy of the "hunting" measurements will drop off by an amount proportional to the degree of divergence between the frequencies of the two oscillators.

4) At any point in the line-scanning plane, perpendicular to the direction of scanning, a thread is stretched or a line drawn, differing in optical density from the basic field of the test card.

5) The scanning system of the transmitting section of the facsimile setup is triggered. A vertical peak periodically appears in the trace on the oscillograph scope, once in the course of each horizontal sweep.

The overshoot on the screen will appear at the instant the thread or line of the test card is intersected by a ray of light emanating from the facsimile scanning system. Observation of the trace overshoot readily shows that it undergoes a displacement ("hunts") along the oscillograph screen in a horizontal direction. The magnitude of the displacement of the overshoot wave front in the horizontal direction is evaluated by comparison with a scaled grid. This magnitude is a multiple of the amplitude of the "hunting" of the scanning system under study.

The formula for computing the magnitude of hunting by the transmitting section of the facsimile apparatus is derived, using the following data as a point of departure:

1) Let the transmitting section of the apparatus produce a line trace of the image L mm in length at a speed of v lines/min, with the current being fed to a synchronous motor from a generator of a frequency of f cps.

2) One period of the generator current is scanned on the oscillograph in a horizontal direction for a length of A mm.

3) The displacement of the leading edge of the overshoot peak, measured on the oscillograph screen by comparison with the scaled grid, is equal to B mm.

Let us establish the proportion: An A mm deviation of the oscillograph sweep beam corresponds to $\frac{1}{f}$ sec, B mm corresponds to x sec, from which

$$x = \frac{B}{Af}.$$

A line of length L mm is traced during the time

$$t = \frac{k}{v},$$

where k denotes a conversion factor, from minutes to seconds, equal to 60.

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Taking this into account, let us establish the other proportion: t sec corresponds to L mm; x sec corresponds to y mm, from which

$$y = \frac{Lx}{t}, \qquad (1)$$

where y denotes the amplitude of "hunting" on the part of the scanning system, given in mm. Substituting the values x and t into the right-hand side of eq.(1), we obtain a formula with which to compute the magnitude of hunting by the transmitting section of the facsimile apparatus

$$y = \frac{BLv}{60Af}. \qquad (2)$$

Example. The transmitting section of the facsimile setup operates at a speed of v = 120 lines/min. The feed to the electrosynchronous drive is carried out by delivering current at a frequency of f = 2 kc. The length of the line to be scanned is L = 220 mm.

On the oscillograph screen, one period of the generator current is traced over a length of A = 100 mm. The leading edge of the overshoot traced on the oscillograph screen is displaced by an amount B = 20 mm.

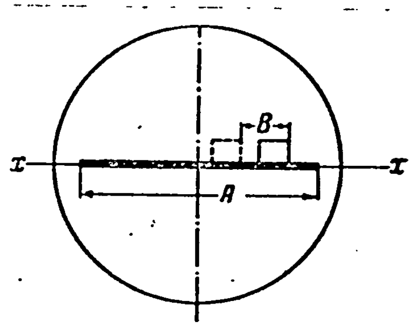

Fig.1

Substituting these data in eq.(2), we obtain the magnitude of the amplitude of the "hunt" expressed in millimeters,

$$y = \frac{20 \cdot 220 \cdot 120}{60 \cdot 2000 \cdot 100} \text{ mm} = 0{,}044 \text{ mm}.$$

A view of the pulses traced on the oscillograph screen is shown in Fig.1.

The error in the amplitude of "hunting", computed on the basis of eq.(2), is determined in the following manner: At first, by taking the logarithm and differentiating eq.(2), a formula is found for determination of the relative error

$$\ln y = \ln B + \ln L + \ln v - \ln 60 - \ln A - \ln f.$$

Differentiating eq.(3), we obtain

$$\frac{dy}{y} = \frac{dB}{B} + \frac{dL}{L} + \frac{dv}{v} - \frac{dA}{A} - \frac{df}{f}.$$

Taking into account the fact that the error in frequency $\frac{df}{f}$ and the error in speed $\frac{dv}{v}$ are equal as to absolute value and always have the same sign, these values can be reduced. The term $\frac{dA}{A}$ should be taken with the plus sign, since the sign of this value was unknown at the outset.

Expressing $\frac{dy}{y}$ in percentage, we obtain the formula

$$\frac{dy}{y}\,\% = \left(\frac{dB}{B} + \frac{dL}{L} + \frac{dA}{A}\right) \cdot 100.$$

It may be inferred that the error in measurements based on the oscillograph screen with the application of the millimeter scaled rule will be of the order of 1 mm. Then, in the case of the example considered above, the relative errors associated with the measured values will be as follows:

$$\frac{dA}{A} = \frac{1}{100}, \quad \frac{dB}{B} = \frac{1}{20} + K.$$

Here, K denotes the error resulting from the irregularity of the cathode-ray oscillograph sweep on which the measurements were carried out, i.e., the error resulting from the linear form of the oscillations of the sawtooth-wave generator and from the nonlinear sweep of the trace beam due to the deflecting plates of the cathode-ray tube. In the worst case, this nonlinearity may reach 5% of the entire length of the line (according to technical data obtained in using the EO-7 oscillograph). Taking into account the fact that the displacement B of the leading edge of the pulse takes place across a section of the line scanned, this error is lessened proportionally, to be exact

$$K\% = 5\%\,\frac{B}{A}.$$

Since $\frac{B}{A} = \frac{1}{5}$, then, accordingly, the magnitude of the error resulting from the

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nonlinearity of the sweep will amount to K = 1%, and the total magnitude of the error will be

$$\frac{dB}{B} = \left(\frac{1}{20} \cdot 100 + 1\right)\% = 6\%.$$

The relative error of the value $\frac{dL}{L}$ is determined by the possible accuracy of the measurement made on the length of line scanned and may in any case be equal to 0.5%, which corresponds to the measurement made with a precision of ± 1 mm. Summing up all relative errors, we obtain the actual error in percent

$$\frac{\Delta y}{y}\% = (1 + 6 + 0,5)\% = 7,5\%.$$

In relation to the measured value of the "hunting"

$$y = 0,044 \text{ mm}$$

the absolute error in the linear measurement amounts to

$$\Delta y = \frac{0,044}{100} \cdot 7,5 \text{ mm} = 0,0033 \text{ mm}.$$

Thus, y ± Δy = 0.044 ± 0.003, and the absolute error in measurements for the facsimile apparatus has no practical significance. Actually, the human eye, viewing an image at a distance of 250 mm (best visibility), distinguishes only those errors in the image which are of the order of ± 0.05 mm.

The value of the flyback of the CR oscillograph beam amounts to less than 1% of the forward trace (at frequencies of 1 to 8 cps). For that reason, the error introduced by the retrace of the beam into the total error is not large and may be safely disregarded.

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SYSTEMS DESIGNED FOR REDUCTION OF THE TELEPHONE SIGNAL BANDWIDTH

by

G.I.Tsemel'

The article considers various systems for speech-spectrum compression, developed recently as well as several similar arrangements developed previously.

Introduction

The application of the general communications theory to speech transmission has shown the possibilities of considerably compressing the spectrum of frequencies transmitted in communications channels.

The increase in demands made on telephone channels has spurred work in the direction of realizing these possibilities in practice. Depending on the principles forming the basis of speech bandwidth compression and the requirements governing the quality of the reproduced speech, the systems considered below achieve a reduction in the transmitted bandwidth by 2 - 100 times.

Vocoder

A typical example of a speech transmission system in which it is basically possible to conserve not only intelligibility but also, to a certain extent, the individual features of the caller's voice, is the arrangement known as the Vocoder, worked out in its primitive form during the prewar period (Bibl.1). In designing the Vocoder, features of a mechanism for remaking the sounds used in speech were employed.

The original speech, at the transmitting end of the vocoder, is transformed, with the aid of the analyzer and its band filters, into a series of separate parameters (Fig.1). One of these represents the character of the sound pronounced by the caller (whether it is voiced or unvoiced or a combination of both). The second par-

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ameter is the fundamental frequency of the sound produced by the vocal chords. Since the frequency of each of its harmonics varies with the fundamental frequency, this

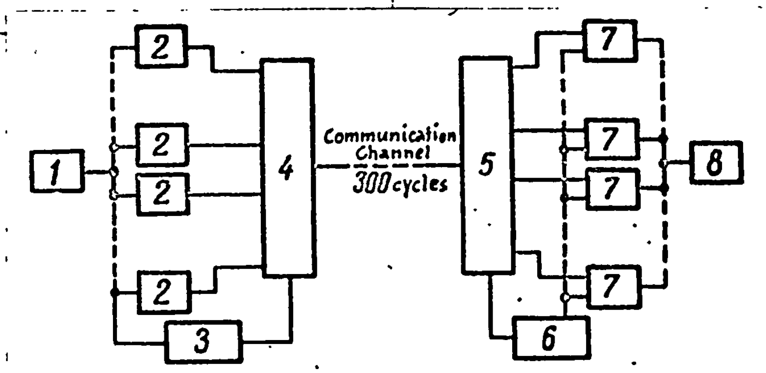

Fig.1

1 - Speech sources; 2 - Band filters and devices for measuring energy; 3 - Analyzer of the fundamental tone and character of the voice; 4 - Combining equipment; 5 - Distributor; 6 - Energy source for reconstructed speech; 7 - Devices for regulating the spectrum of individual signal components; 8 - Speech receiver.

one parameter alone is sufficient to give all the harmonics. The third parameter is the speech energy distribution over an energy spectrum, characterizing the individual formants of vowels and voiced consonants, as well as the remaining consonants. The indicated speech parameters are transmitted in coded form down a communication channel with a bandwidth of roughly 300 cps. At the receiving end, the control signals set up a fundamental frequency in the speech generator (the energy source for the synthetic speech) and control the energy spectrum of the reorganized speech to correspond with the original. In the case of unvoiced sounds, instead of the sound generator, a noise generator is inserted.

A Vocoder setup was worked out in England, on a plan similar to the one constructed earlier in the USA. A paper (Bibl.2) describing the British variant of this system outlines the circuitry of its separate units in great detail.

In the earlier Vocoder, the received speech was not sufficiently natural and had diminished intelligibility. Later improvements led to a successful application of the Vocoder in wartime. Comparatively recently, this system was improved to such a degree that it is often difficult to tell the difference between the transmitted (original) speech and the received (reconstructed) speech (Bibl.3). The differences in question are often easier to detect in comparing the spectrograms of transmitted and received speech, although the latter are also quite similar in appearance. A correct reconstitution of the fundamental tone of the transmitted speech still encounters difficulties. Research on this problem is being undertaken at the present time, leading in several directions.

Speech Transmission Systems Without Preservation of Individual Voice Features

Further noticeable narrowing of the speech spectrum, beyond that achieved with the Vocoder, can be attained where only the requirements of speech intelligibility are met, without preserving the individual features of the caller's voice. The problem, in that case, consists in recognition of the sound (phoneme), for example by its spectrum, or by transmission of the number of phoneme and reconstruction of the latter at the receiving end, with the aid of a source for artificial speech.

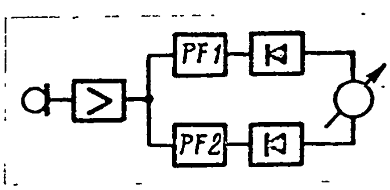

Fig.2

Consider first the setups where recognition of sounds has been in application earlier. The principles of the objective recognition of speech sounds, based on the differences in the energy distribution associated with different phonemes along the spectrum, are presented in a paper by L.L.Myasnikov (Bibl.4). If the speech source is connected with a DC voltmeter to two circuits each consisting of a band filter and a rectifier (Fig.2), then, depending on the sound uttered, the pointer of the measuring device will deflect to the right or to the left, or will remain at neutral. The direction of the pointer deflection depends on the passband of the filters and not on the loudness, timbre, or intonation characteristic of i

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voice. For example, where the passband of PF 1 is 500-700 cycles and that of PF 2 is 800-1000 cycles, the sounds ah, yeh, eh will cause a deflection in one direction, while the sounds oh, yoo will cause a deflection in the other. The sound ya will result in no perceptible deflection of the voltmeter pointer. If the passbands of the filters are taken at 1250-1500 cycles and 4000-5000 cycles, respectively, then, in response to the sound eh, the direction of the deviation will change, while for ah it will remain unaffected. At a proper choice of the passband, for several pairs of filters, all the vowels and consonants can be divided. Utilizing the above principles, L.L.Myasnikov designed a device to effect the recognition of speech sounds, namely a dynamic analyzer, consisting of the analyzer proper and of a recognizer (a relay-type decoder). The result of the action of the device is recorded by the flashing of the corresponding light on the panel. This has resulted experimentally in establishing the possibility, in principle, of actuating a small typewriter through the dynamic analyzer.

This possibility could be made a reality by using a "phonetograph", intended for the phonetic transcription of speech, using the conventional letters (Bibl.5). The device consists of an "electronic ear", carrying out the spectral analysis of speech; a relay-type decoder, seeking out the letter corresponding to the phoneme in the spectrum; a small typewriter with a solenoid-controlled keyboard drive. Determination of the sounds is performed with the aid of six filters, analyzing a band from 150 to 4000 cycles and separating out the formants of the vowels. Three additional filters are required to accomplish the sorting of the consonants. The speech gains access to the instrument through the microphone and is automatically interrupted 5-15 times a second. The frequency of alternation should be equal to the number of phonemes per second. After passing through the filters, the pulses are rectified and differentiated. Negative peaks are used for synchronization (in order to make it possible to observe the spectrum of the phoneme), while positive pulses serve to form the spectrum of the phoneme, which spectrum is used for making a single-valued

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determination of the phoneme; the pulses finally are fed to the decoder. Only 5% of the range of the sound frequencies is used for selection of the letters. The printing recorder prints 30 letters, representing French phonemes. The setup may be adapted to any language. Pauses are expressed by spaces between the letters. On a 200-cycle telephone channel, 15 "phonetograms" (spectra of phonemes) can be transmitted simultaneously. Accordingly, here a speech bandwidth compression of roughly twenty is achieved. A model of the device was demonstrated in October of 1952, in Geneva.

In 1952, a device (Bibl.3,6) was developed to effect the recognition of numbers uttered by a voice (Audrey; automatic digit recognizer). In this setup, the spectrum of the sequence of sounds associated with the pronounced digit was analyzed. The data obtained were compared with the spectra of the sound sequences of all ten digits, recorded earlier. As a result of the comparison, the device "decided" to which number the spoken word corresponded most closely. The device operated properly in response to the utterances of numbers by a voice to which it was tuned, and also to voices of a similar timbre.

Its response to other voices, to faulty pronunciation, or to an accent sometimes produced a wrong result. An analysis of the errors showed that they occurred mainly in response to an analysis of a whole word, rather than to individual sounds. It was found that, in its response to the pronunciation of a given number (as, for that matter, its response to any other words), the sequence of sounds, depending on the pronunciation by various persons, was different.

An accurate study of discrete sounds of speech showed that the instrument was useful for the designing of a new system of communications.

In the USA, a system of speech transmission has been developed which is based on the recognition of sounds at the transmitting end and on transmission of only the number of the phoneme along the communication channel (Bibl.3). At the receiving end, the sounds are reconstructed by the artificial speech source, analogous to the

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receiving section of the Vocoder, except that the fundamental tone of the vocal-chords imitator (artificial larynx) does not change (Fig.3). As already indicated, this leads to a loss of the individual voice features of given persons, although some of the features of enunciation naturally remain. This type of reconstructed speech therefore contains somewhat more information than the conventional telegraph message. The first such system was constructed with the contributions of only 10 phonemes. The remade speech proved to be quite understandable, especially when very familiar words were pronounced. Inasmuch as the rate of repetition of phonemes in normal speech does not exceed 10 utterances per second, extremely narrow-bandwidth channels are adequate for the transmission of speech by this system. According to other data available (Bibl.7), the width of the channel band, as suggested, should be reduced to 30 cycles.

A system similar to the one considered above was demonstrated in May of 1956 in England, and was described in two brief notes (Bibl.8,9). The speech spectrum in

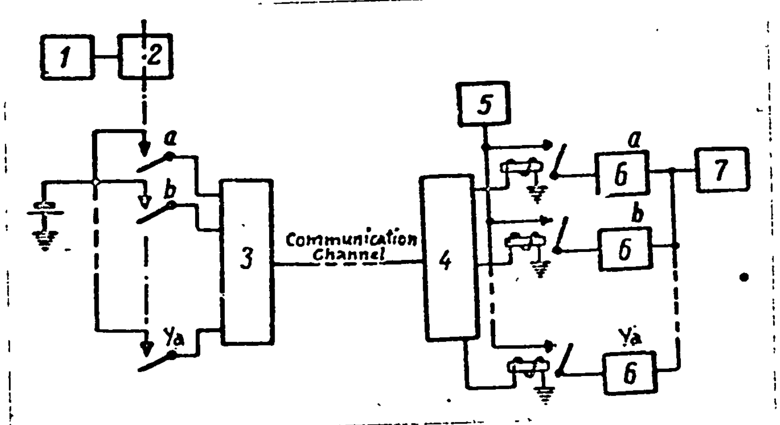

Fig.3

1 - Speech source; 2 - Phoneme analyzer; 3 - Combining equipment; 4 - Distributor; 5 - Energy source for artificial speech; 6 - Devices used to establish the bandwidth of discrete phonemes; 7 - Speech receiver.

this case is compressed roughly a hundred times and is comparable with the spectrum of a telegraph signal. During the demonstration, the transmitting and receiving

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parts were connected by a six-conductor transmission line, although, in principle, signals could be coded for transmission along a single channel. The device is still in its earlier stages of development and, in evaluating the quality of the sound produced, the impression created is something like a person speaking with his "mouth full of potatoes".

Speech Transmission Systems Using Frequency Dividers

The increasing demand for telephone channels during wartime led to the division, in a number of cases, of the normal channel bandwidth into two narrow bandwidths, each ranging from 300 to 1700 cycles. A certain number of the divided channels, in spite of their reduced transmission quality, are still in use in the USA, up to the present time. This shows that the working out of systems permitting a lesser compression of the speech bandwidth, but possessing simplicity of design, reliability, and good transmission quality will stimulate interest.

In the USA, a speech bandwidth reduction system known by the name "Vobanc" and achieving a bandwidth compression by two times, has been designed (Bibl.3,10). The operation of this system is based on the fact that vowel sounds are entirely determined by three formants. The first formant usually lies below 1000 cycles; the second, in the 1000-2000 cycle bandwidth; the third; above 2000 cycles. At the transmitting end, the signal bandwidth, after being shifted into the 107.8-104.8 kc range, is divided by band filters into three bands (Fig.4). The divider-circuit modulator is hooked up to a ring circuit. When the modulator signal at the input is below a certain minimum, no signal appears at the output. As indicated (Bibl.10), the frequency divider halves the frequency of the most powerful sinusoidal component in each of the three bands. The dynamic range remains linear within the limits of 30-35 db. The other components, found in the neighborhood of the maximum-power component, are shifted with the same shift in frequency as at the divider input. In that way, the division of the frequency of the maximum-power component does not mean that the entire bandwidth is simultaneously divided by the same value. The reduction of the three bands by a factor of two, effected by the divider filter, leads to a loss of several components which are, however, components of low power and having little effect on the quality of transmission.

It may be inferred that, for example, the two-divider setup could consequently be used for achieving a further bandwidth compression by two times. However, in that

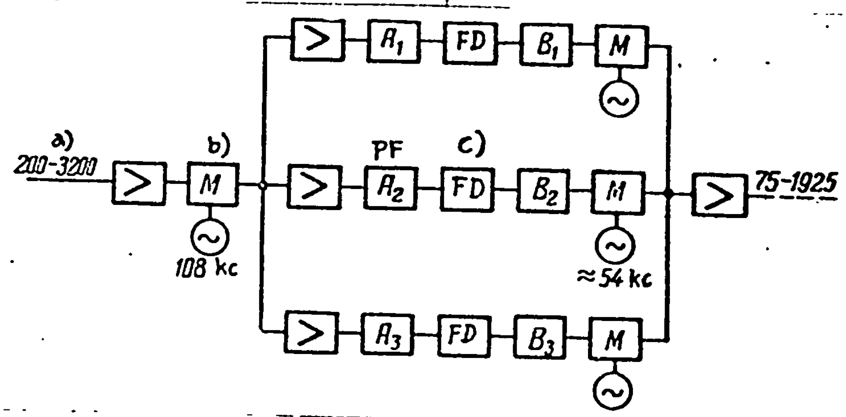

Fig.4

a) Input 200-3200; b) Modulator; c) Frequency divider

case, the width of each band, after passing through the divider filter, would extend to 250 cycles, which would cause cutoff of most of the harmonics of the fundamental frequency, of which the formants are composed.

Separate modulators were inserted for each of the three bands after the divider, and also at the receiving end, with the aim of obtaining improved selectivity and avoiding transient noise. At the receiving end, the transformation of the speech spectrum proceeded in reverse order (Fig.5).

Two variants of the Vobanc have been divised for duty on the normal telephone channel bandwidth with multiplexed transmission: one, taking up a bandwidth of 75-1925 cps, the other, a bandwidth of 2075-3925 cps. In articulation tests, the Vobanc was compared with the normal telephone channel of 3500-cycle bandwidth. In the first test, the coefficient of articulation amounted to 79% and 89%, respectively and increased to 91% and 93% in the fifth test. The listeners thus quickly adapted themselves to the Vobanc and the noticeable difference became negligible. When subjected to comparison with direct transmission using a narrow-bandwidth channel (200-1700 cycles), the two methods produced the following results from a second team of listeners (48 persons): Vobanc: 79.7%, narrow bandwidth: 65.6%. In some cases, characteristic noise distortion was perceived, taking the form of "burbling"

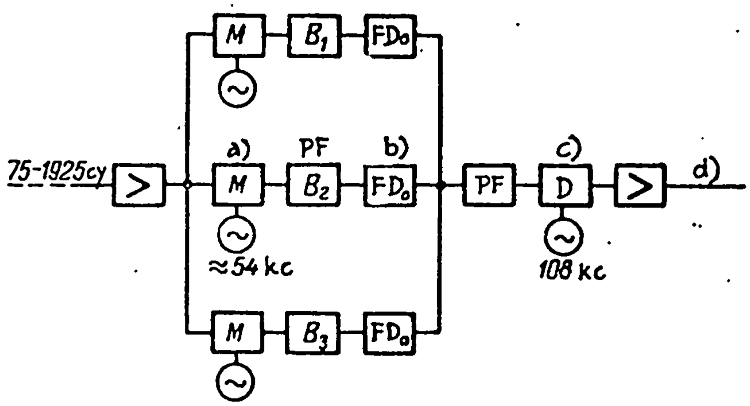

Fig.5

a) Modulator; b) Frequency doubler; c) Demodulator; d) Output

in response to input noise and fricative sounds. The noise in question always affected the quality of reception more than it did the intelligibility. Accordingly, Vobanc in some respects is inferior to the normal channel and considerably outperforms the narrow channel with respect to direct transmission.

In 1955, in a paper read before the International Conference on Information Theory, Marcou and Daguet outlined a new method of speech transmission (Fig.6), entailing a reduction of roughly three in the transmitted bandwidth (Bibl.11,12,14).

The authors considered the speech signal as a complex-modulated (modulated both with respect to amplitude and frequency) oscillation $A(t) \cos \varphi(t)$. Where the signal spectrum of $\varphi(t)$ is actually narrower than the speech signal spectrum, the system in question can be used with comparatively narrow channels. In their report, Marcou and Daguet failed to provide any indication, in particular, of the extent to which the channel bandwidth could be compressed, as a function of the frequency

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scaling ratio. In addition, there is some indication of a deterioration of the signal after it is passed through the frequency-divider filter. This filter must have a smooth crossover from the passband to the stop band, which leads in practice to some degree of widening of the transmission band. It should also be noted that, in

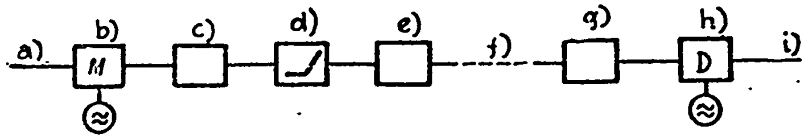

Fig.6

a) Input; b) Modulator; c) Limiter; d) Low-pass filter; e) Frequency divider; f) Communications channel; g) Frequency multiplier; h) Demodulator; i) Output

the later publication of the report, the section devoted to narrowing the speech bandwidth by means of frequency dividers is entirely left out (Bibl.13).

Thus, the effectiveness of the bandwidth compression system proposed by Marcou and Daguet may be cleared up only as a result of further research. Experimental studies carried out recently have not yet confirmed the notion that the spectrum of $\varphi(t)$ is materially narrower than the speech-signal bandwidth.

In conclusion, it should be kept in mind that the problems involving compression of the speech bandwidth present great interest from both the theoretical and the practical side, and new work in this area of investigation is to be anticipated in the immediate future.

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RESOLUTIONS OF THE EIGHTH PLENARY SESSION OF THE CCIR ON PROBLEMS OF TELEVISION

by

M.I.Krivosheyev

Documents and resolutions on problems of television which were considered and adopted by the Eighth Session of the CCIR in the fall of 1956 are presented.

The Eighth Plenary Session of the International Consultative Committee on Radio (CCIR) took place in August and September 1956 in Warsaw. The XI research commission is the department of the CCIR charged with handling problems of television. Because of the large number of problems which were to be discussed at the meeting, the XI commission was divided into five subcommittees, subcommittee XI-A for color television; subcommittee XI-B for the development of requirements to be met by communication channels for long-distance transmission of television signals; subcommittee XI-C on black-and-white (monochrome) television standards; subcommittee XI-D for quality of television pictures; subcommittee XI-E for interference ratios in allocating frequency channels to television stations.

Color Television

The demonstrations of various systems of color television in the USA, England, Holland, and France in the spring of 1956 preceded the sessions of the XI research commission at the Eighth Plenary Session of the CCIR in Warsaw. The results of the demonstrations were cited in the report of the president of the XI commission (Bibl.1). An All-Europe color television system was to have been established in Warsaw.

The discussion on this problem indicated that the majority of the countries, upon carrying out research on the development of a color television system, give preference to an adaptation to European conditions of the system developed in the USA in 1953 by the National Television Systems Committee (NTSC).

For instance, the delegates of the Federal Republic of Germany and of Switzerland reported that tests on the modified NTSC system with 625 lines of resolution are in progress in their countries. The Danemark delegation expressed its preference for the NTSC system. The Holland delegation also stated that the NTSC system largely meets the requirements made on color television systems, as formulated by the XI commission in 1955 at its Brussels meeting. Simultaneously, the Holland delegation reported that, in the spring of 1956, the "Philips" system with two subcarriers was demonstrated in order to show the possibility of developing color television systems which have certain advantages over the NTSC system (for instance, absence of color distortions due to phase distortion since color television is transmitted over long-distance communication lines). However, the delegation indicated that the advantages of the future single European color television system are more substantial and that Holland does not insist upon the adoption of its system.

The delegations of the USSR reported that, in the Soviet Union, the basic efforts are directed toward the development of a fully compatible color television system with one color subcarrier and square components, since this is the most fully tested system at present.

The delegation from France and of England objected to the selection of any color television system at the Eighth Plenary Session, motivating their objection by the insufficient amount of research which had been carried out. At the same time, the delegation from England reported that investigations of the NTSC system for 405- and 625-lines standards are being carried out in England as well, but that the results obtained so far are insufficient for a prospective color television system to be selected.

The French delegation, in contrast to the other delegations, reported that laboratory investigations of systems different from the NTSC system are being carried

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out in France and that no final conclusions can be drawn as yet.

In the course of the discussion, the delegations from the Federal Republic of Germany and Italy reported that the frequency channels in the bands I (48-100 mc) and III (174-224 mc) are not satisfying their countries' needs for the development of a black-and-white television network and that, therefore, they will be forced to allocate the bands IV and V (470-940 mc) to black-and-white television stations, without waiting for the choice of an All-Europe color television system to be made.

The delegations from England and France and the representative of the European Broadcasters' Union (EBU) deplored the declarations of the Federal Republic of Germany and of Italy as to their refusal to adhere to their obligation to set aside the frequency bands IV and V for a single Pan-Europe color television system, an obligation which they contracted in 1955 in Brussels.

The representatives of the Federal Republic of Germany and of Italy, nevertheless, confirmed that, in the near future, black-and-white television stations will be put into service on bands IV and V; at the same time, they indicated that if the single color television system for all European countries will be adopted, they will adopt this system for transmitters which, at that time, will operate in the bands IV and V.

The delegation from France insisted that a comparison of the NTSC system and of the color television systems being developed in France should be made (Bibl.1). Despite the fact that the discussion of this question lasted through three sessions of the subcommittee XI-A, it yielded no definitive results and was discontinued due to the difficulties that arose in carrying out a comparison of the qualitative and technical characteristics of the NTSC system, which is already in service and has been tested sufficiently, with the systems being developed in France, which are essentially at the stage of laboratory experiments.

The delegation of the USA reported that the USA does not intend to change their point of view on the selected NTSC system.

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As a result of the discussion on color television systems, the committee arrived at the conclusion that, at the present time, only the NTSC system has been sufficiently tested in practice to be acceptable for consideration as a basis for standardization. However, this must not constitute an obstacle to the continuation of the investigations to the possibility of the acceptance of other standards if their practical tests are satisfactory.

Thus, due to the lack of sufficient data for a color television standard to be selected, as well as to the lack of unity of opinion on this problem, the Eighth Plenary Session of the CCIR adopted no definitive resolutions on a color television system for Europe. It was decided that this problem will be reconsidered at the next session of the XI commission, which is to take place in 1958 in the USSR.

The second European Conference for the revision of the Stockholm plan for allotment of frequency channels for television and ultrashort wave FM broadcasting in the bands I, II, and III is supposed to take place after the session of the XI Commission in 1958 as well. At this conference, a new plan for the allotment of channels in the bands IV and V must also be drawn up. By that time, the CCIR research committee V is to prepare precise documents on the propagation of radio waves in these bands.

Requirements on Communication Channels for Long-Distance Transmission of Television

The discussion of the above-mentioned standards which was taking place in the subcommittee XI-B, concerned the proposed standards defined in 1955 in Brussels (Bibl.2), which included new propositions by several countries. As a result, a number of new requirements as to the qualitative characteristics of television channels of radio-relay lines and cable lines were developed (Document 907). These included methods of measuring the nonlinearity of the amplitude characteristic, the amplication factor of the communication channel, as well as the allowances for transient characteristics in the domain of low and medium video-frequencies, etc.

All these characteristics have been established for black-and white television signal transmission, as applied to a long-distance line of 2500 km along which de-modulation of the received signals and the consequent modulation of the carrier frequency at intermediate points will not take place more than twice. It was decided that, at the present time, that the outlined requirements for such a line should be used rather than the standards established for the existing and the planned long-distance lines designed for the transmission of television signals.

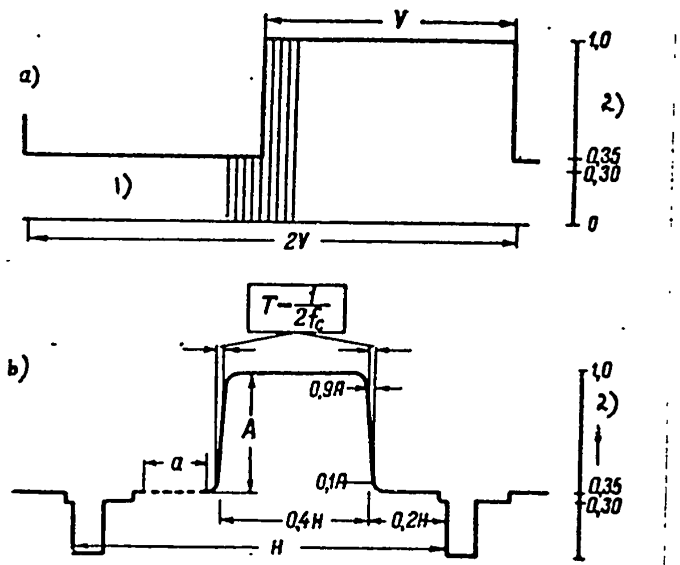

Fig.1

1) Synchro-pulse; 2) Volts

The committee gave preference to methods of evaluation of linear channel distortions by means of transient characteristics. The transient characteristic of a channel must be measured at three frequency ranges: low, medium, and high. The test signals, shown in Fig.1, are recommended for measuring the transient characteristic of a channel. The test signal shown in Fig.1a is used for testing the transient characteristic in the low-frequency range. The signal is composed of rectangular symmetric pulses, in synchronization with the field frequency. For the clamping circuits of the channel to operate normally, these pulses are mixed with the blanking and synchronizing pulses. The test signal (Fig.1b) is used for testing the transient characteristic in the range of medium and high frequencies, as well as for measuring modulation of the received signals and the consequent modulation of the carrier frequency at intermediate points will not take place more than twice. It was decided that, at the present time, that the outlined requirements for such a line should be used rather than the standards established for the existing and the planned long-distance lines designed for the transmission of television signals.

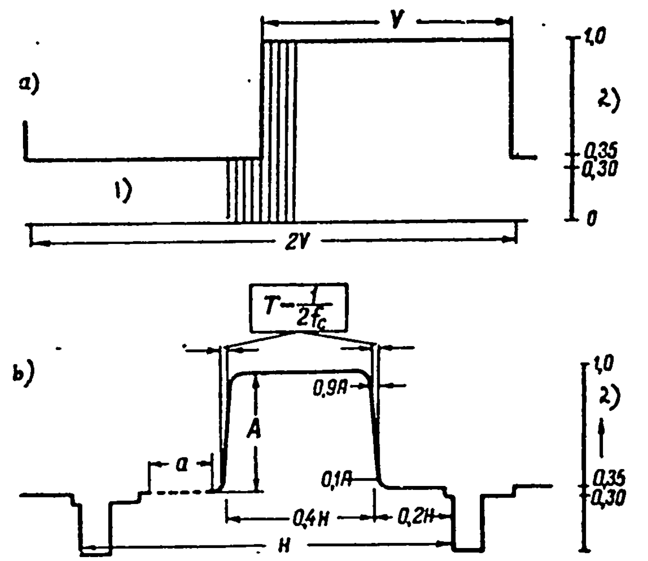

Fig.1

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The build-up time T of the leading edge must be equal to $T = \frac{1}{2f_c}$, where f_c is the highest limit frequency for the given television system. In addition, the use of pulses with a build-up time of the leading edge of 2T is permissible. Both diagrams show a protective section whose magnitude is 5% of the total signal amplitude, a section which is between the level of the test pulses and the level of the blanking pulses. However, in order to simplify the test equipment, a reduction to zero of this section is permissible. In order to regulate the build-up time of the leading edge of the rectangular pulse, which is the pulse of line frequency, the filter described by Thomson in the British magazine POIEE, Vol.99, Section III, 1952, p.373 is recommended as an example. Detailed data on such a filter, used for measurements in a system with f_c = 3 mc, are given in the minutes of the meeting, No.154.

a)

b)

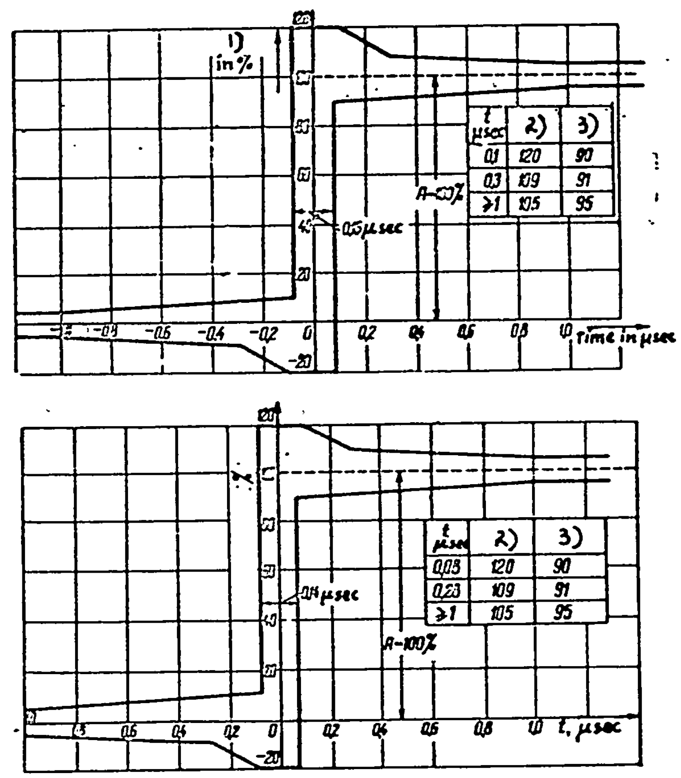

Fig.2

1) Amplitude; 2) Lower limit; 3) Upper limit

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Allowances common for all standards are proposed for low and medium frequencies. Thus, for low frequencies, the slope of the horizontal portion of the test signal (Fig.1a) must not exceed ± 10%. For medium frequencies, the slope of the horizontal portion of the test signal (Fig.1b) must not exceed ± 5%. Simultaneously, it is indicated that it is desirable to lower this tolerance to ± 3%. The indicated measurements at medium frequencies must be carried out on sections of the signal that are between the intervals $t_1 + 1\,\mu$ sec and $t_2 - 1\,\mu$ sec (where t_1 and t_2 are the instants of time at which the rectangular pulse voltage passes values that correspond to one half of its magnitude. The allowances for the transient characteristic in the high-frequency range are given for the 625- and 819-line systems.

Figure 2a shows the time allowances for this characteristic, for the television standard adopted in the USSR (625 lines, channel width: 8 mc).

For a 625-line system which uses a 7-mc channel and for the Belgian 819-line system, the allowance for the transient characteristic of the channel, under the test signal of Fig.1b are given in Fig.2b.

In Fig.1b, a section a is provided so that additional test signals can be inserted if necessary, for instance, sine-squared pulses, bunches of high-frequency sinusoidal voltages, etc. In particular, in England (405-line system) sine-squared pulses are transmitted in this interval (Bibl.3). Allowances for distortion of these signals are given in document No.154. Allowances for the amplitude-frequency and the phase-frequency characteristics (Bibl.2) are kept only as additional information for system designers.

Nonlinearity of the Amplitude Characteristic. No single method of measuring the nonlinearity of the channel amplitude characteristic was determined on the basis of the proposed standards which were developed in Brussels in 1955. Several methods (Bibl.2) were proposed for this purpose. At the session in question, it was decided to recommend that the nonlinearity for all standards be measured by means of the test signal shown in Fig.3a.

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The test signal in the range of video-signal transmission is composed of a sawtooth voltage with a superimposed sinusoidal voltage. In order to test the characteristic in the full voltage range, the variations of the constant component being

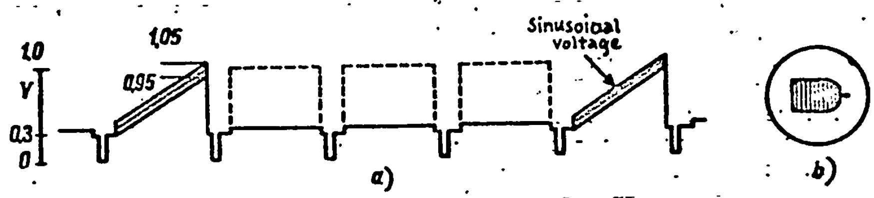

Fig.3

taken into consideration, the signal is alternated with three lines which may correspond either to the transmission of a black field or to that of a white field. This yields the conditions which correspond to a variation of the constant component. The frequency of the sinusoidal voltage for a 405-line standard is taken to be 2.5 mc and for 625- and 819-line standards, 4 mc.

At the output of the channel, a filter is installed which passes frequencies of 2.5 and 4 mc. Packets of sinusoidal voltage are observed on the oscillograph when the frequency of its horizontal sweep is equal to the frequency of the lines (Fig.3b). A change in the amplitude of this voltage indicates that the channel amplitude characteristic is nonlinear. The measurements are carried out twice, a first time when lines which correspond to the transmission of a black field are given (Fig.3a, continuous lines) in the intervals between the test sawtooth signals, and a second time, when white field lines are given in the intervals (dotted lines). The amplitude of the high-frequency signal on the oscillograph screen (Fig.3b) must not drop below 80% of its maximum magnitude.

When the above-described variations of the test signal take place, the oscillation amplitude of the synchronization pulses must not exceed + 10%, -30% of the rated value of -0.3 volts.

Channel Amplification Factor. The amplification factor of a television channel, when it is initially established and measured by the method given below, must re within the limits of ± 1 db. When originally installed, the oscillations of the channel amplification must not exceed the following limits:

a) For changes occurring within a short time (for example, one per second), ± 0.3 db;

b) For changes occurring in a medium period of time (for example, one per hour), ± 1 db.

It was decided that the measurement of the channel amplification factor be carried out by means of the test signal which contains blanking and synchronizing pulses and corresponds to the transmission of vertical black-and-white bands, each of which having the length of half a line (Fig.1b). When only the amplitude of the test signal at the input and output of the channel is measured, the amplitude between the white and black levels is taken into consideration (the synchronization signals are disregarded).

In order to avoid errors due to the distortion or the sloping of the horizontal portion of the test signal, the amplitude at the input as well as at the output of the channel must be measured between the points corresponding to the middle of each of the bands.

Black-and-White Television Standards

The subcommittee XI-C revised the report on the black-and-white television standards which are used in various countries. The basic parameters of the OIR (International Broadcasting Organization) television standard provide for a 625-line resolution of the picture, when the channel band is 8 mc and the spread between the carrier frequency of the television transmitter and the mean frequency of the sound transmitter is 6.5 mc. It was decided that the parameters of the OIR standards be included in this CCIR report.

Several countries have included tolerances for their gamma-characteristic coefficients and their television transmitters which, when the modulation characteristic of the receiving tube is taken into consideration, must be less than unity. F instance, in England, the coefficient is γ = 0.4 – 0.5; in the USA, γ = 0.45; in France, γ = 0.6; etc.

The subcommittee prepared a program for the investigation of problems referring to the correction of television signal distortions when a single sideband is transmitted (quadratic distortions, phase distortions in the transmitter and receivers, etc.). As these problems are studied, an answer must be provided to the problem of properly locating the correcting circuits for these distortions (in transmitters or receivers).

Quality of Television Images

The subcommittee XI-D discussed the papers on this problem, which were presented by the delegations of the USSR, the USA, the Federal Republic of Germany, and other countries. The discussion on the problem of the quality of a television picture showed that the lack of widely accepted television measurement methods constitutes a considerable obstacle to the establishment of single standards for the basic characteristics of television images.

In connection with this, it was decided that the methods for measuring the characteristics of television equipment as well as the various methods for evaluating the quality of television images, at present, do not depend on the television standards adopted in the various countries.

During discussion of the perception of television images, it was pointed out that the resolving power of the eye and the differentiating sensitivity in viewing moving television pictures, must be investigated.

A representative of France, Prof. Boutrie, remarked that a comparison of the quality of color images obtainable in the various systems must precede the selection of a pan-European color television system. Therefore, unified methods of rating the quality of television images must be developed. He accepted the proposition of the chairman of the CI commission to organize the coordination of the activities of the XI commission along this line.

Interference Ratios in Allotment of Frequency Channels

The subcommittee XI-E examined papers on this problem, presented by England, Poland, the Federal Republic of Germany, Czechoslovakia, and others.

The Polish paper (Doc.209) gave the results of an experimental investigation of the interference ratios necessary for a black-and-white 625-line television system with an 8-mc radio channel. The determination of these ratios was based upon the fact that they must satisfy the viewer's requirements as to picture quality during the first performances as well as in the future when the demands will become more severe. Therefore, person experienced in viewing television pictures were invited for consultation. The receiver which was used in the tests had the frequency characteristic shown in Fig.4. A picture corresponding to the signal being received was watched on the television screen; simultaneously, an interfering signal from the output of the test oscillator was supplied to the receiver input. The frequency of the interfering signal was changed in steps of 1 mc. The measurements were carried out in a dark room. The viewers were placed at a distance 4 times the height of the screen. The illumination on white was 3 nits, the picture contrast range was 10. The task of the viewers was, first of all, to determine the level at which the distortion becomes hardly noticeable and, secondly, to determine the distortion level which may be considered acceptable. During the measurements, in order to set up conditions close to reality, titles, test patterns, scenes from the studio, motion pictures were being transmitted. As a result of individual experiments with the test pattern, average numbers were obtained which are expressed by the curves shown in Fig.5. Curve 1 corresponds to the threshold discernible distortion, while curve 2 shows the intolerable distortion level. Curve 4, which was recommended by the delegation of Poland, corresponds to the levels of tolerable distortion; it was obtained from curve 1 by means of subtracting 6 db from its ordinates. The interference-ratio curve 3, recommended by the CCIR (Bibl.2, Report 34) is also given here for comparison (f_1 is the frequency of the distortion, f_2 is the carrier frequency

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of the picture transmitter).

It may be seen from Fig.5 that the recommended curve of the interference ratios reaches its maximum, approximately equal to 43 db, at a frequency difference of $f_1 - f_2 = 0.5$ mc. In the frequency band of $0.5 < f_1 - f_2 < 5.5$ mc, on logarithmic coor-

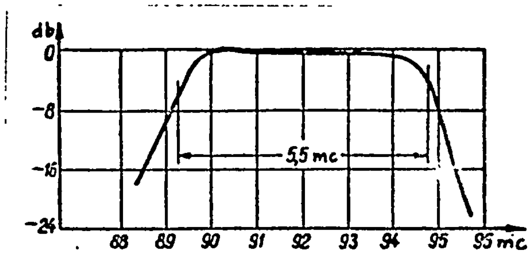

Fig.4

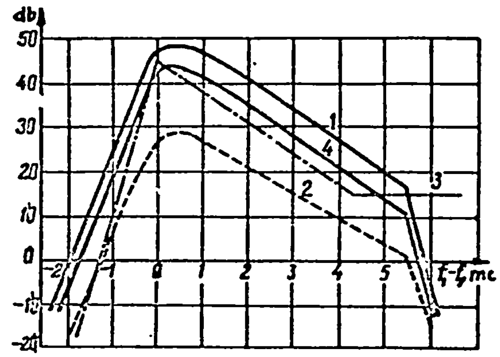

Fig.5

dinates, the curve falls off linearly and is parallel to the CCIR curve. Thus, it was proposed that the interference ratio be increased by 2 – 3 db as compared to the CCIR curve. At the point 5.5 mc, the curve falls to 11 db and after that slopes quite abruptly, bisecting the abscissa (0 db) at the point 5.7 mc. In the frequency band $f_1 - f_2 < 0$, the slope of the curve is steeper than that of the CCIR curve.

Figure 6 shows interference ratio curves for the case when the difference between the f_1 and f_2 is equal to the odd number $\frac{f_{cmp}}{2}$ (1 represents the threshold of discernible distortions; 2 the tolerable distortions; 3 the intolerable distortions). When these curves were being determined, the frequency of the distortion was changed in 1 mc steps, and the accurate determination of the oscillator frequency was carried out in accordance with the minimum of distortions on the screen of the television set. The following conclusions can be drawn from an analysis of these curves: at the difference $f_1 - f_2 \approx 0.5$ mc, the curve reaches a maximum of 30 db. The interference ratio at the carrier frequency may be lowered to 27 db if the same channel is used, which approximately corresponds to the CCIR standards. When $f_1 - f_2$ exceeds 1 mc, the interference ratio curve asymptotically approaches the CCIR curve, and the two curves coincide as they approach the frequency $f_1 - f_2 \approx 5.5$ mc.

It follows from the foregoing that, in the cases when the channels of the transmitters are superimposed, the use of a system in which the carriers are displaced is of practical advantage in the full range of the frequencies of the channel subject to distortions. If the large number of television centers being built and the limited

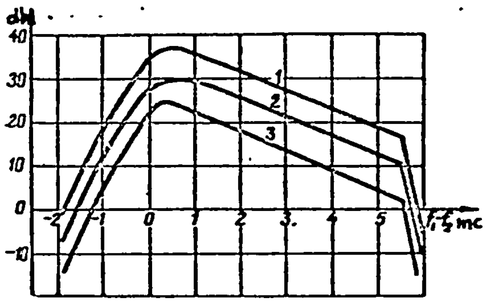

Fig.6

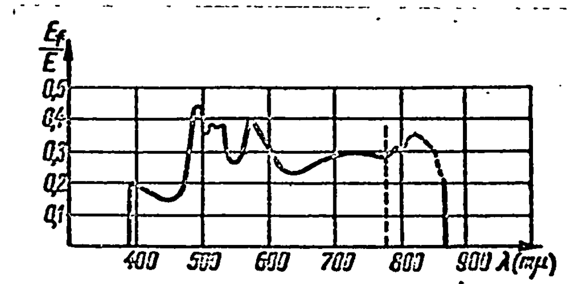

Fig.7

number of frequency channels for television is taken into consideration, the use of the carrier phase-shift method is of great interest. However, the advantages of this method depend upon the stability of the line frequency of the received picture as well as on the stability of the carrier of the picture in television transmitters and upon other factors.

Therefore, the resolution by the Assembly mentions the necessity of investigating the following questions: conditions and frequency bandwidth where the use of carrier phase-shift is advantageous; the advantages of this method when the difference between the frequencies of the interfering transmitters is considerable and when the synchronization connected to the feeding network frequency is used; the effect of the receiver vertical sync unit on the quality of the reception when phase shift of the carrier frequencies of the picture transmitters is used.

As a result of discussing the problem of the interference ratios, when channels are assigned to black-and-white television stations, the committee found it necessary to stress the fact that, upon determining these ratios, the frequency characteristics of television receivers must by all means be taken into consideration since

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the curve of the protective relationships given in the CCIR documents (Fig.5, curve 3) was plotted without sufficient regard to the latter.

The paper presented by the delegation of the Federal Republic of Germany (Doc. 240), gave the results of investigations of the United Committee of Radiobroadcasting Companies, investigations which expose the difficulties that may be encountered upon introducing color television based on a system analogous to the NTSC system when the frequency channels are handled by the bands I, II, and III, an allocation which now exists in Europe and was adopted at the European Radiobroaucasting Conference of 1952 in Stockholm. It was then assumed that the color subcarrier was located in the 4 mc region. The ranges of maximum saturated textile dyestuffs, typographical colors, etc. were used as basis for determining the relationships between the amplitudes of the color signals and the color wavelengths, which correspond to definite colors (Fig.7).

It follows from the diagram that the signal amplitude of a color of maximum intensity and frequency (E_f) comprises only 40% of the amplitude of the black-and-white video signal E, i.e., is at least 8 db less than the latter (the narrow section in the blue-green color range is neglected). The range between the black and white levels in the television radio signals makes up approximately 65% (3.7 db) of its total range (from zero to the level of synchronization signals). Accordingly, it may be expected that the maximum amplitude of color signals in radiotransmission will be approximately 12 db (3.7 db + 8 db = 11.7) below the level of the picture carrier.

The following particular cases are considered in the paper:

Black-and-White Television Distortions due to Color Television Signals. The measurements showed that the interfering signal which falls in the band of black-and-white television video frequencies in the 4-mc region, must be 25 db below the level of the picture carrier of the station being received in order that the distortions remain within the tolerance limits, and 32 db below in order that they remain im-

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perceptible.

In accordance with the Stockholm plan, the black-and-white television stations which operate on a common channel must guarantee 45 db interference ratios when the frequencies of distortion are close to the picture carrier frequency of the received station; when using a transmitter carrier phase-shift system, a 30 db decrease in this ratio is allowed. Since the color signals are 12 db below the corresponding picture carrier, they will not distort the reception of black-and-white television if the stations operating on a common channel adhere to the above-indicated interference ratios.

Distortion of Color Television Reception due to Black-and-White Television Stations. If the frequency of the distortion is close to the frequency of the color subcarrier, the low-frequency pulses generated in the receiver will be quite noticeable in the picture. Figure 8 shows curves of the minimum allowable interference ratios of the useful signal to the interfering signal as a function of the frequency difference between the carrier of the station being received and that of the interfering signal. These curves were borrowed from another paper (Bibl.6) and were recalculated so as to be applicable to a 625-line system (radio-channel width: 7 mc). It can be seen from these curves that in the radio-frequency band extending up to 2 mc from the picture carrier, the

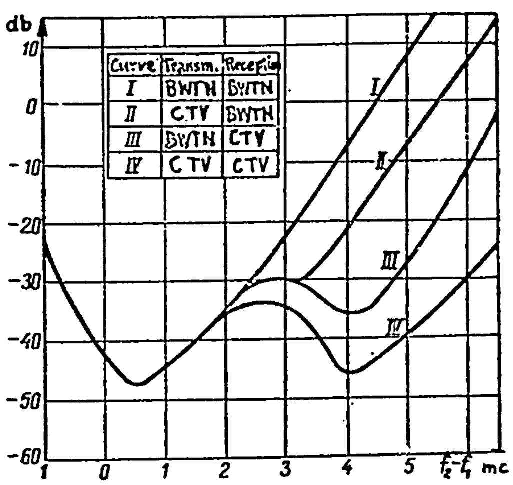

Fig.8

interference ratios remain the same for color and for black-and-white television. At frequencies close to that of the picture carrier, a 45-db interference ratio is necessary. Since the indicated ratio must be ensured by the existing black-and-white television stations, it may be considered that the signals of black-and-white television stations will not impose distortions on the color television stations which operate on the same channel. At the same time, it is indicated that this is valid for stations which operate with carrier phase shifts at an interference ratio of 30 db.

Distortions due to Color Television Stations. In such a case, in addition to the foregoing, it is necessary to determine which distortions of color signals may be expected from similar signals of the interfering station. In Fig.8 (curve IV), it can be seen that the interfering signal which falls into the color subcarrier region must be 45 db below the picture carrier of the received station. If only the distortions due to color signals of the interfering station which are 12 db below the picture carrier are considered, interference ratio of 45 db - 12 db = 33 db between the two color signals will be sufficient when the stations operate on a common frequency channel. In such a case, it is claimed that, since an interference ratio of 46 db must be maintained between the picture carriers, no mutual distortion must take place between the stations.

At the present time, black-and-white television stations operate with carrier phase shifts at a protection of 30 db. Color television, as shown above, requires a protection of 33 db in the subcarrier region. Due to the fact that the color subcarriers are also shifted, the actual interference ratio required will be less than 30 db; for this reason, it is assumed that no distortions will take place in such a case.

The measurements carried out have also shown that no distortions of color television, due to stations operating on adjacent channels, are expected.

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