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HEINRICH HERTZ

On the Centenary of his Birth

(1857 - 1957)

by

G.A.Levin

A hundred years have passed since the birth of Heinrich Rudolf Hertz.

"Heinrich Hertz" - was the content of the first radio message in the history of mankind, transmitted by A.S.Popov in 1895.

The name of Hertz was used for denoting the unit of oscillation frequency, the most important characteristic of any oscillatory processes.

For what did Heinrich Hertz merit his immortality?

In the Eighteen Seventies of the last century, advocates and opponents of the theory of action-at-a-distance in the science of electricity and magnetism were engaged in stubborn controversy.

On the theoretical plane, James Clark Maxwell struck the most serious blow at the remote action theory, having created his own amazingly orderly theory of electromagnetic phenomena, from which it followed of necessity that no kind of action-at-a-distance exists and that electromagnetic energy is propagated in wave form with a velocity equal to the velocity of light. However, so long as the new theory was not confirmed experimentally, abundant opportunities remained for every kind of skeptical judgements about Maxwell's work.

Maxwell himself did not make such an experimental verification. To do this fell to the lot of Heinrich Hertz.

By his famous experiments in 1888 Hertz proved indisputably the existence of electromagnetic waves with finite velocity of propagation in space. It was further found that this velocity was equal to the velocity of light and that in general elec-

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tromagnetic waves are, in their properties, almost completely identical to light waves. They are reflected from conductors, are refracted during transition from one dielectric to another, are polarized, and are capable to produce interference phenomena.

The experimental results of Hertz became instantly sensational. The idealistic theory of action-at-a-distance suffered a crushing defeat. Simultaneously, the electromagnetic nature of light became obvious. The services of Hertz to radio technology are, however, not exhausted by this.

For his experiments, Hertz used equipment which in a large measure anticipated the apparatus of the simplest radio stations.

This equipment comprised an open oscillatory loop, the prototype of future antennas; a generator of damped oscillations of high frequency, which greatly resembles the oscillator of future spark-radio stations; a receiving device in which a spark served as the indicator of the presence of oscillations.

Thus, without setting for himself the task of accomplishing radio communication, and evidently not even suspecting such a possibility, Hertz put into our hands something very much like a radio station, in which the primitive open loop was to be replaced by a real antenna, and a more modern indicator of the presence of oscillations was to be employed in the receiver instead of a spark.

The greatest service rendered by A.S.Popov in this respect was that he was first to perceive how close Hertz, without himself being aware of the fact, was to a solution of the problem of radio communication and along what path the research by Hertz should be continued in order to realize radio communication.

In his experiments, Hertz made extensive use of the phenomenon of resonance, so characteristic for all radio technology. Hertz was a great theorizer. He is the originator of the theory of the so-called Hertz dipole, in which the problem of emission of radio waves is solved by an open oscillator of the simplest form. To the present time, this theory is the basis on which any theory of the emission of radio

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waves by antennas is built.

From the modern viewpoint, Hertz's theory of emission should be closely examined also as to possible linking of the emission phenomenon with the behavior of electrons in the atom. The state of physics in Hertz's times, however, excluded the possibility of such development of the emission theory by Hertz himself. This proved possible only at a later time, on the basis of quantum mechanics.

Hertz was the first in the history of physics to investigate the photoelectric effect. In the last years of his life, Hertz studied cathode rays. Early death - Hertz died in 1894 at the age of 37 years - cut short the fruitful scientific activity of this gifted man.

In its development, radio technology has repeatedly deviated greatly from the ideas of Hertz. The wave band which Hertz used for his experiments (ultrashort and decimeter waves) proved of little use for radio communications over long distances. At first, therefore, the radio technology of long waves saw rapid development.

The invention of the cathode-ray tube produced an opportunity to solve many such problems of which Hertz could not even dream. The development of radio broadcasting entailed full mastery of the medium-wave band. Extreme-distance radio communication was reliably realized by the use of short waves, with their special mechanism of propagation.

In recent decades, however, radio technology has again returned to the Hertzian wave band. The development of television, radar, and radio-relay lines of communication required utilization of ultrashort decimeter and centimeter waves.

The transmission and reception equipment for waves of superhigh frequencies, of course, has so far developed in the meantime in comparison with the equipment of Hertz, that there cannot be even the remotest quantitative or qualitative comparison. Certain elements of Hertzian equipment, however, have been retained to the present time. These include the Hertz doublet itself, parabolic reflectors, and so forth.

The point is not at all that contemporary radio technology has, in certain re-

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spects, gone back to Hertz.

Important is the tremendous impetus which the work of Hertz as a whole gave later generations of scientists and mainly to A.S.Popov, the creator of radio technology of wireless communication. From here began the rapid scientific progress which led to present-day radio technology with its astonishing achievements.

Hertz is indisputably one of the classic leaders of physics in the field of the science of electromagnetic oscillations. In this consists his great role in the history of science.

In the days of celebrating the centenary of the birth of Heinrich Rudolf Hertz, we remember with gratitude the name of the great scientist and render due respects to his remarkable creative work in science.

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A SCHEMATIC FOR RECEPTION OF SIGNALS

by

A.A.Kharkevich

One of the possibilities of "total" reception of a binary signal is discussed, with a description of the basic diagram of a device which accomplishes this possibility. The reception is done visually, by observation of the screen of a cathode-ray tube.

To realize the possibilities offered by interference-killing codes, it is necessary to receive the signal (in the form of some code combination) as a whole, i.e., the entire combination in toto, and then to collate it with a great number of transmitted signals and identify it with those from which the received signal differs least. This permits detecting and correcting of errors.

From the viewpoint of the geometric theory, any signal is a point in space of a corresponding number of measurements. The superimposition of interference displaces this point. To avoid errors, i.e., to identify the received signal not with the actually transmitted signal but with other possible signals, it is necessary to increase the distance between points that represent possible transmitted signals.

The technology of receiving the signal in its "entirety" must in general consist of the following: 1) the received signal is remembered; 2) the signal is checked with all possible transmitted signals which must be known in the reception side of the system of communications and must be stored in a certain memory device; 3) of the possible transmitted signals the one from which the received signal differs least is recorded - this signal is also considered as the transmitted signal (the expression "differs least" determines the operating system of the receiver).

The general wiring diagram of a receiver of "total" signals is pictured in the following manner: There is a memory in the form, for example, of magnetic recording,

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in which all possible transmitted signals are recorded in advance; there also is the possibility of separately recording the received signal. After that the received signal is reproduced many times; simultaneously, each time one of the possible transmitted signals is reproduced.

The comparison can be done in various ways; either by subtracting, squaring the difference, and summarizing according to the formula

$$d_k^2 = \int_0^T [y(t) - x_k(t)]^2 \, dt, \tag{1}$$

where $y(t)$ is the received signal;

$x_k(t)$ is one of the possible transmitted signals;

and the transmitted signal is determined by the minimum d_k;

or by multiplying the signals and then summarizing according to the formula

$$R = \int_0^T y(t) x_k(t) \, dt \tag{2}$$

and finally determining the transmitted signal according to the maximum R. Both variants of the method of checking are theoretically equivalent*. The quantity d_k in the geometric sense is the distance between the received signal and any possible transmitted signal. The quantity R is the factor of cross correlation between the received signal and any possible transmitted signal. Equations (1) and (2) refer to continuous signals; T designates the duration of the signals. For discrete signals, eqs.(1) and (2) are replaced by corresponding sums:

$$d_k^2 = \sum_{i=1}^{n} (y_i - x_{ik})^2, \tag{1'}$$

$$R = \sum_{i=1}^{n} y_i x_{ik}, \tag{2'}$$

*For signals of equal energy both variants simply coincide.

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where y_i, x_{ik} are the discrete values ("signs", symbols), respectively, of the received signal and any of the possible transmitted signals, while n is the number of signs.

The above-described system has a sufficiently universal character and is applicable for any signals. But the resulting technical solution is rather cumbersome. In addition, the operation of checking and selection requires time; this time increases rapidly with an increase in the duration of signals, since an increase in the duration (or the number of signs, in the case of discrete signals) causes the total number of possible signals to increase according to an exponential law. The necessary capacity of the memory also grows correspondingly. Meanwhile the advantages of interference-killing codes are realized precisely for large segments of signal. The most advantageous correlations are derived in the range when T (or n) tends toward infinity.

Consequently, such a universal system is not of such interest as a simpler system for reception of "total" signals. We will describe this circuit applicably to the reception of discrete and, in particular, to binary signals.

The idea consists of transfering the multidimensional image of a set of transmitted signals to a plane, i.e., to a space of two dimensions. This permits using an ordinary cathode-ray tube as the terminal link of the receiving device; the set of transmitted signals is presented as a system of points arranged in a certain plane lattice.

Any n-digit binary number of the total quantity

$$N = 2^n \qquad (3)$$

can be replaced by a two-digit number according to a number system with the base a. For complete mapping of the set N, it is necessary only to satisfy the condition

$$M = a^2 \geqslant N, \qquad (4)$$

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from where also a is determined. On the basis of eqs.(3) and (4) we find

n	3	4	5	6	7	8	9	...
a	3	4	6	8	12	16	23	...

For the start, we will analyze the case of n = 3, a = 3. The three-digit binary signal is geometrically represented by the vertices of a cube; the set of transmitted signals contains $N = 2^3 = 8$ combinations.

Each combination is expressed by a three-digit binary number. On the other hand, these binary numbers can be expressed by two-digit ternary numbers; one is found in excess, since

$$M = 3^2 = 9.$$

Thus, we have the following one-to-one correspondence:

Number of Combinations (vertices of cube)	0	1	2	3	4	5	6	7
Binary Recording	000	001	010	011	100	101	110	111
Ternary Recording	00	01	02	10	11	12	20	21

Pictured in Fig.1a is a cube with correspondingly numbered vertices, and in Fig.1b, a two-dimensional table for ternary recording.

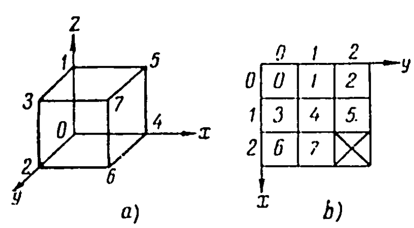

Fig.1

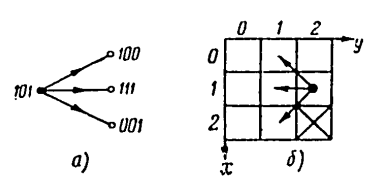

Fig.2

Let us assume that the binary number 101 is converted to the ternary number 12 (problems of the technology are discussed below). We consider this number as a point on a plane with the coordinates x = 1, y = 2. This point coincides with the lattice cell for No.5 and depicts the vertex No.5 of the three-dimensional cube (with the coordinates x = 1, y = 0, z = 1).

It is clear that, if during the transmission of the binary signal a single error occurred (i.e. an error in one sign), then the received combination coincides with one of the other possible combinations. For example, a single error may convert signal No.5 into one of three others, according to the diagram in Fig.2a, which corresponds to the situation marked by points in Fig.2b. Consequently, an error under such conditions cannot be corrected or detected.

In order to detect a single error, it is obviously sufficient to select binary combinations so that they differ by not less than two signs. In a geometric model in the form of a three-dimensional cube, this corresponds to a selection of vertices spaced at a distance of two edges (i.e. lying on diagonals of the faces). This means that we use only half of all possible combinations; the remaining combinations are forbidden. Let the combinations 000, 011, 101, 110 be allowed, corresponding to the

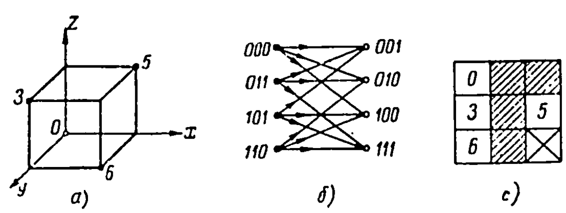

Fig.3

cube vertices Nos.0,3,5,6 (Fig.3a). A single error transforms the allowed combinations into forbidden ones according to the diagram in Fig.3b. In the two-dimensional Table in Fig.3c, the sites of the forbidden combinations are hatched; the incidence of the point of the received signal on one of the hatched cells indicates an error, which is thus automatically detected. A binary error cannot be detected since it translates any allowed combination into another allowed one.

In order that a single error can be corrected, the allowed combinations must differ in not less than three signs. We can construct only two three-digit binary signals that satisfy this condition. The corresponding points are lying in the diagonals of the cube (for example, 000 and 111, as shown in Fig.4a). The transition circuit in a single error is shown in Fig.4b. Thus each of the forbidden combinations formed as a result of a single mistake, is connected only with one of the two allowed combinations. Therefore, the two-dimensional Table assumes the form of Fig.4c. The base positions (in the absence of an error) of signals Nos.0 and 7 are marked by circles; these numbers are repeated in other cells to which the signals might be transmitted as a result of a single error. Thus, a single error is corrected.

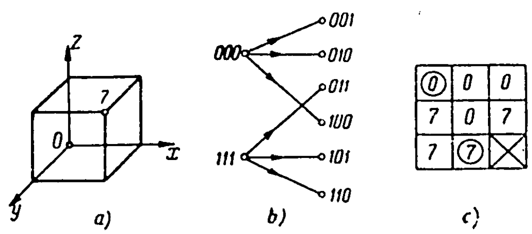

Fig.4

We are interested, however, in a larger set of signals requiring combinations with a great number of signs during binary recording. As an example, let us take the telegraph code used for the transmission of 32 letters. The ordinary code is five-digit (Baudot code), but we at once construct the simplest code detecting a single error, after supplementing the five-digit combination by another 0 or 1, with a calculation such that the derived six-digit combinations have an even number of units (or zeros). For conversion of the six-dimensional space to a plane we employ an octonary number system. Besides,

$$N = 2^6 = M = 8^2 = 64.$$

In the Table below are given the letters, their binary recording in the form of six-digit combinations developed from the ordinary Baudot code, and the corresponding two-digit octonary recording.

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The flat Table corresponding to the last line has 8 × 8 squares like an ordinary chess board. Half of the squares are occupied by letters, while the remaining half corresponds to forbidden combinations (with an uneven number of zeros or units); these squares are hatched (see Fig.5). In a similar manner, the binary code which

Letter	A	B	C	D	E	F	G
Binary Recording	100001	001100	011011	010100	111100	010001	000110
Octonary Recording	41	14	33	24	74	21	06

corrects the single error can also be constructed. This will be a nine-digit code (Bibl.1); a Table of 23 × 23 = 529 squares will serve as its two-dimensional mapping.

The basic diagram shown in Fig.6 is useful for a technical realization of the reception device, with reflection of signals in a plane. The binary signal, received together with the superposed interference, enters the quantizing device Kv which directs the received signal to one of two levels and determines what, actually, is received: 0 or 1. Here also occurs a possible error whose probability is estimated in the usual manner if the probability distribution of the interference is known. From the quantizing device,

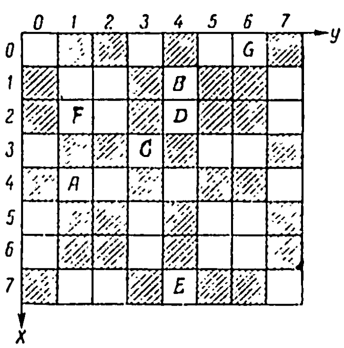

Fig.5

the binary signal goes to the coincidence circuit CC; here also enter the pulses from the pulse generator IG, operating in synchronism with the signal (which can be started with starting pulses). At the output of CC a binary signal of standard level is obtained. This signal goes to one of two decoding devices D_1 or D_2, depending on the position of the switch K. This latter is controlled from the pulse counter Co.

The action of the system consists in that, for example, in the six-digit code the first set of three binary signs determines the first figure of the two-digit number. When this has been determined, the switch is reset, and the next set of three binary signs, arriving at the second decoding device, determine the second figure of

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the two-digit number. The decoding devices are conventional decoding circuits KIM, containing an RC segment with a time constant, selected so that during the cycle period the potential in the capacitor is smoothly cut in two. This principle is

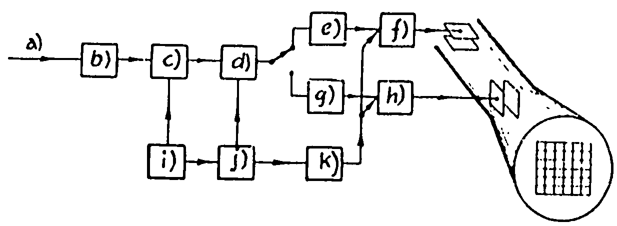

Fig.6

a) Signal plus interference; b) Quantizing device Kv; c) Coincidence circuit CC; d) Switch K; e) Decoder D_1; f) Storage cell N_1; g) Decoder D_2; h) Storage cell N_2; i) Pulse generator IG; j) Pulse counter Co; k) Dropping circuit Cd

widely known. At the output of the decoding devices the storage cells N are connected, which maintain the definitive value of the potentials resulting at the end of the operating cycle of the decoding devices. From the storage cells the voltage is fed directly to two pairs of deflecting plates. By means of the dropping circuit Cd, the voltage is taken from the storage cells at the required instant which latter is likewise set by the pulse counter. The screen of the tube is superposed by a transparency in the form of a grid (like Fig.5).

The above description is a tentative outline of an apparatus conceived primarily as a demonstration unit, but also useful for certain investigations. The future will show whether these or other elements of a similar device can find application in the technology of communications.

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IMPROVEMENT OF THE METHODS OF MEASURING NONLINEAR DISTORTIONS

by

N.L.Bezladnov

Means are examined for obtaining the greatest correspondence between the results of measurement and the degree of subjective perception of nonlinear distortion. For this purpose, measurements are recommended of relative distortion levels realized during the reproduction of a sound transmission or when using test spectra which simulate the transmission. It is proposed that, during the measurements, corrections in the distortion volume be taken into account and also their possible masking by signals of the transmission and noises in the reproduction rooms. The principles of constructing the corresponding measuring circuits are examined.

Introduction

At present, measurements of the nonlinearity of the reproducing circuit and measurements of nonlinear distortions introduced by it are in use.

One of the main shortcomings of the existing methods of measuring nonlinear distortions in sound-reproduction devices is the discrepancy of the results of measuring the degree of subjective perception of distortions. This shortcoming in fact lessens the value of the measurement. To overcome it, the principles used, as basis for distortion measurement must be established, proceeding from factors which determine the subjective perception of distortions. Such factors are chiefly the distortion volume and their masking by signals of the transmission and also by noises in the rooms where the reproduction is made. In turn, these factors are determined by the character of the nonlinearity of the devices being examined and also by the character of the sound transmission being reproduced (distribution of the dynamic levels of transmission in time and according to frequency).

It should be especially emphasized that, inasmuch as the point under discussion is not an investigation of the nonlinear characteristics of the reproducing devices but of the nonlinear distortions introduced by them, the character of the reproduced transmission acquires decisive importance, and failure to take this circumstance into account inevitably leads to substantial errors.

Basic Principles of Measurement of Nonlinear Distortions

In accordance with the foregoing, the following principles must be used as basis of measurements:

1. The measurements must be made in the process of reproducing the sound transmission (dynamic conditions) and during the admission, at the input of the device under study, of specially selected test spectra which simulate definite fragments of sound transmission. For purposes of simplifying the measurements, discrete frequency spectra can be used, which correspond to periodic oscillations of constant magnitude (static conditions).

2. For a numerical characteristic of the distortions, it is expedient to measure the relative level of the distortion volume. For this, it is sufficient to measure the relative level of the power of nonlinear distortions and to introduce corrections in the curves of equal volume, carried out in accordance with the normal level of reproduction (for example, applicably to reproductions with a maximum loudness level +90 decibels). The expressions for the relative levels of the distortion power can be represented in the following form:

$$R_d = 10\lg\frac{\int_{\omega_{i1}}^{\omega_{i2}} p_i(\omega)\,d\omega}{\int_{\omega_1}^{\omega_2} p(\omega)\,d\omega} = 20\lg\frac{\sqrt{\int_{\omega_{i1}}^{\omega_{i2}} u_i^2(\omega)\,d\omega}}{\sqrt{\int_{\omega_1}^{\omega_2} u^2(\omega)\,d\omega}},$$

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$$R = 10\,\lg \frac{\sum_{1}^{N_i} P_i}{\sum_{1}^{N} P} = 20\,\lg \frac{\sqrt{\sum_{1}^{N_i} U_i^2}}{\sqrt{\sum_{1}^{N} U^2}}.$$

Here R_d is the dynamic relative level of the distortion power;

R is the static relative level of the distortion power;

P_i, U_i are the effective values of the power and voltage of the products of nonlinear distortions;

$p_i(\omega)$, $u_i(\omega)$ are the continuous frequency spectra of the power and voltage of the products of nonlinear distortions;

P, U, $p(\omega)$, $u(\omega)$ } is the same for the fundamental oscillations;

ω_{i1}, ω_{i2} are the edges of the spectra of distortions;

ω_1, ω_2 are the edges of the spectra of fundamental oscillations.

The transformation to the expressions for the dynamic and static level of the distortion volume R_d' and R' can be made by the introduction, respectively, of the correction functions $m_i(\omega) = 10^{\frac{b_i(\omega)}{10}}$, $m(\omega) = 10^{\frac{b(\omega)}{10}}$ or of the correction factors $m_i = 10^{\frac{b_i}{10}}$, $m = 10^{\frac{b}{10}}$. Here $b_i(\omega)$, $b(\omega)$, b_i, b are the differences between the intensity levels and the volume levels for corresponding frequencies (determined according to curves of equal volume). Thus the dynamic relative level of the distortion volume is

$$R_d' = 10\,\lg \frac{\int_{\omega_{i1}}^{\omega_{i2}} p_i(\omega)\, m_i(\omega)\, d\omega}{\int_{\omega_1}^{\omega_2} p(\omega)\, m(\omega)\, d\omega},$$

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while the static relative level of the distortion volume is

$$R' = 10 \lg \frac{\sum_{1}^{N_i} P_i m_i}{\sum_{1} P m}$$

It is evident from the foregoing that, in the measuring circuit, the change-over to the levels of distortion volume can be made by supplementing the zones of distortions and transmission with attenuations depending on the power levels in these zones and on the frequency.

3. In view of the fact that, in the process of reproducing a sound transmission, the level of nonlinear distortions varies continuously it is useful, for the case of a full-value characteristic of the distortions, to determine the instantaneous and mean value of the dynamic level of the volume of nonlinear distortions. By instantaneous value is meant the distortion level measured by an instrument with a time of integration, responding to the minimum signal duration sufficient for perception of nonlinear distortions (tens of milliseconds). Of greatest interest is the measurement of the instantaneous dynamic distortion levels, which correspond to the maximum level of transmission. This permits a more reliable control of the limit of the allowable use of the equipment than by measurement of the maximum dynamic voltage levels. Actually in the latter case only an indirect judgement as to the magnitude of nonlinear distortions is possible. However, the rare and transitory increases of distortion at transmission peaks cannot have a substantial effect on the quality of reproduction as a whole, and in this respect the mean value of the dynamic distortion level $R_{d\ mean}$ is more characteristic. This means the average statistical value of the distortion level, determined on the basis of a distribution in time of the dynamic levels of sound transmission.

4. For augmenting the correspondence between the measurements and the subjec-

tive perception of distortions, the masking of distortions by the transmission signals and also by noises in the reproduction rooms must be taken into account.

For the most accurate determination of the masking of distortions, the frequency range of sound reproduction must be divided into a series of frequency bands and the level of perception in each of the bands must be reduced by the magnitude of the masking. The latter is determined with consideration of the effect of the masking spectrum within the limits of the given band and of all underlying frequency bands. As is generally known, the portion of the spectra which corresponds to the overlying frequencies has no practical effect on the masking.

A well-known method of rating the masking of spectra by spectrum (Bibl.1) consists in finding, for each of the bands, the equivalent level of perception of the masking spectrum, which also determines the masking in the given band. This level is equal to the sum

$$B_{n1}(+)B_{n2}(+)\dots(+)B_{nK}(+)\dots(+)B_{n-1,n}(+)B_{nn},$$

where $B_{nK} = B_K - Q_{nK}$, B_K being the integral level of perception of the masking spectrum within the limits of the given band, while Q_{nK} is the constant interband attenuation which characterizes the reduction of the masking effect in the band n of the integral level in the band K, in comparison with its effect in the same band K. The signs (+) signify that, rather than the levels B_{nK}, the quantities $10^{\frac{B_{nK}}{10}}$ are summarized, which are proportional to the power of the masking oscillations.

The rating cited for the case of the masking of distortions by transmission signals can be modeled in the circuit by supplying the zone, divided into distortion frequency bands (depending on the transmission levels) with attenuations equal to the masking in the given band and lowering the level of perception for the masking spectrum of distortions in this band by its magnitude. The indicated attenuations must be determined by the summary power of the transmission spectrum for the given and underlying frequency bands. At the same time, the action of transmission powers of the

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underlying frequency bands must be obtained through constant interband attenuations (a_{nk}).

Such a circuit is obviously fairly complex. However at some lowering of the recording accuracy of masking, considerable simplifications are apparently possible. Thus, the number of frequency bands can be reduced, the effect of the masking levels need be considered only in the given band and in one of two closest underlying frequency bands (or even only in the given frequency band), and finally, instead of introducing a smoothly varying attenuation, the blanking of the frequency channels of the distortion zone can be accomplished when sufficiently high transmission levels are present in the corresponding frequency band.

The rating of the masking action of noises in the reproduction rooms can be realized in a similar manner, with the difference that all introduced attenuations have a constant magnitude.

Principles of Designing Test Circuits

The principal problem in the development of test circuits is the selection of means for isolating the distortion products at the output of the device being investigated.

It is obvious that the method of filtration customarily used with multifrequency oscillations, such as a sound transmission or the test spectra simulating it are, is not always applicable inasmuch as the distortion spectra and the fundamental frequency spectra may considerably overlap one another.

In the latter case, the following means of isolating distortion spectra are possible: a) compensation of the fundamental frequencies in the test channel at the output of the device being investigated; b) blanking of the test channel for fundamental frequencies at the output of the device being investigated.

The compensation method is apparently the only method which permits isolating the entire effective distortion spectrum, which is formed under dynamic conditions, without any variation and without lowering the quality of the transmission reproduction. In principle, the compensation method can also be applied in static measurements. The compensation channel (Fig.1), within the limits of the reproduced frequency range, must have the same transmission factor $\frac{U_{out}}{U_{in}}$ as does the fundamental channel. Since, under the examined conditions, the transmission factor uniquely determines the transition function, the compensation conditions will be retained also under dynamic conditions. For a control of the transmission factor of the compensation channel, a suitable extension arm and amplitude-frequency and phase-frequency correctors (AK and FK) must be provided. Compensation is realized in the circuits of the compensator K.

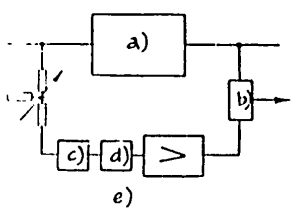

Fig.1

a) Device being measured; b) Compensator K; c) Amplitude-frequency corrector AK; d) Phase-frequency corrector FK; e) Compensation channel

It is difficult in practice to obtain an exact compensation within the limits of a wide frequency band. It is, therefore, expedient to have several compensation channels available, each of which accomplishes compensation within the limits of a relatively narrow frequency band. Such a structure of the compensation zone facilitates switching to the levels of distortion volume, and also permits an easier evaluation of the effect of masking.

Blanking the test channel for fundamental frequencies can be accomplished with a consecutive reproduction of the amplitudes of the frequency components at the output of the measuring device, by means of an analyzer with a sliding frequency of the type of a spectroscope. This method can be applied in static measurements.

Apart from the reviewed methods of isolating the distortion spectra in multifrequency oscillations, another useful method is that proposed by V.M.Vol'f (Bibl.2), permitting an isolation of the products of distortion within the limits of a narrow frequency band at the output of the examined device during cut-out of this band from the spectra of the input signal.

Dynamic Method of Measuring Nonlinear Distortions

This method is designed for measuring nonlinear distortions under dynamic conditions of reproduction of a sound transmission and is based on the compensation of fundamental frequencies in the measuring zone, which is divided by filters into a series of frequency channels. The method considers the correction in correlations of volumes, and also the effect of the masking of distortion spectra by the transmission spectrum.

The block diagram of the measuring unit is shown in Fig.2, where F_1 and F_1', F_2 and F_2', F_3 and F_3' are band-pass filters, AK_1, AK_2 and AK_3 are amplitude-frequency

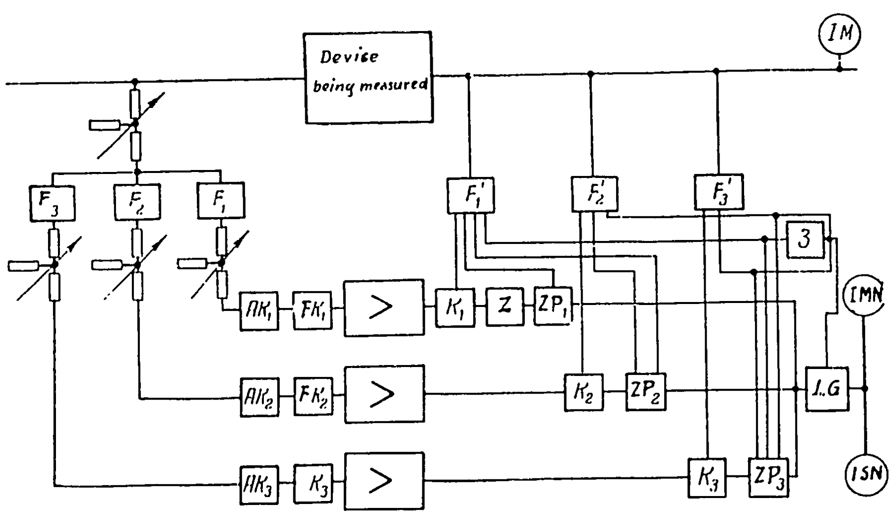

Fig.2

correctors, FK_1, FK_2 and FK_3 are phase-frequency correctors, K_1, K_2 and K_3 designate compensators, Z denote nonlinear loops for introducing attenuations in the low-frequency channels at low transmission levels, ZP_1, ZP_2 and ZP_3 represent blanking loops, LG is a logarithmic logometer (electronic), while IMN is a meter for maximum dynamic levels of nonlinear distortion volumes, ISN is a meter for the mean dynamic levels of nonlinear distortion volumes.

The filters of each pair (F_1 and F_1', F_2 and F_2', etc.) must have strictly identical damping characteristics and phase-frequency characteristics, in order that the amplitude and phase correctors AK and FK take into account only the corresponding distortions in the device being examined. At the output of the compensators K remain only the products of nonlinearity (and background) formed by the device being measured. The nonlinear loops Z, connected in the low-frequency channels of distortions and of fundamental frequencies, introduce, at low levels, attenuations which take into account corrections in the curves of equal volume. Estimates show that, under other conditions, corrections in volume can be disregarded. The effect of masking, in a simplified way, is taken into account by the loops ZP_1, ZP_2, and ZP_3 which blank corresponding frequency channels when sufficient levels of fundamental oscillations are present in these channels and in channels of lower frequencies.

The loops Z connected in the low-frequency channels effectively suppress the background harmonics of the device being examined (also during the pauses). The loops ZP suppress products of nonlinearity with frequencies that coincide with the fundamentals and therefore do not cause nonlinear distortions.

The blanking loops must be inertial to avoid additional nonlinear distortions. Their time parameters must be selected on the basis of a compromise between the requirement for a low magnitude of additional nonlinear distortions and sufficiently fast reaction with respect to recording the distortion dynamics. It should be mentioned that certain inevitable additional distortions will not greatly affect the accuracy of measurement inasmuch as the products of nonlinearity rather than the fundamental frequencies are subject to these secondary distortions.

As far as the amplifiers of the compensation channels are concerned, the nonlinear distortions introduced by them can easily be made negligibly small, since their output power will not exceed tenths of a watt, and since the input is fed only

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with the portion of the fundamental frequency spectra which corresponds to one of the frequency bands of the compensation zone.

The minimum number of frequency channels must be defined as that at which the recording of the masking is sufficiently accurate. At the same time, the required accuracy of compensation of the fundamental-frequency voltage must be taken into consideration.

The entire spectrum of the distortion products and also the fundamental-frequency spectrum are fed from the output of the device being measured to the electronic logarithmic logometer LG*. The latter gives a rectified current, proportional to the logarithm of the ratio of integral volumes of nonmasked products of distortions to the fundamental frequencies or, in other words, it measures the dynamic levels of the volume of nonmasked distortion products. The logarithmic logometer must also be of the inertia type. The time of integration is selected on the basis of the necessity of measuring instantaneous dynamic distortion levels.

As a meter for mean dynamic distortion levels (ISN), an automatic recording instrument giving a level-gram of distortions can be used or an ordinary electrodynamic meter**. It is desirable to use the meter for maximum dynamic voltage levels IMN.

In using the proposed circuit, the compensation may be disrupted not so much because of the nonlinearity of the device being examined but because of a variation in its amplitude and phase-frequency characteristics. It is, therefore, desirable to have a possibility of checking the transmission factor of the device not only during the intermissions but also during actual reproduction.

The described measuring unit can be used primarily for rating the single-valued

*More precisely, both the fundamental-frequency and the distortion spectra are present at the output of the device being measured; however, at relatively small distortions, this is known to have very little effect on the accuracy of measurement.

**The sufficient accuracy of the meter readings, when it was used for a similar purpose, was verified experimentally under the direction of the author.

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correspondence between the given subjective opinion of the expert and the numerical indices of nonlinear distortions.

In addition, a similar measuring unit, possibly somewhat simplified (reduced number of compensation frequency channels, maximally simplified recording of masking, etc.) can be employed in most cases for effective operating control of distortions. The presence of such a control permits judging any possible defects of the equipment and any disturbance of the fixed diagram of levels because of imperfections of operative and nonoperative adjustment of the transmission levels. At the same time, an instrument may be required for continuous tracking of any variations in the transmission factor and for automatic restoration of the balancing of the fundamental-frequency voltage.

The proposed dynamic method of measuring nonlinear distortions, in comparison with the method of V.M.Vol'f (Bibl.2), has the main advantage that the entire developing spectrum of nonlinear distortions is taken into account by it simultaneously, whereas the method of Vol'f requires the carrying out of a series of consecutive measurements for the various frequency bands and therefore cannot be used under dynamic conditions, without repetition of the reproduced transmission.

Static Method of Measuring Nonlinear Distortions

In the development, with industrial and episodic operational control of sound-reproducing equipment, speed and accuracy of measurements are essential under definite conditions stipulated by GOST (State Standards) or technical requirements. Moreover, the measuring equipment must be sufficiently simple to handle. It is therefore expedient to measure nonlinear distortions during the input of discrete frequency spectra, simulating the real oscillations of the transmission, i.e., to measure the static distortion levels.

In selecting such test spectra it is necessary to determine:

a) The number of frequencies entering the composition of the spectrum;

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b) The edges of the spectrum and the values of the separate frequencies forming it;

c) The voltage amplitudes of the frequency components of the spectrum and their distribution in frequency.

It is evident that measurements must take into account the conditions most unfavorable with respect to the formation of nonlinear distortions, and this corresponds to a transmission with a large number of frequencies, i.e. to a multivoice performance (orchestra, chorus). But even in speech transmission, the number of frequency components is considerable (measured in tens).

Therefore, with the object of maximum approximation to real conditions relative to the quantity, to the power of the products of nonlinear distortions being formed, and to the character of their distribution in the frequency spectrum, it is desirable to select a sufficiently large number of frequencies of the test spectrum. At a large number of the indicated frequencies, however, the resulting measuring unit would be extremely complex and cumbersome. To avoid this, it is necessary to determine the minimum number of frequencies of the test spectrum at which the spectrum of emerging distortions still sufficiently corresponds in character to the distortion spectrum, formed by real multifrequency oscillations.

The criteria of this correspondence can be selected in the following manner:

It is known (Bibl.3,4) that, during the action of multifrequency oscillations on a nonlinear quadripole, a spectrum is formed of nonlinear distortions, whose components are divided into primary and secondary products. The frequency of the distortion product ω_i has the form $\omega_i = a\omega_1 \pm b\omega_2 \pm c\omega_3 \pm \ldots z\omega_k \pm \ldots$, where ω_1, ω_2, ... ω_k, ... are frequencies of the components of the input potential (fundamental frequencies). The number of fundamental frequencies participating in the formation of the frequency of the distortion product ω_i, can be different and is determined both by the order and kind of product and by the total number of fundamental frequencies. To the primary products correspond coefficients with frequency components of a = b = c ... = z = ... = 1; in the case of secondary products at least one of the coefficients before the frequencies ν_1, ν_2, ν_3, ... ν_k ..., which form the frequency of the primary product of the same order, is greater than unity, and at least one of them is equal to zero (the harmonics of the fundamental frequencies are a particular case of secondary distortion products). It can be shown that during an unlimited increase in the number of fundamental frequencies, the summary power of the secondary products tends toward zero. Thus, the characteristic criterion of the distortion spectrum of multifrequency oscillations is the concentration of the distortion power in the primary products of nonlinearity.

By this is determined the means of forming the frequencies of the distortion products (frequencies of the type $\nu_1 \pm \nu_2 \pm \ldots \pm \nu_n$) and consequently their frequency spectrum, and also the correlation with the fundamental frequencies. By this, at a given characteristic of nonlinearity and at given amplitudes of the fundamental frequencies, is also determined the character of the distribution of the power of the distortion products in the frequency range. Therefore, in selecting the number of frequencies n of the test spectrum, the main criterion must be the ratio of the power of the primary distortion products to the power of seconproducts. It is essential to note that this ratio does not depend on the character of nonlinearity of the system.

Order of Products of Distortions	Ratio of the Power of Primary Products of Distortions to the Power of Secondary Products — Number of Fundamental Frequencies	
	n = 5	n = 10
2	16.0	36.0
3	7.8	21.0
4	3.38	13.1
5	1.16	8.3

Calculations made under condition of equality of the voltage amplitudes of the fundamental frequencies led to the results shown in the Table.

It can thus be assumed that, already at n = 10, the overwhelming portion of the power of nonlinear distortions is concentrated in the primary products of nonlinearity and, consequently, the distortion spectrum is a spectrum characteristic for multifrequency oscillations.

In selecting the edges of the test-frequency spectrum one can proceed from: 1) filling in with frequencies the entire operating frequency range (or its greater part); 2) filling in with frequencies only a certain section of the operating range.

The first alternative has the advantage that it permits simultaneously taking into account the effect of nonlinearity over the entire frequency range.

The second alternative, in which the high-frequency section of the range is filled in with frequencies and the section of lower frequencies in which masking of distortions has no effect is left unfilled, will probably permit taking into account the case of greatest differentiability of distortions and, as is shown below, leads to a simpler circuit solution. As far as the voltage amplitudes of the frequency components of the test spectrum are concerned, they can be identical or variable in accordance with the probable distribution of the levels of the frequency transmission.

The concrete selection of the edges of the test spectrum, of its frequency components, and their amplitudes requires an accumulation of pertinent experimental data.

It should be noted that, knowing the dependence $R = \varphi(b)$, where b is the level of the input potential, one can judge the average dynamic level of distortions. For example, a frequently encountered form is the dependence $R_d = R_{d\ max} + \frac{a}{2}b$, where $b = 20\ \lg \frac{U_{in}}{U_{in\ max}}$, $R_{d\ max}$ corresponds to $b = 0$, and a is a constant positive magnitude. It can be shown that the average dynamic distortion level is also equal to the static distortion level when the dynamic transmission level is average, i.e., at the level $b_{d\ mean}$, corresponding to the mean of its dynamic range ($b_{d\ mean} = -0.5B_d$).

Circuits for measuring the static levels of nonlinear distortions can be designed with application of compensation means, filtration, or blanking of the fundamental-frequency voltages at the output of the measuring unit.

A shortcoming of the compensation circuit in static measurements is the complexity of its tuning (compensation realized in several frequency channels).

If such equipment must be employed for measuring various devices, then the overregulation of compensation channels may greatly complicate the operation.

In the case of a partial filling of the frequency range of the measuring unit with the test spectrum, the filtration means for fundamental frequencies can be used, which makes it possible to realize an extremely simple measuring circuit (Fig.3).

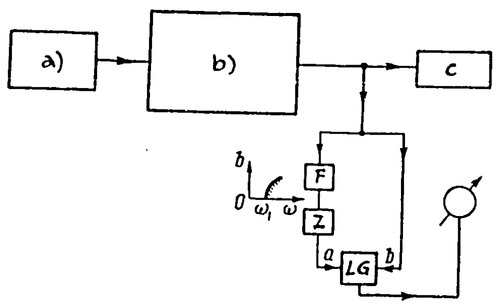

Fig.3

a) Oscillator of test spectrum; b) Measuring unit; c) Load

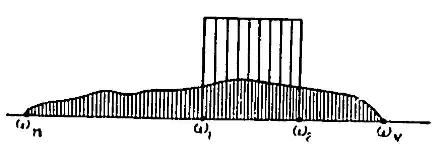

Fig.4

In this circuit, the low-frequency filter F filters out the spectrum of test frequencies $\omega_1 - \omega_2$, together with the products of nonlinear distortions which are in the frequency band of $\omega_1 - \omega_v$ and are masked by this spectrum (Fig.4). Therefore, at the input a of the logarithmic logometer LG only the nonmasked products of distortions, grouped in the band $\omega_k - \omega_1$, are admitted. The input b of the logometer is supplied with the entire spectrum of fundamental frequencies and distortions (which, as is known, does not lead to substantial errors).

In case it is not impossible to assume that the distortion products in the frequency band $\omega_2 - \omega_v$ are masked by the fundamental-frequency spectrum, the low-frequency filter must be replaced by a band-rejection filter in the frequency band $\omega_1 - \omega_2$. The nonlinear inertial loop Z introduces attenuation which takes into account the correction in volume necessary in the channel of distortions in view of the low frequencies and the low levels of nonmasked products of distortions.

A further improvement of the static methods of measurement can be obtained by

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providing the possibility of visual observation of the distortion spectrum. For this purpose, a spectroscope (Bibl.5) can be used, whose circuit structure is shown in a dotted frame in Fig.5. With partial filling of the range by the test spectrum,

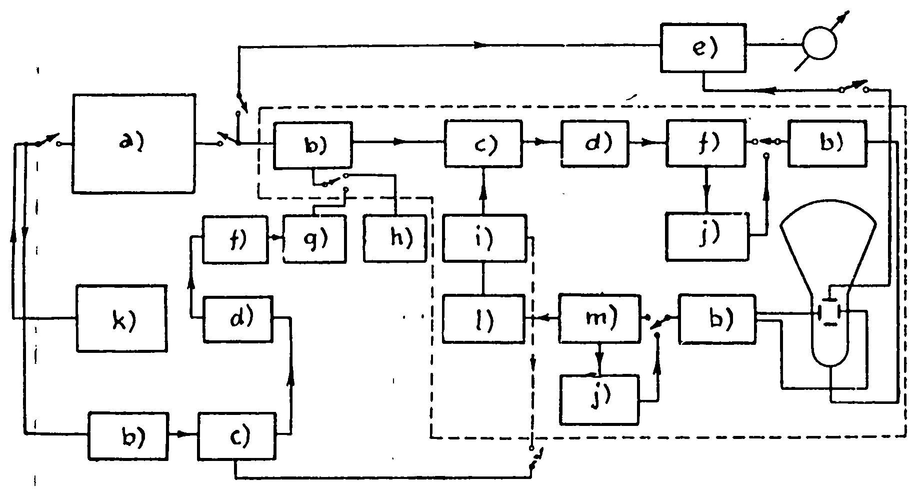

Fig.5

a) Measuring unit; b) Amplifier; c) Mixer; d) Frequency-band amplifier; e) Logarithmic logometer; f) Detector; g) Circuit trigger; h) Multivibrator; i) Auxiliary oscillator; j) Logarithmic device; k) Oscillator of test spectrum; l) Frequency modulator; m) Deflection oscillator

exclusion of the fundamental frequencies can be obtained simply by restricting the limits of frequency variation by using a deflection oscillator. At the same time, the indicated limits must correspond to the frequency band occupied by the spectrum of the nonmasked distortion products (for example, the frequency band $\omega_n - \omega_1$, Fig.4). With solid filling of the range by test spectra, a trigger circuit can be used, fed by the test-spectrum oscillator through an auxiliary reproduction channel, and blanking the primary reproduction channel of the spectroscope for all frequencies of this spectrum. In view of the large number of distortion products, another means is also possible, based on the application of a device of the multivibrator

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type, which periodically unblanks the reproduction channel for short intervals of time, sufficient for operation of the spectroscope, through intervals of the order of 100 cps. The mentioned multivibrator can also be used for blanking the reproduction channel relative to the fundamental frequencies. In this case, it is essential that the frequency intervals of unblanking the reproduction channel contain no fundamental frequencies.

It should be noted that, in static measurements, the method by V.M.Vol'f (Bibl.2) can also be used; the method is improved in that the measurements of the distortion products in separate narrow frequency bands are made in sequence and automatically with sufficiently high speed.

In conclusion I consider it my duty to express gratitude to V.M.Vol'f and to V.K.Iofe for valuable counsels.

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THE STATE OF COLOR TELEVISION ABROAD

by

A.K.Kustaryev

A survey is given of the state of color television in the USA, France, England, and Holland on the basis of data collected during a visit to these countries by a group of CCIR delegates.

In March - April 1956, about a hundred delegates of the Eleventh Study Commission of the CCIR, including six delegates of the USSR, traveled to the USA, France, England and Holland with the object of studying color television. Demonstrations were arranged for the delegates, showing the operation of actual and projected systems of color television; also shown were various experiments for substantiating the principles of these systems and experiments in measuring the quality of the color picture, accompanied by reports, etc.

Color television in these countries is in various stages of development: in the USA for example, a system of color television known as NTSC system has been standardized and is in operation for more than two years; but in the remaining countries experimental work is still in progress in the study of original systems of color television or in the adaptation of the NTSC system to the television standards in effect in these countries. Since the USA at present holds a leading position in the field of color television and is the only country in which color TV broadcasting has been realized, the visit to the USA and an inspection of the NTSC system and various equipment for its operation were of special interest.

USA

At the present time black-white television is most developed in the USA. At the time the CCIR delegation visited the USA, the operating television stations numbered 475 and the black-white TV receivers about 40 million. Actually, all the most important regions of the country are provided with several programs of TV broadcast and about 75% of all families have TV sets. Television programs can be transmitted over the entire country by a network of radio-relay lines. The network of inter-urban lines comprises 109,000 channel-km and the network of local lines, 68,000 channel-km.

Color television as yet plays a still negligible role, but it is thought that, because of the imminent saturation of the country with black-white TV sets, the share of color television in the total volume of TV set sales will increase rapidly. From 1954 to 1955, the sale of color TV sets increased from 5000 to 35,000 units. It was expected that, in 1956, the sales will amount to 150,000 units and by 1964 may reach 9 million units. An obstacle to wide installation of color television was the high cost of color receivers. A color receiver with a 53-cm screen costs $700 and higher, whereas black-white receivers cost from $100 to $300. As is known, the most important part of a color receiver is the tricolor picture tube. These tubes are still extremely complex and expensive, but there is a trend toward reduction in their price as a consequence of improvement in the technology of their manufacture and the adjusting of mass production. Thus, from the beginning of 1955 to the present time the price of a tricolor RCA kinescope with shadow mask was reduced from $175 to $85. It is thought that in the near future the prices of color receivers can be expected to drop to $500 and even to $300, which will boost sales.

For the production of color programs, 23 studios (in March 1956) have been equipped and 70 more have been supplied with equipment for the transmission of color motion pictures or slides. It is planned that an additional 19 new stations will be equipped by 1956 for color programs. The production of color programs is as yet still negligible. Thus, in the NBC network in 1955 only 40 hours of color programs a month were transmitted. It was planned to double this figure by 1956.

Adopted as the standard system of color television in USA is the NTSC system,

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being a simultaneous compatible system, with transmission of color information in one color subcarrier with interlacing of spectra of brightness and color signals. A brief description of the basic principles of the NTSC system was given in the journal "Electrosvyaz'" for June 1956 (Bibl.4); we will therefore not describe it here. The NTSC system was the result of three years of work by the 30 largest American radio-engineering firms, whose efforts were combined by the National Television System Committee (NTSC) and directed toward the creation of a single standard system, approved at the end of 1953 by the Federal Communications Commission. The work on creating a standard system of color television included investigations of the principles selecting the main parameters of the system, establishing tentative specifications for the system, manufacturing the equipment according to these specifications and finally, extensive field tests with the system, including the use of transmitters and receivers under conditions of wireless broadcasting, after which the final standard was developed.

During the stay in USA from 5 to 22 March, the CCIR delegates attended demonstrations of color television organized by the National Television Committee and by various firms. The delegates visited scientific research laboratories, studios and transmitters of color television, and also radio-relay lines for the transmission of color television. They observed the operation of color TV receivers with various tricolor kinescopes and were shown the manufacture of tricolor kinescopes with shadow mask, at the RCA factories in Lancaster.

The visit to color TV studios left the delegates with the impression that the work on creating color television programs has ceased to bear an experimental character, and that in this field much experience has already been accumulated.

In comparison with black-and-white TV studios, the work in color studios is more complex and expensive. Here more service personnel are required, great care is necessary in preparing and adjusting the camera, the choice of illumination, and so forth.

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Stiffer demands, in comparison with black-and-white television, are made on transmitters for color broadcasting and on radio-relay lines that transmit color programs. The most important additional demands as to radio-relay lines in the transmission of color television are the requirement for a smooth frequency characteristic in the range of the color subcarrier (3.58 megacycles) and the requirement for minimal distortions of the differential phase, i.e., distortions of the phase of the signal of the color subcarrier depending on the magnitude of the brightness signal. The delegates had the possibility of comparing, with the original, color pictures after transmission over interurban radio-relay lines; in all cases the quality of the pictures was rated as good.

In the color receivers which were shown during demonstrations, various types of tricolor kinescopes were used. The operating principles of two types, namely, of the color kinescope with shadow mask and of the Lawrence tube (chromatron), were examined in the above-mentioned article (Bibl.4). We note that the tricolor three-ray RCA kinescope with shadow mask and the 53-cm screen of the 21AXP22 type is the only color tube produced at present for sale. Its shortcomings are the difficulty of providing good convergence of the three rays over the entire screen and the large loss of brightness, caused by the mask blocking a large part of the electrons, so that only 15% reach the screen.

In the tube with post-acceleration, developed by the General Electric as well as in the chromatron, a line screen of three-color phosphors is used, together with a wire grid in front of them for focusing the rays. In contrast to the chromatron, however, the tube with post-acceleration has three rays. All the wires of the grid are connected here and are impressed here with one and the same potential of 5 kilovolts. Three guns are arranged in one plane, perpendicular to the strips of the screen and at a slight mutual slant, so as to have each ray strike the appropriate phosphor strip. The advantage of this type of tube over the tube with a shadow mask consists in that about 85% of the electrons reach the screen.

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The Philco Corporation has developed a tricolor kinescope with an indicating or tracking ray. This kinescope has one electron gun and a screen consisting of vertical phosphor strips. The electron ray created by the gun is resolved into two parts which consist of a working and a tracking ray. On the screen, the tracking ray intersects strips of some secondary-emission substance and creates a signal containing information on the position of the working ray at any given instant of time. This signal, after proper processing, is mixed with the signal supplied in the working ray for getting the color picture. The summary signal modulates the working ray in amplitude and phase so that the latter alternately excites the color phosphor in the proportion necessary for true reproduction of the color picture. The current of the tracking ray is low and gives no noticeable strobing of the picture. The advantage of the tracker tube over other types is the simplicity of design of the tube itself. But the control circuits outside the tube are fairly complex. Focusing of the ray on the screen in a small spot is also complex. The latter is necessary to prevent overlapping of several phosphor strips which would cause dis-saturation of colors.

In the laboratories of the Hazeltine Co., a series of interesting experiments were shown, confirming the standards used as basis of the NTSC system. The series of experiments was devoted to selection of the components of a full-color signal and widths of the assigned bands. The color pictures obtained with the NTSC signal, are comparable to pictures obtained with three simultaneous color signals at a bandwidth of 4 mc each. By means of color rings of various sizes, which were viewed at various distances, the properties of color vision, used as basis in selecting the signals I and Q, were demonstrated. Comparisons were also made of color pictures during variation in the bandwidths of the signals I and Q. The experiments in rating the effect on the color picture quality of the accuracy of maintenance of the amplitude and phase values of the color subcarrier showed that, for retention of good quality of the color picture, the phase variations must not exceed $\pm 5^{\circ}$ and the amplitude variations, ± 1.5 db. In a series of experiments on the action of noise and

and interference on color and black-and-white pictures, it was demonstrated that methods of constant radiance and frequency interlacing are highly favorable. Experiments on the interference resistance of color synchronization in the NTSC system showed that color synchronization is preserved even when the signal-to-noise ratio is insufficient to get a good picture. On the whole, the experiments confirmed the validity of the reason in the choice of the NTSC system.

France

The demonstration in France was organized by the Public Service of Radio and Television Broadcasting. The delegates were shown three experimental systems of color television, developed in France: the system of the line-alternating type known as "Anré de France" and developed by the RBV Society; the system of point-alternating type known as the system with "dual communication" developed by the Laboratories of Electronics and Physics; and the simultaneous code system of Valance. All three systems are still in the experimental stage and have not yet been developed to their final form. Therefore, the purpose of the demonstration was only to show these systems and not to recommend any kind of definite system or standard.

The Anré de France system is a line-alternating system with respect to red and green components of the picture. The full-color picture (one frame) is put together from two fields of scanning, each of which contains 409 lines and is transmitted in the course of 1/50 of a second. The lines in the two fields of one frame are not interlaced but are superimposed, i.e., progressive scanning takes place. The frame of the picture thus consists of 818 lines instead of 819 lines according to the French standard for black-and-white television. The picture during the first field contains 205 red lines, alternating with 204 green lines. During the second field of each frame, 205 green and 204 red lines are transmitted, which are superimposed, respectively, on the red and green lines of the first field. The blue signals exist continuously. They are transmitted in a subcarrier and create a blue component of

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the picture in each field. Thus, a picture is produced at the receiving end. In the transmitter to the alternately transmitted red and green signals, a blue signal is added in corresponding proportion. This is done for improving the compatibility since, in the reception of a color transmission on a black-and-white receiver, the picture is created by signals of all three primary colors in each line. The red-blue and green-blue components are transmitted with the 7-mc band, while the blue component is transmitted in the 8.47-mc subcarrier with double sideband modulation at a total 2-mc signal bandwidth of the modulated subcarrier. The amplitude of the subcarrier is maximally equal to 15% of the amplitude of the red or green signal. In the color receiver, after detection of the subcarrier, the separated blue signal is fed to the corresponding gun of the tricolor receiving tube, and is also fed with reversed sign to the red-blue and green-blue components. Thus restored, the red and green signals are fed to their guns of the tube. The separation of the red and green signals in the receiver is accomplished by the asymmetry introduced into the line-synchronizing pulses.

In the Anré de France system, in consequence of rejection of interlaced scanning, the vertical definition is diminished by half. The horizontal definition is also diminished because of the reduction of the frequency band of the video signal from 10.4 mc (the standard of French black-and-white television) to 7 mc. A corresponding reduction in definition occurs during reception of color transmissions in a black-and-white receiver. With few saturated colors and with saturated blue, the system operates like a system with 409 lines and 50 frames per second with progressive scanning, but on transmission of saturated red and green colors, it operates like a system with 409 lines and 25 frames a second with interlaced scanning. The visibility of the color subcarrier on the screen of the black-and-white receiver is reduced by selecting the subcarrier frequency as an odd multiple of half the line frequency. In the reception of black-and-white transmissions on color receivers a vertical definition is preserved of 819 lines with interlaced scanning.

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The system with "dual communication", similar to the Anré de France system, is a system of alternating type for red and green colors; however, in contrast to the latter, this system is of the point-alternating type. The main feature of this system is the use of one subcarrier with its seemingly double amplitude modulation. This is achieved in the following manner: Two series of sinusoidal pulses with a repitition rate equal to the frequency of the subcarrier, are amplitude-modulated by red and green signals. Then, the series modulated by red signals is shifted 180° in phase and, in reversed polarity, is combined with the series modulated by the green signals. As a result, a signal is obtained with a subcarrier frequency whose positive envelope contains information on the green and whose negative envelope gives data on the red. During such transmission, large cross interferences occur between the red and green signals, which can be considerably reduced by transmitting [instead of red (R) and green (G)] signals of the type $R + \varepsilon G$ and $G + \varepsilon R$, where ε is a small magnitude. This correction of signals must be introduced after correction of gamma, since it presupposes a linear system. The blue signal is transmitted, as it is also in the Anré de France system, on a separate subcarrier outside the spectrum of video frequency. The full-color picture is composed of four fields, during which line-and-point interlacing takes place, i.e., the number of full-color pictures per second amounts to 12.5. The green and red signals are separated in the receiver by dual detection; in one of them the positive half-waves of the red-green subcarrier are reinserted and in the other, the negative half-waves. The chief defects of this system are, first, the presence of two closely spaced subcarriers, the blue subcarrier being close to the sound carrier, and, second, that the green and red signals can each be used only to one-half the total modulation depth. The system is not wholly compatible, since the picture in the black-and-white receiver will be created only by the green signal. The reception of black-and-white transmissions by color receivers can be accomplished normally, with deflection of filtration circuits and dual detection. But the system is designed for operation at new high-frequency

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ranges in new black-and-white compatible receivers. These receivers will be similar to color receivers with the exception of the tubes; in these, the three-color signals will be used for composing the brightness signal which is then fed to the tube creating the normal black-and-white picture. Of course, such receivers will be more expensive than the ordinary black-and-white receivers.

The demonstrated color pictures still had many defects such as color fringe, insufficient definition, etc., which can be attributed to the fact that the work on the systems is still incomplete.

Apart from the demonstration of the systems, the delegates were shown a series of interesting experiments on determining the quality of the color picture in the two systems described, by means of special test charts. The prepared charts were designed for determining such qualitative indices of the color-television channel as linearity of the transmission of amplitude, cross distortions between the three-color channels, fidelity of the color reproduction, and picture definition. Also shown were experiments in the study of picture flicker in the three separate primary components. The experiments were made at frequencies of 33.3 and 50 cps and showed, in passing, that flicker practically does not depend on the blue primary, and that it decays to zero in the white and yellow, when the red and green primaries are in the antiphase.

The operating principle of the third system, i.e., the Valance system, was briefly described in another paper (Bibl.4).

England

The present state of television in England is characterized by a rather broad development of black-and-white TV broadcasting. The BBC television network, using five frequency channels in a range to 100 megacycles, was by the end of the summer 1956 to have supplied 97% of the population with television broadcasts. The country has 6 million TV receivers, i.e., about 40% of all families have receivers. Si STAT

September 1955, a commercial television organization also transmits a television program. Owing to the presence of a large network of black-and-white TV receivers England, just as the USA, also is interested in introducing a compatible system of color television.

The first experiments in creating a system of color television were directed toward developing a system of the NTSC type, adapted to the English television standard (405 lines, bandwidth of video signal 3 megacycles with a channel width of 5 megacycles). This work was begun about two years ago and was carried out jointly by the BBC, the General P.O. Department, and the radio industry. The BBC investigated in detail the NTSC system and the problems involved in its conversion to the British standard, and in October 1955 began to conduct experimental transmissions of color television from the telecenter in Alexander Palace. The radio industry developed studio and radio relay equipment, transmitters and other apparatus for standards at 405 and 625 lines, and also some experimental types of color receivers in 405 lines. The General P.O. Department examined problems associated with the transmission of color TV programs by cable and radio-relay lines, and with the frequency distribution when color television is introduced.

It should be noted that, under the English standard, it is a great deal more difficult to use combinations of the spectra of the brightness and coloration signals than it is under European or American standards, because of the more narrow band of the picture signal (3 mc against 5 and 4.2 mc, respectively). In England there are advocates for a transition to a black-and-white television standard higher in definition (which would facilitate the combination of spectra) and of the use, for color television, of higher frequency bands, where it would be possible to employ separate transmission of the brightness and coloration signals. These questions are as yet undecided just as is also the question as to the expediency of adopting a system of the NTSC type under a standard at 405 lines.

The demonstrations of color television in England included a showing of a STAT tem of the NTSC type with standards at 405 and 625 lines, a showing of a series of experiments, in particular, on the effect of various types of interference on black-and-white and color receivers, and a showing of various equipment, in particular, a color studio and a radio-relay line for transmission of color television. In the majority of color receivers, tricolor RCA kinescopes with shadow mask and 53-cm screen were used.

In the demonstration of the BBC system of the NTSC type, remodeled for the standard at 405 lines, the following alternates of signal transmission were shown: transmission of three-color signals in three separate channels with a bandwidth of 3 mc for each; transmission of the brightness (Y) and coloration (I and Q) signals in separate channels with a width of 3.1 and 0.4 mc, respectively, i.e., with an overall bandwidth of 4.4 mc; transmission of the same signals with a reduction in the bandwidth for the signal I to 0.4 mc; and, finally, transmission by the NTSC type, i.e., transmission of three signals in a common video band of 3 mc. The frequency of the color subcarrier, in this case, was about 2.66 mc and the bands for the signals I and Q were 1 and 0.4 mc, respectively.

The EMI Co. demonstrated several alternates of transmission, with conversion of the NTSC system to 625 lines: broad-band system with transmission of the signals Y, R - Y and B - Y in three channels of 5 mc each; system with average bandwidth, with transmission of the signal Y in a band of 5 mc; and signals I and Q with bands of 1.5 and 0.5 mc, respectively, during quadrature modulation of the subcarrier arranged outside the spectrum of the brightness signal; and, finally, a narrow-band system with a channel bandwidth of 7 mc at a subcarrier frequency of about 4.4 mc.

The EMI Co. also demonstrated equipment for conversion of field-alternating signals into simultaneous signals by means of optical-electronic transcription of signals. This device, analogous to what is known as the "chromacoder" developed in America by the Columbia Broadcasting Company, was conceived with the object of avoiding the difficulty of creating color television cameras with three transmitting

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tubes (the difficulty of exact recording of three pictures, large dimensions and weight of the camera, complex controls, etc). When the converting device is available in the studio, small and simple cameras can be used of the field-alternating type with rotating disks with light filters. The three alternating light signals from such a camera are fed to three kinescopes, each of which reproduces the picture in one of the primary colors. In front of the screen of each kinescope stands a transmitting tube, scanning the respective picture. From the outputs of the three transmitting tubes simultaneous color signals are taken, which can then be processed in the usual manner. In such conversion of signals, all difficulties of recording are transferred from the camera to the transcription unit, which can be one or several cameras. The control of the picture from each camera can be accomplished by monitors of the field-alternating type. The transcription unit can be used for conversion of the standards of scanning. Thus in the device demonstrated, the color camera operated on a standard at 405 lines, while the transcription was made at a standard of 625 lines. The transcription unit had automatic balancing of variations of the sensitivity of transmitting tubes according to the photocathode surface in each of the converter channels.

Holland

The demonstrations in Holland included visits to the PTT laboratories at Leydendam and the television installation at Rosendal, and acquaintance with the research laboratories of the Phillips Co.

The Phillips Co. demonstrated two alternates of the NTSC system at 625 lines and a system developed by the firm, with two color subcarriers. In the first alternate of the NTSC system, the frequency of the subcarrier was about 4.43 mc, and transmission of color was accomplished by means of the signals I and Q with bands of 1.3 and 0.5 mc, respectively. In the second alternate of the NTSC system, the subcarrier with a frequency of about 4.1 mc was modulated by the color signals R - Y

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and B - Y with bandwidths of 0.9 mc each, in transmission with two sidebands. The transmission of coloration signals in equal bands is known to eliminate the causes for the rise of cross interference, but leads to a narrower frequency of the subcarrier which makes it more noticeable.

In the system with two color subcarriers, the brightness signal is transmitted in the capacity of fundamental signal, while the red and blue color signals are transmitted in the two subcarriers. The frequencies of the subcarriers in the system at 625 lines were selected as approximately 3.59 and 4.64 mc. The main difficulty in this system consists in reducing to a minimum the effect of interference from the subcarriers on the screen. This interference is created by each carrier separately as well as by the beat interference between them, which is especially harmful since the beat frequency is lower than the frequency of each individual subcarrier. To reduce the conspicuousness, the frequency of the lower subcarrier is taken equal to an odd multiple of half the line frequency, and moreover, its phase is shifted alternately by $\pm 90^{\circ}$ at the beginning of each field. The frequency of the upper subcarrier is a multiple of the line frequency, and its phase is shifted by $\pm 180^{\circ}$ at the beginning of each field. Such a selection of subcarriers reduces the conspicuousness both of the subcarriers themselves and of the beat interference between them. A transmission is used in subcarriers with suppression of the upper sideband. The width of the red video signal amounts to 2 mc and that of the blue signal to 1 mc. The green signal is derived in the receiver by means of a matrix.

Demonstrations on a large projection screen permitted a comparison of the three systems: transmission of each color in a 7-mc band; a system of the NTSC type with a subcarrier of 4.43 mc; and a system with two subcarriers. The difference between the first system and the remainder was negligible, even at a short distance from the screen. This can be partially explained by defects of the projection system, since the line structure of the picture was not noticeable even at a short distance.

During their visit to the color television studio, the delegates had an oppor-

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tunity to see two types of three-color cameras - in image orthicons and in vidicons.

The demonstrations were ended with a showing of tricolor receiving tubes of the RCA type with shadow mask and tubes of the Phillips Co. patterned on the type of the above-described General Electric and Philco tubes.

In conclusion, it should be noted that the best and most developed system at present is the NTSC system. The other original systems described in the present survey have not as yet been developed to such a stage that they might be compared with the NTSC system. For the time being, no definite advantages of any of these systems over the NTSC system can be detected.

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TECHNICAL CALCULATIONS OF ERRORS OF ATTENUATORS

by

D.D.Voyeykov

A method is described of computing the errors of attenuators, operating with constant input voltage. The final calculation results are reduced to Tables and Graphs which permit determining the error of an attenuator as a function of the tolerances of its elements and the magnitude of the attenuation introduced.

Introduction

In the design of attenuators, one of the main problems is providing the required accuracy of the transmission factor. The method of approximate estimate of the errors of attenuators, which is prevalent at present (Bibl.1), leads to excessively rigid tolerances in resistances or to lowering of the accuracy of the attenuator transmission factor. The method proposed in the present article, while not inferior in simplicity to that previously adopted, permits, at the same time, a derivation of mathematically substantiated results.

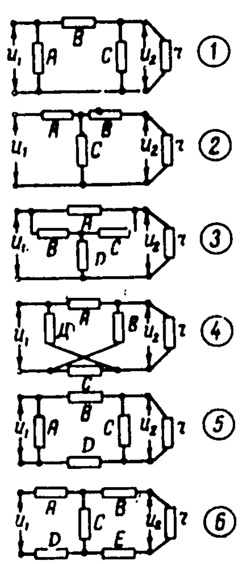

Fig.1

Analysis of Errors of Attenuator Sections

For simplification of the analysis, attenuators will be examined that operate at constant input voltage rather than at the constant electromotive force of a generator. Such attenuators are used in the Russian signal generators ZG-11, ZG-12, I-101, and others. As shown below, the derived results can, with sufficient

Table 1

Circuit Type and Number in Fig.1	Formula of Transmission Factor $T = \frac{u_1}{u_2}$	Relationship between Sections, when R = r
Pi-network (No.1)	$1+\frac{B}{r}+\frac{B}{C}$	$A=C=r\,\mathrm{ctg}\frac{\theta}{2}=$ $=r\frac{T+1}{T-1}$ $B=r\,\mathrm{sh}\,\theta=r\frac{T^2-1}{2T}$
T-network (No.2)	$\frac{AB+A(r+C)+C(B+r)}{Cr}$	$A=B=r\,\mathrm{tg}\frac{\theta}{2}=$ $=r\frac{T-1}{T+1}$ $C=r\frac{1}{\mathrm{sh}\,\theta}=r\frac{2T}{T^2-1}$
Bridged T-network (No.3)	$1+\frac{A(BC+BD+Br+CD)}{r(AD+BC+BD+CD)}$	$B=C=r$ $A=r(T-1)$ $D=r\frac{1}{T-1}$
Bridged (No.4)	$\frac{ABC+ABD+ACD+ACr+ADr+BCD+BCr+BDr}{r(AC-BD)}$	$B=D=r\,\mathrm{tg}\frac{\theta}{2}=$ $=r\frac{T-1}{T+1}$ $A=C=r\,\mathrm{cth}\frac{\theta}{2}=$ $=r\frac{T+1}{T-1}$
O-network (No.5)	$1+(B+D)\left(\frac{1}{r}+\frac{1}{C}\right)$	$A=C=r\,\mathrm{cth}\frac{\theta}{2}=r\frac{T+1}{T-1}$ $B=D=\frac{1}{2}r\,\mathrm{sh}\,\theta=$ $=r\frac{T^2-1}{4T}$
H-network (No.6)	$\frac{(A+C+D)(B+C+E+r)}{Cr}-\frac{C}{r}$	$A=B=D=E=\frac{1}{2}r\,\mathrm{tg}\frac{\theta}{2}=$ $=\frac{1}{2}r\frac{T-1}{T+1}$ $C=r\frac{1}{\mathrm{sh}\,\theta}=r\frac{2T}{T^2-1}$

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accuracy, be applied to any cases of operation of attenuators, if the errors of the sections of the latter do not exceed 0.5%.

In the general case, the circuit of any attenuator consists of a series of very simple sections pictured in Fig.1. The errors of the individual sections can be determined by the differentiation of the transmission factor.

We introduce the following symbols:

u_1, u_2 = input and output voltages, respectively, of the attenuator;

R, r = input and output resistances, respectively, of the attenuator;

A, B, C, ..., M = resistances of attenuators according to Fig.1;

T = f (A, B, C, ..., M) = transmission factor;

N = attenuation in decibels;

θ = attenuation in nepers.

It is obvious that

$$T=\frac{u_1}{u_2},\quad T=10^{\frac{N}{20}},\quad N=8{,}68\theta.$$

Applying Kirchhoff's laws, we derive the formulas for the transmission factor, given in Table 1.

Differentiating these formulas, we derive the particular errors

$$\frac{\partial T_A}{\partial A},\ \frac{\partial T_B}{\partial B},\ \frac{\partial T_C}{\partial C},\ \ldots,\ \frac{\partial T_M}{\partial M}.$$

and through them the total relative error of the transmission factor of a section

$$\varepsilon=\frac{\Delta T_A}{T}+\frac{\Delta T_B}{T}+\frac{\Delta T_C}{T}+\ldots+\frac{\Delta T_M}{T}, \tag{1}$$

and the probable relative error of the transmission factor of a section

$$\sigma=\sqrt{\left(\frac{\Delta T_A}{T}\right)^2+\left(\frac{\Delta T_B}{T}\right)^2+\left(\frac{\Delta T_C}{T}\right)^2+\ldots+\left(\frac{\Delta T_M}{T}\right)^2} \tag{2}$$

The cumbersome expressions for ξ and σ are substantially simplified for the case R = r, and in such form are they represented in Table 2.

Table 2

Circuit Type and Number in Fig.1	Error Equations
Pi-network (No.1)	$\xi = \frac{(T-1)^2}{2T^2}\left[b_C + \frac{2T}{T-1}b_B + \frac{T+1}{T-1}b_r\right]$ $\sigma = \frac{(T-1)^2}{2T^2}\sqrt{b_C^2 + \frac{4T^2}{(T-1)^2}b_B^2 + \frac{(T+1)^2}{(T-1)^2}b_r^2}$
T-network (No.2)	$\xi = \frac{T-1}{2T^2(T+1)}[2T^2 b_A + (T^2+1)b_B + 2T(T-1)b_C + (T+1)^2 b_r]$, $\sigma = \frac{T-1}{2T^2(T+1)}\sqrt{4T^4b_A^2 + (T^2+1)^2b_B^2 + 4T^2(T-1)^2b_C^2 + (T+1)^4b_r^2}$
Bridged T-network (No.3)	$\xi = \frac{1}{2T^2}[(T^2-1)b_A + (T-1)b_B + (T-1)^2b_D + (T^2-1)b_r]$ $\sigma = \frac{1}{2T^2}\sqrt{(T^2-1)^2b_A^2 + (T-1)^4b_B^2 + (T-1)^4b_D^2 + (T^2-1)^2b_r^2}$
Bridged (No.4)	$\xi = \frac{T^2-1}{8T^2}[(T+1)b_A + (T-1)b_B + (T+1)b_C + (T-1)b_D + 4b_r]$ $\sigma = \frac{T^2-1}{8T^2}\sqrt{(T+1)^2b_A^2 + (T-1)^2b_B^2 + (T+1)^2b_C^2 + (T-1)^2b_D^2 + 16b_r^2}$
O-network (No.5)	$\xi = \frac{T-1}{2T^2}[Tb_B + Tb_D + (T-1)b_C + (T+1)b_r]$ $\sigma = \frac{T-1}{2T^2}\sqrt{T^2b_B^2 + T^2b_D^2 + (T-1)^2b_C^2 + (T+1)^2b_r^2}$
H-network (No.6)	$\xi = \frac{T-1}{4T^2(T+1)}[2T^2(b_A + b_D) + (T^2+1)(b_B + b_E) +$ $+ 4T(T-1)b_C + 2(T+1)^2b_r]$ $\sigma = \frac{T-1}{4T^2(T+1)} \times$ $\times \sqrt{4T^4b_A^2 + 4T^4b_D^2 + (T^2+1)^2(b_B^2 + b_E^2) + 16T^2(T-1)^2b_C^2 + 4(T+1)^4b_r^2}$

The Table contains the following specifications:

$$b_A = \frac{\Delta A}{A},\ b_B = \frac{\Delta B}{B},\ b_C = \frac{\Delta C}{C}, \ldots,\ b_M = \frac{\Delta M}{M}.$$

The magnitudes b_A, b_B, b_C, ..., b_M can mean either the tolerances in the corresponding resistances or the relative errors of resistances. If b_A, b_B, b_C, ..., b_M are expressed in percentages, then ξ and σ are also expressed in percentages. Most frequently, all resistances of a section have identical tolerances. After assuming $b_A = b_B = b_C = \ldots = b_M = b$, the derived formulas can be considerably simplified

Table 3

Circuit Type and Number in Fig.1	Error Equations	Maximum Errors
Pi-network (No.1)	$\xi = 2b\left(1-\frac{1}{T}\right)$; $\sigma = b\left(1-\frac{1}{T}\right)\sqrt{1{,}5+\frac{1}{2T^2}}$;	$\xi_{max} = 2b$ $\sigma_{max} = 1{,}22b$
T-network (No.2)	$\xi = b\left(1-\frac{2}{T+1}\right)\left(3+\frac{1}{T^2}\right)$; $\sigma = b\left(1-\frac{2}{T+1}\right)\sqrt{2{,}5-\frac{1}{T}+3\frac{1}{T^2}+\frac{1}{T^3}+\frac{1}{T^4}}$;	$\xi_{max} = 3b$ $\sigma_{max} = 1{,}58b$
Bridged T-network (No.3)	$\xi = 2b\left(1-\frac{1}{T}\right)$; $\sigma = b\left(1-\frac{1}{T}\right)\sqrt{1+\frac{1}{T^2}}$;	$\xi_{max} = 2b$ $\sigma_{max} = b$
Bridged (No.4)	$\xi = \frac{1}{2}b\left(1-\frac{1}{T^2}\right)(T+1)$; $\sigma = b\,\frac{T}{4}\left(1-\frac{1}{T}\right)\sqrt{1+\frac{5}{T^2}}$;	$\xi \to \infty$ $\sigma \to \infty$
O-network (No.5)	$\xi = 2b\left(1-\frac{1}{T}\right)$; $\sigma = b\left(1-\frac{1}{T}\right)\sqrt{1+0{,}5\frac{1}{T^2}}$;	$\xi_{max} = 2b$ $\sigma_{max} = b$
H-network (No.6)	$\xi = b\left(1-\frac{2}{T+1}\right)\left(3+\frac{1}{T^2}\right)$; $\sigma = b\left(1-\frac{2}{T+1}\right)\sqrt{1{,}87+\frac{1}{T}+2{,}75\frac{1}{T^2}+\frac{1}{T^3}+0{,}375\frac{1}{T^4}}$	$\xi_{max} = 3b$ $\sigma_{max} = 1{,}36b$

For any section circuit, the data of Table 3 can be represented in the form:

$$\xi = k_1 b, \tag{3}$$

$$\sigma = k_2 b. \tag{4}$$

Proceeding from this and knowing the allowable error of the transmission factor α and the maximum attenuation of the section, the tolerance in the section resistance can easily be determined by the formula

$$b = \frac{\alpha}{k_1}. \qquad (5)$$

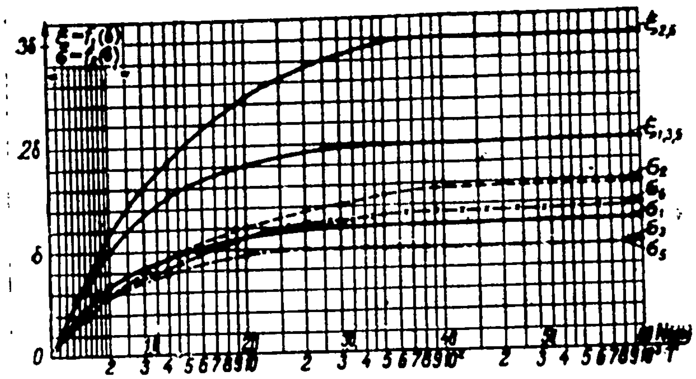

Fig.2

The values of the factor k_1, and if necessary also of k_2, are easily determined according to Fig.2 or Table 3.

Analysis of Errors of Complex Circuits of Attenuators

It is easy to prove that the total relative error of the transmission factor of an attenuator, consisting of a series of very simple sections, is expressed by the equation

$$\xi_T = \xi_{T1} + \xi_{T2} + \xi_{T3} + \ldots + \xi_{Tn}, \qquad (6)$$

where ξ_{T1}, ξ_{T2}, ξ_{T3}, ..., ξ_{Tn} are the total relative errors of individual sections of the attenuator.

Analogously,

$$\sigma_T = \sqrt{\sigma_{T1}^2 + \sigma_{T2}^2 + \sigma_{T3}^2 + \ldots + \sigma_{Tn}^2}, \qquad (7)$$

where σ_{T1}, σ_{T2}, σ_{T3}, ..., σ_{Tn} are the probable relative errors of individual sections of the attenuator.

The magnitudes ξ_{T1}, ξ_{T2}, ξ_{T3}, ..., ξ_{Tn} and σ_{T1}, σ_{T2}, σ_{T3}, ..., σ_{Tn} are determined according to Table 2 or Table 3 and Fig.2.

It should be noted that, if all elements of a section have tolerances of $\pm b$, then the input resistance of the section differs from the rating by not more than $\pm b$.

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This should be taken into account during calculations of the errors of complex circuits of attenuators. Moreover, this circumstance permits applying the graph of Fig.2 for computing attenuators operating at constant electromotive force of a generator, when the tolerances on elements of the sections are not more than 0.5%.

Conclusions

1. The errors of the transmission factor of attenuator sections can be unilaterally determined according to the cited Table 2, Table 3, and the graph in Fig.2.

2. The errors of the transmission factor of complex circuits of attenuators can be determined on the basis of the errors of sections according to eqs.(6) and (7).

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PULSE DISTORTIONS IN VOICE-FREQUENCY TELEGRAPH CHANNELS WITH PHASE MODULATION DURING THE ACTION OF INTERFERENCE

by

A.M.Zingerenko

Distortions of the pulse period in PM channels during harmonic, fluctuation, and pulse interference are determined. It is demonstrated that, with respect to distortions due to fluctuation and pulse interferences, the PM channel, while having considerable advantage over the AM channel, is inferior to the FM channel.

Introduction

The system of transmitting telegraph pulses by variation of the initial phase of the carrier oscillation by $\pm\pi$ has not yet emerged from the stage of experimental tests. The difficulty of realizing such a system is known to consist of the need to have, at the receiving end, a generator of the carrier frequency, which is synchronized in phase with the oscillator of the transmitter. With random variation of the phase of the received signals or the carriers, negative reproduction of the received signals is observed in the receiver.

Despite the mentioned difficulties, attempts have been made to realize voice frequency telegraphy with phase modulation (PM). In this connection, it is expedient to compare the stability of the given system of voice-frequency telegraphy with the stability of AM and FM systems with respect to pulse-duration distortions, during the action of interference of various types.

Transient Conditions during Phase Keying

The transient current at the output of a filter, during variation of the ini-

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tial phase of the source of a unit electromotive force by $\pm\pi$, can be represented by the equation

$$i(t) = \sin\omega_1 t - 2i_1(t), \tag{1}$$

where $i_1(t)$ is the transient current at the filter output when a unit electromotive force with a frequency ω_1 is switched on.

In the general case, the current $i_1(t)$ can be represented in the form

$$i_1(t) = g_1(t)\sin\omega_1 t + g_2(t)\cos\omega_2 t, \tag{2}$$

where $g_1(t)$ and $g_2(t)$ are, respectively, the cophasal and quadrature components of the current $i_1(t)$.

Substituting the value $i_1(t)$ in eq.(1), we derive

$$i(t) = [1 - 2g_1(t)]\sin\omega_1 t - 2g_2(t)\cos\omega_2 t. \tag{3}$$

We represent $i(t)$ in the form of an oscillation with transient amplitude and initial phase

$$i(t) = A(t)\sin[\omega_1 t + \Theta(t)]. \tag{4}$$

In accordance with eq.(3),

$$A(t) = \sqrt{[1 - 2g_1(t)]^2 + 4g_2^2(t)}, \tag{5}$$

$$\Theta(t) = -\operatorname{arctg}\frac{2g_2(t)}{1 - 2g_1(t)}. \tag{6}$$

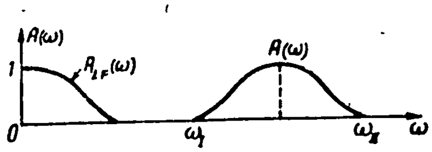

Fig.1

As is known, if the filter pass band is relatively narrow, the phase characteristic is linear, and the transmission-factor characteristic is symmetric relative to ω_1, then

$$g_2(t) \approx 0,$$

$$g_1(t) \approx \frac{1}{2} + \frac{1}{\pi}\int_0^{\omega_{LF}} A_{LF}(\omega)\frac{\sin\omega t}{\omega}\,d\omega.$$

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Here $A(\omega_1)$ is the value of $A(\omega)$ at ω_1;

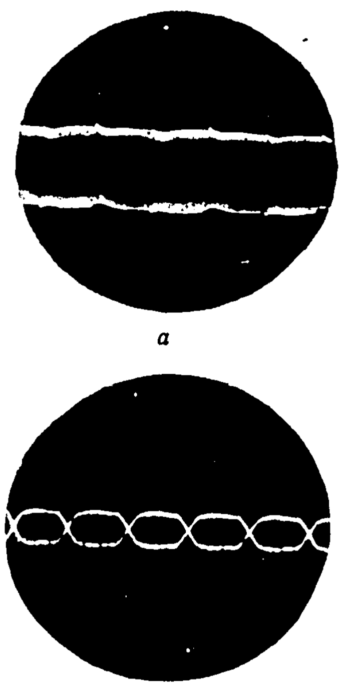
$A_{LF}(\omega)$ is the amplitude-frequency characteristic of the equivalent LF filter (Fig.1).

In this case,

$$A(t)=\frac{2}{\pi}\int_0^{\omega_{LF}} A_{LF}(\omega)\frac{\sin\omega t}{\omega}\,d\omega. \tag{7}$$

$$\Theta(t) = \pm\pi. \tag{8}$$

Consequently, with a relatively narrow pass band and a symmetrical arrangement of ω_1 in the pass band, the oscillation phase will be set instantly. But the oscillation envelope at the instant of phase variation will assume a zero value (Fig.2), i.e., t = 0 necessitates that A(t) = 0. Shown in Fig.2a and b are voltage oscillograms during the absence and presence of the filter (pass band, 170 cps).



a

b

Fig.2

The steepness of the build-up of the envelope can be determined from the derivative A(t) according to time, when t → 0. Besides,

$$S_{rel.\,PM}=\lim\frac{dA(t)}{dt}=\frac{2}{\pi}\int_0^{\omega} A\quad(\omega)\,d\omega,$$

or

$$S_{rel\ PM}=\frac{\omega_{II}-\omega_I}{\pi}=2(f_{II}-f_I), \tag{9}$$

where $f_{II} - f_I$ is the equivalent filter pass band.

With sudden switching on of alternating electromotive force the relative steepness of the amplitude build-up is

$$S_{rel\ AM}=f_{II}-f_I. \tag{10}$$

As is logical with PM, the relative steepness of build-up of the pulse amplitude will be twice as great as with AM.

Distortions during Harmonic Interference

During transmission of signals by variations in the initial phase, the voltage of the signal can be represented by the equation

$$u_c = A_c(t)\sin[\omega_0 t + \Theta(t)] = A_c(t)[\sin\omega_0 t\cos\Theta(t) + \cos\omega_0 t\sin\Theta(t)],$$

where $A_c(t)$ is the signal envelope at the output of the receiver filter;

$\Theta(t)$ is variation of the initial phase during transmission of the signal.

Assume that in the pass band of the receiver a harmonic interference acts, with a voltage of

$$u_n = U_n\sin[(\omega_0 + \Omega_n)t + \varphi_0] = U_n[\cos(\Omega_n t + \varphi_0)\sin\omega_0 t + \sin(\Omega_n t + \varphi_0)\cos\omega_0 t],$$

where Ω_n is the difference in angular frequencies of the signal and interference;

φ_0 is the initial phase of interference.

The total voltage of signal and interference is

$$u = u_c + u_n = [A_c(t)\cos\Theta(t) + U_n\cos(\Omega_n t + \varphi_0)]\sin\omega_0 t + [A_c(t)\sin\Theta(t) + U_n\sin(\Omega_n t + \varphi_0)]\cos\omega_0 t.$$

We represent this voltage in the form

$$u = A(t)\sin[\omega_0 t + \Psi(t)],$$

where

$$A(t) = A_c(t)\sqrt{1 + K_n'^2 + 2K_n'\cos[\Theta(t) - \Omega_n(t) - \varphi_0]},$$

$$\Psi(t) = \operatorname{arctg}\frac{\sin\Theta(t) + K_n'\sin(\Omega_n t + \varphi_0)}{\cos\Theta(t) + K_n'\cos(\Omega_n t + \varphi_0)},$$

$$K_n' = \frac{U_n}{A_c(t)}.$$

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As is known, the voltage of the signal at the output of the phase detector will be equal to

$$U = A(t)\sin\Psi(t).$$

Within the limits of a pulse of a given sign, $\Theta(t)$ is equal to 0 or π. Then,

$$A(t) = A_c(t)\sqrt{1 + K_n'^2 \pm 2K_n'\cos(\Omega_n t + \varphi_0)}, \quad (11)$$

$$\Psi(t) = \pm \operatorname{arctg}\frac{K_n'\sin(\Omega_n t + \varphi_0)}{1 + K_n'\cos(\Omega_n t + \varphi_0)}. \quad (12)$$

The greatest value $\psi(t)$ will be at $\sin(\Omega_n t + \varphi_0) = \pm 1$. In this case, at $K_n' < 1$.

$$|\Psi| = \operatorname{arctg} K_n' < \frac{\pi}{4}.$$

Consequently, when $K_n' < 1$ the interference cannot cause variations in the pulse sign.

The greatest variation in the amplitude $A(t)$ will be at $\cos(\Omega_n t + \varphi_0) = \pm 1$. In this case,

$$A(t) = A_c(t)(1 \mp K_n') = A_c(t) \mp K_n'$$

$$\Psi(t) = 0.$$

Thus, the pulse-duration distortions in the channel with PM will be determined (just as in the channel with AM) by the greatest increase of the signal envelope

$$\Delta A_{max} = U_n.$$

We now determine the magnitude of the pulse-duration distortion. In accordance with Fig.3, the variation of pulse duration is

$$\Delta\tau = 2t_0 = \frac{2}{S_{rel\ PM}}\frac{U_n}{U_c} = \frac{2K_n}{S_{rel\ PM}},$$

where

$$K_n = \frac{U_n}{U_c}.$$

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The pulse-duration distortions are equal to

$$I_{PM} = \frac{\Delta\tau}{\tau}\,\% = \frac{2K_n}{S_{rel\ PM}}\, B\cdot 100\%.$$

Substituting the value $S_{rel\ PM}$, we derive

$$I_{PM} = \frac{K_n}{f_{11} - f_1} B\cdot 100\,\% \qquad (13)$$

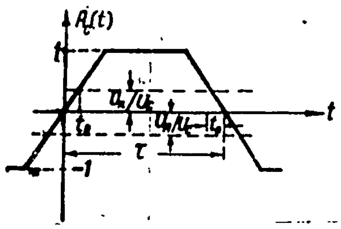


Fig.3

If the increase of the filter attenuation (Δb), with variation of the interference frequency is considered, then

$$I_{PM} = \frac{K_n e^{-\Delta b}}{f_{11} - f_1} B\cdot 100\,\%. \qquad (14)$$

At $\Omega_n = 0$, when the interference frequency coincides with the average frequency of the filter pass band, the distortions will be maximal ($\Delta b = 0$) and will be determined by eq.(13).

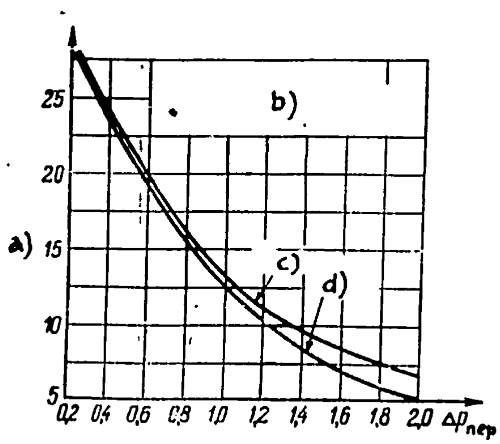


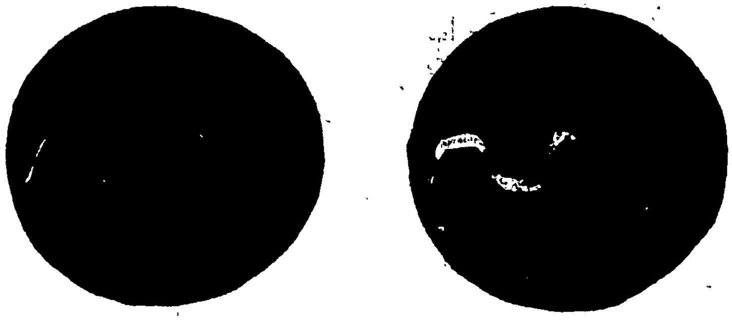
Fig.4

a) Maximum Distortions in %; b) Channel width, 170 cps; Speed of telegraphing 60 bauds; c) Measurement; d) Calculation

Plotted in Fig.4 are the experimental and calculated curves for the dependence of maximum distortions on the difference of the levels of the signal and the interference $\Delta p = \ln\frac{1}{K_n}$. Figure 5a and b shows oscillograms of the voltage at the output of the phase detector in the absence of interference and in the presence of interference, with a level lower than the signal level by 0.6 nep.

It is obvious that the experimental data agree well with the calculated values.

As shown previously, in the case of phase modulation, the steepness of pulse build-up at equal channel width will be twice as great as in the case of amplitude modulation. Therefore, the distortions of pulse duration in a channel with PM, at equal



a *b*

Fig. 5

channel width and equal level of interference, will be twice lower than in the channel with AM, i.e.,

$$I_{PM} = 0{,}5\, I_{AM}. \qquad (15)$$

Approximately the same correlation of maximum distortions occurs for channels with FM and AM. Consequently, the following comparative data can be compiled for the maximum distortions due to harmonic interference, for various kinds of modulation

$$I_{PM} = I_{FM} = 0{,}5\, I_{AM}. \qquad (16)$$

Distortions during Fluctuation Interference

The voltage of the fluctuation interference at the filter output can be represented in the form

$$u_n = U_1(t)\sin\omega_0 t + U_2(t)\cos\omega_0 t, \qquad (17)$$

where ω_0 is the average frequency of the filter pass band,

$$U_1(t) = \sum_{-\Omega_I}^{\Omega_{II}} U_{mq} A(\Omega)\cos(\Omega_q t + \varphi_q), \qquad (18)$$

$$U_2(t) = \sum_{-\Omega_I}^{\Omega_{II}} U_{mq} A(\Omega)\sin(\Omega_q t + \varphi_q). \qquad (19)$$

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Here U_{mq} denotes the amplitude of components of the continuous interference spectrum (identical at all Ω_q);

ω_q is the angular frequency of these components;

$\Omega_q = \omega_q - \omega_o$ is the amplitude-frequency characteristic of the transmission factor, offset in the axis of frequencies to the left by ω_o (Fig.6);

$$\Omega_{\rm I} = \omega_0 - \omega_{\rm I}; \quad \Omega_{\rm II} = \omega_{\rm II} - \omega_0.$$

The signal voltage, within the limits of a pulse of given sign, reads

$$u_c = A_c(t)\sin\omega_0 t. \tag{20}$$

In accordance with eqs.(17) and (20), the resulting voltage at the output of the receiver will be

$$u = [A_c(t) + U_1(t)]\sin\omega_0 t + U_2(t)\cos\omega_0 t.$$

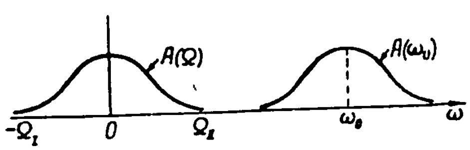

Fig.6

We represent this voltage in the form

$$u = A(t)\sin[\omega_0 t + \Psi(t)]. \tag{21}$$

The envelope of interest to us is equal to

$$A(t) = \sqrt{A_c^2(t) + 2A_c(t)U_1(t) + U_n^2(t)}, \tag{22}$$

where

$$U_n^2(t) = U_1^2(t) + U_2^2(t).$$

At $\overline{U_1^2(t)} \ll U_c^2$ we derive the approximate equation

$$A(t) \approx A_c(t) + U_1(t). \tag{23}$$

Consequently, the increment of the signal envelope, during the action of a fluctuation interference, is equal to

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$$\Delta A(t) = U_1(t). \tag{24}$$

Using eq.(13), which determines pulse-duration distortions in harmonic interference, and substituting $K_n = \frac{U_1(t)}{U_c}$ for $K_n = \frac{U_n}{U_c}$, we derive

$$I = \frac{U_1(t)}{U_c(f_{11} - f_1)} B \cdot 100\%. \tag{25}$$

The function $U_1(t)$ is a random quantity, subject to the normal law of distribution from a center equal to zero. Consequently, the magnitude I will also be a random quantity, with the same law of distribution. At the same time, the probability distribution of distortions is wholly determined by the root-mean-square value of distortions, i.e., by the value

$$\sigma_{PM} = \sqrt{\overline{I^2_{PM}}} = \frac{\sqrt{\overline{U_1^2(t)}}}{\sqrt{2}\, U_{cd}(f_{11} - f_1)} B \cdot 100\%,$$

where U_{cd} is the effective value of the signal voltage.

Since $\overline{U_1^2(t)} = U_n^2(t)$, where $U_n^2(t)$ is the mean-square effective value of the interference voltage, then

$$\sigma_{PM} = \frac{\sqrt{\overline{U_n^2(t)}}}{\sqrt{2}\, U_{cd}(f_{11} - f_1)} B \cdot 100\%. \tag{26}$$

The quantity $\sqrt{U_n^2(t)}\ U_{cd}$ determines the ratio of the RMS values of the voltages of the fluctuation interference to the signal.

We denote this ratio by

$$K_n = \frac{\sqrt{\overline{U_n^2(t)}}}{U_{cd}}. \tag{27}$$

Then,

$$\sigma_{M} = \frac{K_n}{\sqrt{2}(f_{11} - f_1)} B \cdot 100\%. \tag{28}$$

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The density of probability of distortions is equal to

$$W(I_{PM}) = \frac{1}{\sigma_{PM}\sqrt{2\pi}} e^{-\frac{I_{PM}^2}{2\sigma_{PM}^2}}. \qquad (29)$$

It is evident that the comparative data for the RMS distortions during fluctuation interference in channels with PM and AM will be the same as for the maximum distortions during harmonic interference. Therefore,

$$\sigma_{AM} = 2\sigma_{PM}. \qquad (30)$$

We compare the derived result with the RMS distortions in the channel with FM. As our investigations (Bibl.2) showed, at a linear characteristic of the discriminator, the magnitude of the RMS distortions can be represented by the formula

$$\sigma_{PM} = \frac{K_n p}{\sqrt{6}(f_{II} - f_I)} \sqrt{1 + \frac{1}{\pi^2}\left(\ln\frac{p+1}{p-1}\right)^2} B \cdot 100\%, \qquad (31)$$

where $p = \dfrac{f_{II} - f_I}{2 f}$;

Δf is the frequency deviation;

K_n is the ratio of effective values of the voltages of interference to signal.

At optimum value of the factor p = 1.4, we have

$$\sigma_{FM} = 0{,}657 \frac{K_n}{f_{II} - f_I} B \cdot 100\%. \qquad (32)$$

Supplementary investigations showed that the real discriminator under its optimum parameters, without much changing the relative steepness of the pulse build-up, ensures a reduction of the effective value of interference, and consequently, also a lowering of the σ_{FM}. Besides, the value σ_{FM} will be approximately equal to

$$\sigma_{FM} = 0{,}5 \frac{K_n}{f_{II} - f_I} B \cdot 100\%. \qquad (33)$$

Combining the equations for σ_{PM} and σ_{FM}, we derive

$$\frac{\sigma_{PM}}{\sigma_{FM}} = \frac{1}{0{,}49\sqrt{2}} \approx 1{,}4. \qquad (34)$$

We thus derive the following comparative data for RMS distortions

$$\sigma_{AM} = 2\sigma_{PM} \approx 2{,}8\sigma_{FM}. \qquad (35)$$

As is evident, with respect to distortion resistance during fluctuation interference, the channel with PM (while having substantial advantages over the channel with AM) is inferior to the channel with FM.

A comparison of the resistance of the voice-frequency telegraph channel to fluctuation interference at various types of modulation will be more explicit if a check is made on the possibility of the appearance of distortions exceeding a value of $I_1 = \sigma_{AM}$.

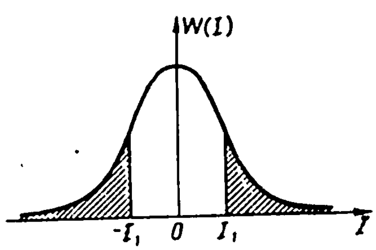

Fig.7

The probability of distortions, exceeding a given value at any type modulation, is equal to

$$P(I \geq I_1) = \frac{2}{\sigma\sqrt{2\pi}} \int_{I_1}^{\infty} e^{-\frac{I^2}{2\sigma^2}}\, dI \qquad (36)$$

(in Fig.7, the quantity P is defined by the hatched area).

We designate $\frac{I}{\sigma} = x$. Then, $dI = \sigma\, dx$ and

$$P(x \geq x_1) = \frac{2}{\sqrt{2\pi}} \int_{x_1}^{\infty} e^{-\frac{x^2}{2}}\, dx. \qquad (37)$$

We then use a known asymptotic expression

$$\int_{x_1}^{\infty} e^{-\frac{x^2}{2}} \approx \frac{1}{x_1} e^{-\frac{x_1^2}{2}}.$$

Besides,

$$P(I \geqslant I_1) = \frac{2}{\sqrt{2\pi}} \frac{\sigma}{I_1} e^{-\frac{I_1^2}{2\sigma^2}}. \qquad (38)$$

By determining the probability of the appearance of distortions, which exceed the RMS distortions in the channel with AM (σ_{AM}) n times, i.e., exceed the value $I_1 = n\sigma_{AM}$, we derive:

a) for the channel with AM

$$P_{AM}(I \geqslant n\sigma_{AM}) = \frac{2}{\sqrt{2\pi}} \frac{1}{n} e^{-\frac{n^2}{2}}; \qquad (39)$$

b) for the channel with FM

$$P_{FM}(I \geqslant n\sigma_{AM}) = \frac{2}{\sqrt{2\pi}} \frac{\sigma_{FM}}{n\sigma_{AM}} e^{-\frac{n^2\sigma^2_{AM}}{2\sigma^2_{FM}}}; \qquad (40)$$

c) for the channel with PM

$$P_{PM}(I \geqslant n\sigma_{AM}) = \frac{2}{\sqrt{2\pi}} \frac{\sigma_{PM}}{n\sigma_{AM}} e^{-\frac{n}{2}\frac{\sigma^2_{AM}}{\sigma^2_{FM}}}. \qquad (41)$$

At the same time, we have

$$\frac{P_{AM}}{P_{PM}} = \frac{\sigma_{AM}}{\sigma_{PM}} \left[e^{\frac{1}{2}\left(\frac{\sigma^2_{AM}}{\sigma^2_{PM}} - 1\right)} \right]^{n^2},$$

$$\frac{P_{AM}}{P_{FM}} = \frac{\sigma_{AM}}{\sigma_{FM}} \left[e^{\frac{1}{2}\left(\frac{\sigma^2_{AM}}{\sigma^2_{FM}} - 1\right)} \right]^{n^2},$$

$$\frac{P_{PM}}{P_{FM}} = \frac{\sigma_{PM}}{\sigma_{FM}} \left[e^{\frac{1}{2}\left(\frac{\sigma^2_{AM}}{\sigma^2_{FM}} - \frac{\sigma^2_{AM}}{\sigma^2_{PM}}\right)} \right]^{n^2}.$$

Substituting the previously cited values σ_{PM}, σ_{FM}, and σ_{AM} we derive:

$$\frac{P_{AM}}{P_{PM}} = 2\,(4{,}482)^{n^2}, \qquad (42)$$

$$\frac{P_{AM}}{P_{FM}} = 2{,}8\,(30{,}625)^{n^2}. \qquad (43)$$

$$\frac{P_{PM}}{P_{FM}} = 1{,}4\,(6{,}753)^{n^2}. \qquad (44)$$

Cited below are the values of the examined ratios for various n:

n	P_{AM}/P_{PM}	P_{PM}/P_{FM}	P_{AM}/P_{FM}
1	—	9,15	—
2	400	$2{,}91\cdot10^{3}$	$2{,}12\cdot10^{6}$
3	$1{,}434\cdot10^{4}$	$4{,}08\cdot10^{7}$	$6{,}22\cdot10^{13}$
4	$5{,}12\cdot10^{10}$	$2{,}65\cdot10^{13}$	$1{,}43\cdot10^{24}$

We note that the cited ratios of probabilities also determine the ratio of the average number of distortion surges per unit time, exceeding the given value σ_{AM}.

Distortions during Pulse Interference

As previously shown (Bibl.3), the maximum distortions of telegraphic pulse duration during pulse interference in channels with AM and FM, are determined by the following equations:

$$I_{pulse\,AM} = \frac{4U}{U_c} B \cdot 100\,\%, \qquad (45)$$

$$I_{pulse\,FM} = \frac{I_{pulse\,AM}}{3{,}5}. \qquad (46)$$

Since, with phase modulation, the steepness of the pulse build-up is twice as great as with AM, then

$$I_{pulse\,PM} = 0{,}5\,I_{pulse\,AM}. \qquad (47)$$

Thus, the comparative data for maximum distortions from pulse interference of

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equal magnitude, at various types of modulation can be represented in the form

$$I_{pulse\,AM} = 2\, I_{pulse\,PM} = 3{,}5\, I_{pulse\,FM}. \tag{48}$$

As is evident, the channel with FM, while being twice as stable as the channel with AM, relative to pulse interference, is considerably inferior to the channel with FM.

Conclusions

Thus, the comparative data of distortions during the action of various interferences in channels of voice-frequency telegraphy with PM, in comparison to channels with AM and FM, are represented in the following manner:

a) For maximum distortions during harmonic interference

$$I_{AM} = 2\, I_{PM} = 2\, I_{FM}.$$

b) For RMS distortions during fluctuation interference

$$\sigma_{AM} = 2\sigma_{PM} = 2{,}8\sigma_{FM}.$$

c) For maximum distortions during pulse interference

$$I_{AM} = 2\, I_{PM} = 3{,}5\, I_{FM}.$$

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PROBLEMS OF COMPUTING CIRCUITS OF T-FILTERS WITH OVERLAP

by

I.I.Petrov

It is shown that T-filters with overlap are a variation of M-derived circuits and can be designed by using the same conversion factor M. At the same time, all considerations remain valid concerning arrangement of the attenuation bump in the usual M-derived filter.

M-derived circuits are an extremely widespread type of circuits of electric filters. These circuits are convenient in that, while varying the magnitude of the conversion factor M from 0 to 1, it is possible to locate the attenuation bump of the filter in the interval of frequencies which should be suppressed as much as possible.

The so-called circuits of T-filters with overlap (Fig.1) are also known to possess properties of M-derived circuits. To the present time, however, T-filters with overlap are computed either by means of converting them to equivalent circuits (Bibl.1) or by means of determining the resistances of open-circuit conditions and the short-circuit of networks of chosen configuration.

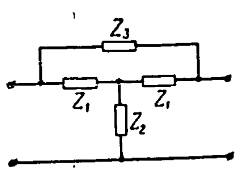

Fig.1

We will examine the usual conversion of a half-section of a filter circuit with constant K (the schematic in Fig.2a, where the resistances of the series leg Z_1 and of the parallel leg Z_2 are opposite in sign of reactance and where the condition $Z_1Z_2 = K^2$ is observed) into a series-derived half-section of the M type.

The condition of conversion is to keep the characteristic resistance of the circuit on the side of the terminals 1 - 2 constant while simultaneously varying the resistances of the series and the parallel legs.

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This condition is satisfied in the schematic shown in Fig.2b. The resistance of the open-circuit conditions in the converted network is

$$Z_{\infty 12} = MZ_1 + \frac{1-M^2}{M} Z_1 + \frac{Z_2}{M} = \frac{Z_1 + Z_2}{M}$$

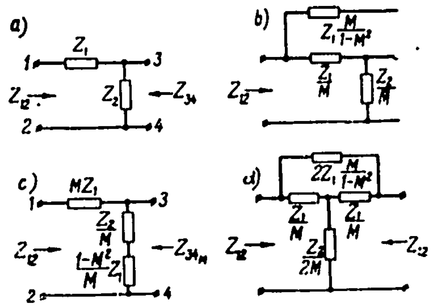

Fig.2

while the resistance of short-circuit is

$$Z_{012} = MZ_1.$$

Thus, the characteristic resistance on the side of the terminals 1 - 2 of the M-derived network is

$$Z_{12M} = \sqrt{\frac{Z_1 + Z_2}{M} MZ_1} = Z_{12}.$$

However, the circuit pictured in Fig.2c* also satisfies the same condition of keeping the characteristic resistance constant.

Here, the resistance of open-circuit conditions is

$$Z_{\infty 12} = \frac{Z_1 + Z_2}{M},$$

while the resistance of short-circuit is

*Strictly speaking, the schematic in Fig.2c permits only a determination of Z_{012} and $Z_{\infty 12}$, by means of which (on the basis of the theorem of bisection) the parameters of a section of a T-filter with overlap can be found. We can, however, formally assume that also schematic in Fig.2c can be connected from the right to a matched load – the second half-section with a load Z_{12}. In this case its input resistance on the side of the terminals 1 - 2 will be equal to the characteristic. Therefore, for convenience of discussion, we retain this inaccuracy and will relate the idea of characteristic resistance $Z = \sqrt{Z_0 Z_\infty}$ both to the circuit of a section of a T-filter with overlap and to the practically unrealizable circuit of a half-section.

$$Z_{012} = \frac{Z_1 \frac{M}{1-M^2} \frac{Z_1}{M}}{Z_1 \frac{M}{1-M^2} + \frac{Z_1}{M}} = MZ_1.$$

Consequently, the characteristic resistance of the circuit on the side of the terminals 1 - 2, just as in the preceding circuit, remains equal to the characteristic resistance of a filter with constant K.

This circuit cannot be realized in the form of a half-section, but two half-sections of it give a real, physically realizable circuit of a section of a T-filter with overlap (Fig.2d).

The conversion of a half-section of the filter K into a parallel-derived half-section of the M type is done under the condition of keeping the characteristic resistance of a half-section of the K type constant on the side of the terminals 3 - 4, at simultaneous variation of the resistances of the parallel and series legs of the half-section.

The condition is satisfied in the schematic pictured in Fig.3a. This is the

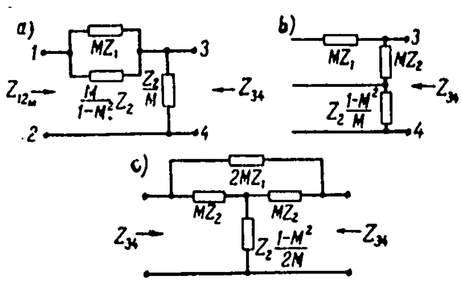

Fig.3

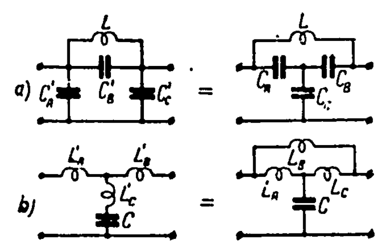

Fig.4

circuit of a parallel-derived half-section of the M type. It is easy to detect that the condition formulated above is satisfied also in the schematic of Fig.3b. The network in Fig.3b is also unrealizable in the form of a half-section, but two half-sections form the circuit of a T-filter with overlap (Fig.3c), which differs from the preceding circuit of such a filter in that the "overlapping" resistance is opposite in sign of reactance to the resistance of the series leg.

The developed circuits of T-filters with overlap, of the type of Figs.2d and 3c, are a consequence of M-derived conversion; their elements can be computed with a given factor M, and their properties by attenuation in the cutoff band are the same as in the sections of usual M-derived circuits, since they have identical expressions for the transmission constant.

As mentioned earlier, the circuit of a T-filter with overlap can, in certain cases, be developed by means of equivalent conversion of an M-derived circuit. Thus, the particular circuit of a parallel-derived filter of the M-type can be converted into the circuit of a T-filter with overlap by the method of conversion of the resistances connected according to a triangle circuit, into an equivalent connection according to a star circuit (Fig.4a), which was shown by T.Ye.Shi (Bibl.1). Acting in a similar manner it is possible to convert also the circuit of a series-derived section M into the circuit of a T-filter with overlap, after converting the star-connected resistances into an equivalent connection according to a triangle circuit (Fig.4b).

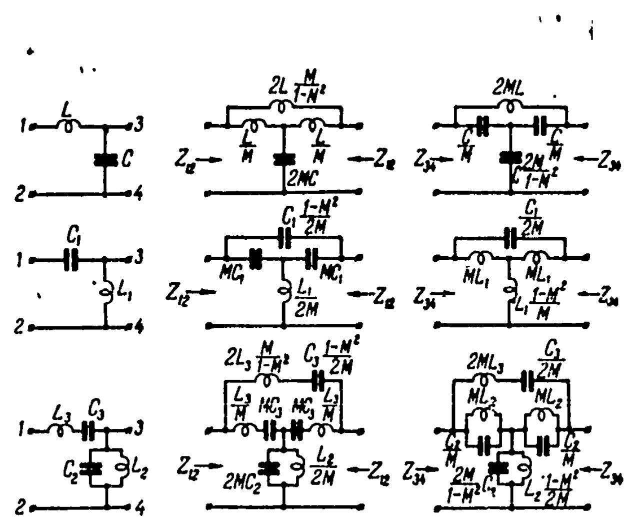

Fig.5

However, such conversions are convenient for filters of lower and higher frequencies, where reactances of identical sign enter the triangle or star connection and are inconvenient for the band filter, where Z_1 and Z_2 are complex and each consists of a parallel or series connection of elements of unlike reactances.

The proposed method of computing T-networks with overlap makes it possible to calculate them simply in all cases, since it reduces to conventional M-derived networks.

Shown in Fig.5 are circuits of initial half-sections of filters with constant K of lower and higher frequencies, and of a band-pass filter, and also corresponding M-derived sections of T-filters with overlap, with specifications for computed magnitudes of individual elements. Not quite clear is the application of the indicated method to the calculation of those T-circuits with overlap, which are M-derived with two coefficients: M_1 and M_2. As is known, in this case during the process of M-conversion of band filters, unlike factors are taken for elements with different sign of reactance, but under the condition that the input characteristic resistance and the frequency of the edges of the pass band of the filter remain invariable.

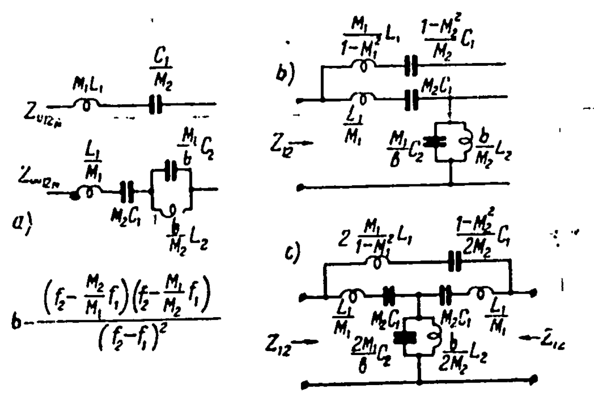

Fig.6

In application to a series-derived circuit of a half-section of a band-pass filter, these conditions lead to the form $Z_{\infty 12M}$ and Z_{012M}, shown in Fig.6a. We apply them in the same manner to the circuit of a T-filter with overlap.

Then the circuit of the half-section will look as pictured in Fig.6b, and the entire section will assume the form shown in Fig.6c. Here f_1 and f_2 are the edges of the pass band.

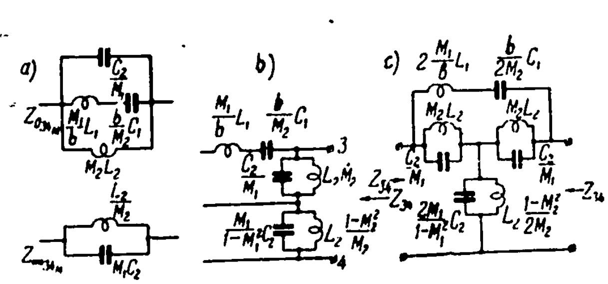

Fig.7

The resistances of the open-circuit conditions and the short-circuit of the parallel-derived network with two coefficients M_1 and M_2 on the side of the terminals 3 - 4 are shown in Fig.7a. For deriving such Z_∞ and Z_0 on the side of the terminals 3 - 4 in the T-network with overlap, it is necessary to have the circuit of a half-section,aas pictured in Fig.7b.

The complete circuit of a section is shown in Fig.7c.

Just as in the usual M-derived filters, the coefficients M_1 and M_2 in the filters in Figs.6c and 7c can be computed as a function of the desired arrangement of frequencies of an infinitely large attenuation $f_{1\infty}$ and $f_{2\infty}$ according to the following formulas:

$$M_1 = \frac{\frac{f_1 f_2}{f_{2\infty}^2} g + h}{1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}}, \quad M_2 = \frac{\frac{f_{1\infty}^2}{f_1 f_2} h + g}{1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}},$$

where

$$g = \sqrt{\left(1 - \frac{f_{1\infty}^2}{f_1^2}\right)\left(1 - \frac{f_{1\infty}^2}{f_2^2}\right)}, \quad h = \sqrt{\left(1 - \frac{f_1^2}{f_{2\infty}^2}\right)\left(1 - \frac{f_2^2}{f_{2\infty}^2}\right)}.$$

In conclusion I would like to express the hope that the proposed method of computing T-filters with overlap will permit more frequent use of this type as a kind of M-derived filter and will increase the possibility of varying the magnitudes of the elements when designing a filter according to a given attenuation characteristic.

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USE OF NONCONTACT ELEMENTS IN DIAL OFFICE CONTROL CIRCUITS

by

V.N.Roginskiy

In the present article circuit diagrams for the use of noncontact elements in the control circuits of ATS (automatic telephone systems) are given, and the principles of designing separate automatic office centers on their basis are discussed,

Introduction

Automatic telephone offices which started their development in the Eighteen-Nineties of the last century, completed the first stage of their development in the Nineteen-Twenties of this century. By that time, a large number of various dial office systems had appeared, based on selectors - electromechanical devices with relatively large moving parts and connection times of the order of tenths of a second. Both selectors and electromagnetic relays also found application in auxiliary circuits. However, the great organic defects of selectors were soon manifest: mainly the poor quality of contact and the frequent failure due to the presence of moving and wiping parts and large inertia.

One of the possible ways of eliminating the shortcomings of selectors was the replacement of selectors by low-inertia mechanisms - relays which are reliable in operation and possess good contact. The use of relays for creating systems of switching speech circuits, however, results in a very large number of relays per office, since the number of relays increases approximately proportional to the square of the office capacity. The relay offices, especially offices of large capacity, proved considerably more expensive and required larger premises than the dial offices with selectors. But for small offices, the use of relays is profitable, since it

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permits creating equipment which does not require constant attendance. In control circuits, the relays successfully replaced selectors in many cases.

At the same time, the development of relay circuits led to the creation of a new switching device, the crossbar connector (Bibl.1) which is a low-inertia device with light-weight moving parts, little movement, low operating time (order of milliseconds), and good contact of the relay type*. Beginning from the Nineteen-Thirties, dial offices with crossbar connectors came into increasingly wider use, and at present the majority of foreign firms have gone over or are going over to such systems. The use of such automatic offices permitted an approximate tenfold reduction in service personnel at dial offices.

At the same time, research was begun on the possibility of creating inertialess automatic telephone office systems, in which moving parts and mechanical contacts would be wholly absent. The extensive research work carried out in many countries of the world, showed that there is at present no fully reliable element which might replace mechanical contact in a speech circuit.

In control circuits, however, where different demands are made on contact than in speech circuits, nonmechanical elements have been used increasingly since the Nineteen-Thirties. The use of valves and gas-discharge tubes in relay circuits made it possible to simplify certain schematics. The introduction of electronic or ionic tubes in matching circuits permitted a simplification of selector sets, as in the American "universal" dial office and in the modernized "Rotary-7-E" system (Bibl.2).

In the Nineteen-Fifties an electronic director (Bibl.3) was created in England, which accomplishes recording of a six-digit number and sets up in exchange the first two signs (office code) to seven pulse trains. The last four signs (subscriber number) are transmitted without change. The two-year operation showed good service qualities of this director. Somewhat later in Belgium, the so-called "mechanoelec-

*Such a system is sometimes called a crossbar system.

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tronic telephone switching system" (Bibl.5) with crossbar switching and completely electronic control circuit was developed. Such an automatic office, with a capacity of 2000 numbers, has been operating in Oslo since 1954. In 1955, a thyratron director (Bibl.4) was introduced in experimental operation at one of the mechanical automatic offices in Leningrad. There are reports on various experimental systems of dial offices with noncontact elements in the control circuits (Bibl.6) and also on the use of these elements in individual automatic telephone office centers (Bibl.7-9).

In the present article, certain examples of the design of automatic office centers with noncontact elements are briefly discussed.

Circuits of Correspondence

In the mechanical automatic dial systems, one of the most complex instruments is the director, and in particular, the circuit controlling the movement of selector brushes and the setup of connection. We will discuss two examples of modernization of circuits of control and correspondence.

The "Universal" dial system (USA) is based on a multipotential system of control by a mechanical selector which has one rectilinear movement.

The director in this system consists of mechanical registers fixing the dialed number, of the starting relay, and the electronic circuit of correspondence. The numeral fixed by the registers is noted by feeding a definite voltage (from -40 to +40 v in stages of four volts each) to the circuit of correspondence. The corresponding outputs of the selector are marked by the same voltages fed to the lamellae of the brushes (c). When the pertinent sign has been dialed and the director connected with the selector, the circuit of the selector clutch magnet (E) in the director circuit (Fig.1) is closed, and the brushes begin to move. So long as the voltages fed from the director and the selector differ, current will flow in the plate circuit of the tubes $L_1 - L_2$. As soon as these voltages become level, the current in the plate circuit is reduced, causing the gas-discharge tube (T) to ignite and the relay (P) to operate, breaking the starting circuit.

In the new system 7-E ("Rotary") (Bibl.2), multiphase control is used for setting up correspondence. The dial office contains a generator of 450-cycle current, having 12 outputs with the current-phase shift of neighboring outputs at 30°. Thus, there are 12 different phases (φ_1, φ_2..., φ_{12}), fed to the contacts of the registers and the selectors.

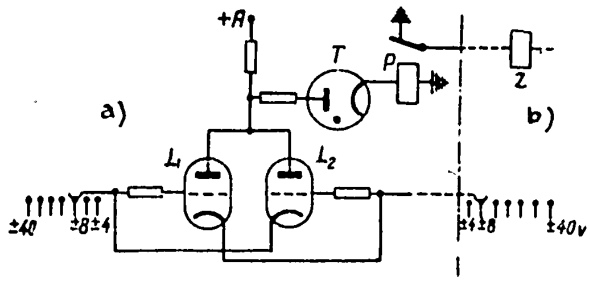

Fig.1

a) Director; b) Selector

The director for setting up correspondence contains a circuit with two three-electrode ionic tubes of special design (Fig.2), in which the ignition voltage between the electrodes (1) and (2) is equal to 70 - 85 v, while it is more than 200 v between the central electrode (3) and the outside electrodes (1) or (2). After ignition, burning is maintained by a potential of 85 v.

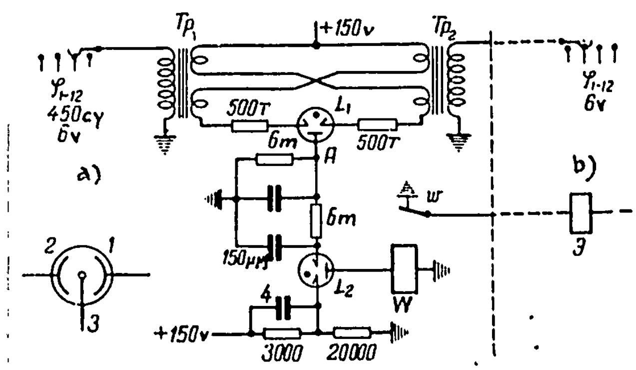

Fig.2

a) Director; b) Selector; c) Brush

For setting up connection in the director through one of the speaking wires, the starting circuit is closed of the selector clutch magnet E, and the selector brushes are set in motion.

So long as the phases of the voltages in the transformers Tr_1 and Tr_2 do not coincide, the alternating current ignites the tube L_1 and the potential of the point A will be close to the voltage of the battery (+150 v). However, when a free output of the sought direction is found, the phases of both transformers become identical and the tube L_1 is extinguished. At the same time, the potential of the point A is lowered, the tube L_2 is ignited and the relay W is tripped, breaking the starting circuit. The elements of the circuit are so selected that the operating time of the relay W, from the moment of closing the circuit by the selector pilot brush, amounts to about 2 milliseconds.

In both cited cases, the use of electronics in the director made it possible to create by simple means an electrical subdivision of the field of selection and to simplify the selector circuit to the maximum. In particular, the function of the test has passed from selector to director. In the 7-E system, the group selectors have no relay at all.

Registering the Number

From the moment the automatic telephone system developed to its present status, the transmission of the number from the subscriber to the dial office has been accomplished in the form of separate pulse trains, with the number of pulses in a train corresponding to the dialed number.

For transmission of the number, the equipment contains a rather complex device, the dial. The pulse method of transmitting the number limits the length of the subscriber line, and the dialing itself of the number proceeds comparatively slowly. Therefore, attempts have repeatedly been made to replace the dial by keys and accomplish the transmission of the number by combinations of frequencies, coded pulses, variations in current strength, and other methods; this problem however has not yet found a practical solution. It should be noted that the possibility of transmitting the number by voice is not a fantasy.

Modern mechanical and crossbar (and also certain step-by-step) automatic telephone exchanges require registration of the number being transmitted. Noncontact elements are very convenient for creating registering devices, and at present many different schematics of such devices exist. We will discuss some of these, which provide for registration and fixation of the number transmitted by means of a dial. For this purpose, the dial office director must contain a register device and a corresponding circuit of division of separate number digits (distributor).

For fixation of separate digits of the number one can have either several registers (similar to what occurs in the director of the mechanical automatic telephone offices of the "Krasnaya Zarya" - Red Dawn - factory), or have one register and several simpler fixators (similar to the register relay).

The majority of such circuits used at present are designed in semiconductor and gas-discharge diodes and triodes. The use of electron tubes was rejected because of their wastefulness (high current consumption and short life).

Attempts to create and apply various multipositional tubes: electron-ray tubes, multicathode gas-discharge tubes (Bibl.8), trochotrons - multianode tubes with electron-ray (Bibl.10) and so forth - have not as yet been fully successful. Recently,

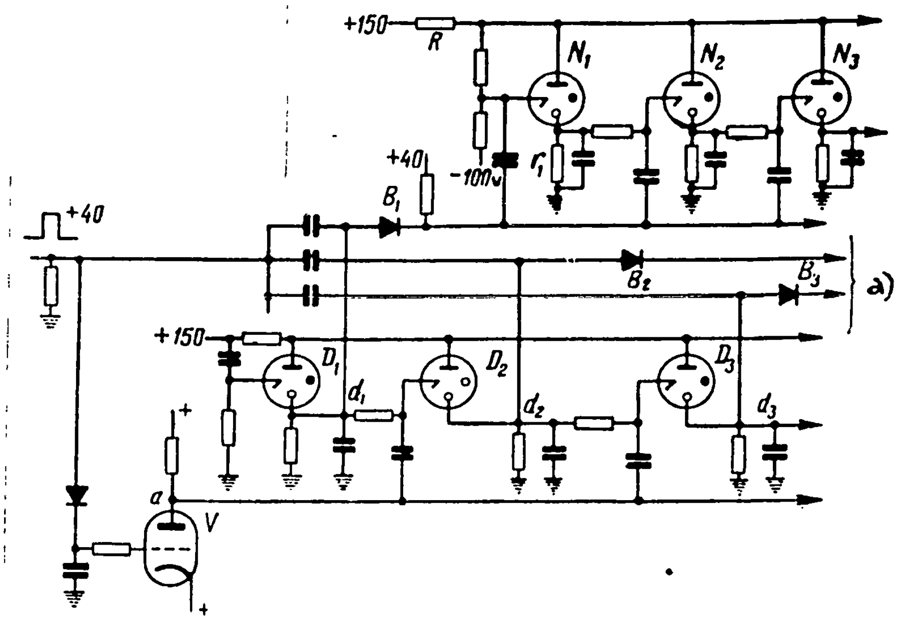

Fig.3

a) To other counting chains

for the purposes indicated above, wide use is being made of magnetic elements with rectangular hysteresis loops, and of capacitors.

We will examine several examples. Pictured in Fig.3 is the circuit of the registering part of the director with ionic tubes (Bibl.7). The counting chain for each digit consists of the tubes N_1, N_2, ... The analogous chain of tubes D_1, D_2, ... plays the role of distributor, while tube V acts as a train relay.

In the state of rest, the tube D_1 is burning, so that the point d_1 has a positive potential and the valve B_1 will be open. The remaining valves B_2, B_3, ... will be closed. The tubes N_1, N_2, ... do not burn, the voltage in the firing electrode of the first tube being somewhat higher than in the others. On arrival of the positive pulse at the input, the tube N_1 is ignited; because of the voltage drop in the resistance r_1 this causes the voltage in the firing electrode of the tube N_2 to rise. At the end of the pulse, the tube N_1 remains in the ignited state. At the next pulse, the tube N_2 is ignited and in consequence of the increase of the voltage drop in the resistance R and the action of the charge of the capacitor in the cathode circuit, the tube N_1 is extinguished. At the next pulse, the tube N_3 is ignited and the tube N_2 extinguished, etc. Simultaneously with the arrival of the first pulse, the tube V is opened, causing the voltage in its anode to drop. In the intervals between pulses, the tube V does not have time to be closed, since sufficient voltage is retained in the grid circuit of the capacitor.

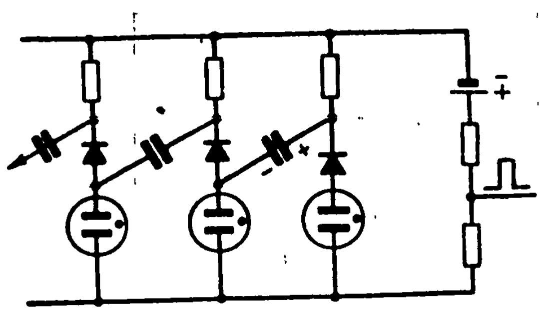

Fig.4

During a long interval between pulse trains, the capacitor in the grid circuit of the tube V is discharged and it is closed. Owing to this the voltage at the point a rises, which forces the tube D_2 to ignite and the tube D_1 to extinguish. This causes the valve B_1 to close, and B_2 to open. The pulses of the next train will now arrive at the second counting chain, etc. The dialed number is thus fixed by the burning of the corresponding tubes in the counting chains. A recording device of such type was applied in an experimental director in England (Bibl.3) and in an ex-

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perimental thyratron director operating in Leningrad (Bibl.4) since 1955.

The schematic of the counting chain (with gas-discharge tubes and valves) is shown in Fig.4. At a negative pulse, one of the tubes is extinguished and the next ignited, due to the charge of the capacitor between the tubes.

The register, whose schematic is shown in Fig.5, is designed on a different principle (Bibl.9). The circuit consists of four triggers of two ionic relays (gas-discharge tubes) X and Y in each. In each pair, normally one of the tubes is burning. With the arrival of a positive pulse in the conductor connected with the firing electrodes of both tubes, the second tube is ignited and the first extinguished. Thus, if in the state of rest the tubes X are burning, the arrival of the first pulse in the first pair of tubes ignites the tube Y_1 and extinguishes the tube X_1. On arrival of further pulses, the tube Y_1 will ignite after odd pulses and the tube X_1 after even pulses. At the same time, at the instant of ignition of the tube X_1, the voltage rise in the cathode of this tube and the charge of the capacitor cause the gas-discharge two-electron

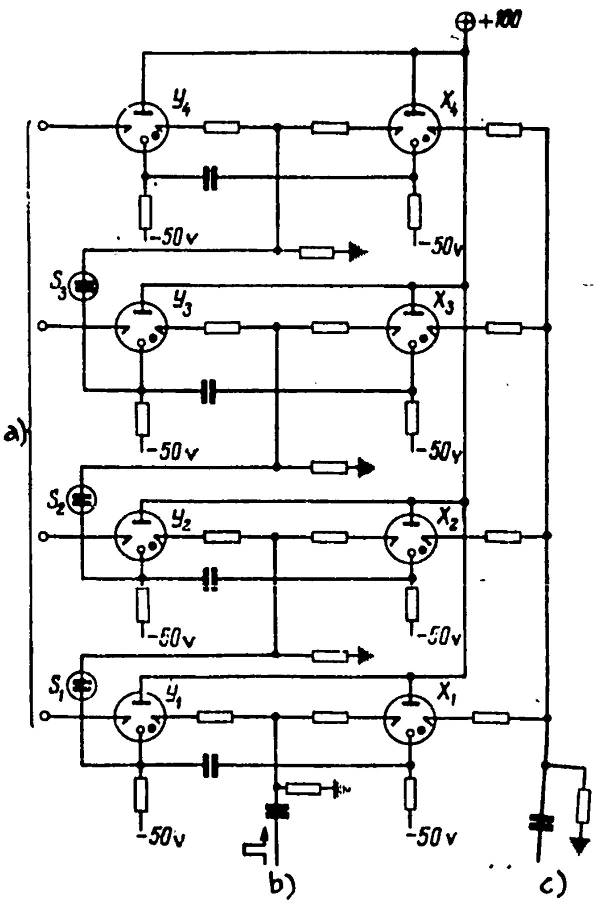

Fig.5

a) To retainers; b) Pulses of dialing;

c) Initial pulse

tube S_1 to ignite briefly and the pulse to be transmitted by the second pair of the tubes $X_2 - Y_2$. After the fourth pulse, the trigger $X_3 - Y_3$ is brought into action, etc. Thus, to each digit corresponds a combination of ignited tubes Y. In all,

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such a register can count 15 pulses.

The retainer consists of four two-electrode gas-discharge tubes. At the end of the pulse train, the tube of the first retainer is connected for a short time to the Y tubes of the register, across the distributor. At the same time, the corresponding tubes are ignited in the retainer. After that, the "initial pulse" is given in the register, causing all X tubes to ignite; the register is then ready for reception of the new train.

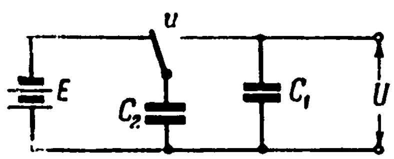

Fig.6

A similar device serves also for transmission of retained pulses.

Figure 6 shows the simplest circuit of metering pulses by means of the capacitor C_1 which, at every pulse, receives across the contact u of the pulse relay a booster charge from the capacitor C_2 or lower capacity. With proper selection of the

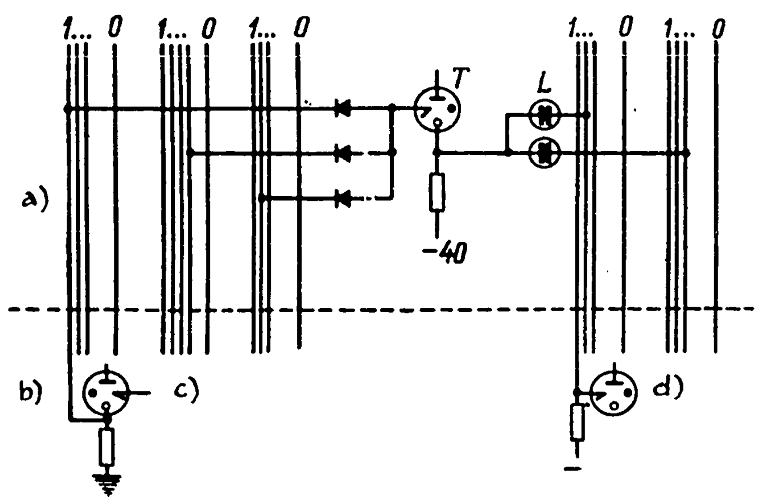

Fig.7

a) Translator; b) Register; c) Retainers of register; d) Retainer of translator

circuit parameters, the voltages U in the capacitor C_1 will be approximately proportional to the number of received pulses. Such a circuit can, for example, replace

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the selectors in the director of a multipotential system, which was discussed above.

There are apparently good prospects for the use in telephony of magnetic cores with rectangular hysteresis loops (Bibl.11,12), which have found wide application in computers and certain other fields. These elements make it possible to create very compact and economical circuits.

In certain cases the translation of a retained number becomes the function of the register. Such a recount can be accomplished by the means employed for counting and fixation. The simplest circuit of translating a three-digit number to a two-digit one is shown in Fig.7 (Bibl.7). For each translated number, the translator contains a thyratron (T) in which the firing electrode is connected across valves with the corresponding outputs of the register retainers, and the cathode is connected with the firing electrodes of the retainers of translated digits across gas-discharge tubes (L). The drawing shows only one circuit for the translation of the numbers 142 to 23. The thyratron T is unlocked only in the case in which the number 142 is retained. In consequence of the rise in cathode potential, the gas-discharge tubes L are ignited, the digits 2 and 3 will be fixed in the retainer of the translator.

Inasmuch as the translation occurs within a very brief interval of time, the translation circuit can be used for several registers connected in series. In the electronic director, already mentioned above (Bibl.3), a translator common to all registers is employed, making up to 48 translations in 600 milliseconds.

Noncontact Elements in Relay Circuits and Selectors

The use of noncontact elements, and primarily of valves, in relay circuits has in many cases permitted considerable simplification of the circuits. Thus, M.A.Gavrilov (Bibl.13) showed that the use of rectifier elements affords an opportunity to reduce the number of contacts in the relay circuit to one change-over contact for each contact of the relay, and gave the theoretical method of the synthesis of such

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circuits.

We will discuss several circuit diagrams for the application of nonlinear elements in separate automatic office centers.

Represented in Fig.8 is a circuit of conversion of dialed pulses, transmitted by

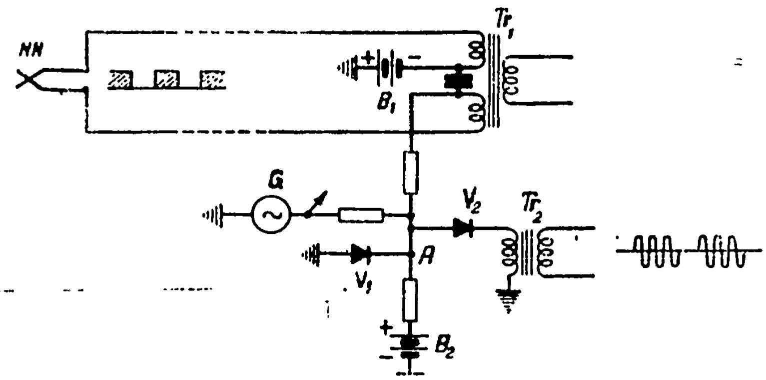

Fig.8

break of a DC circuit, into pulses of alternating current. So long as the DC circuit in the contact nn of the dial is closed, the point A has a negative potential relative to the ground (on account of the battery B_1), so that the valve V_1 is open and the valve V_2 is closed. When the circuit in the contact nn is broken, the voltage at the point A is increased because of the voltage of the battery B_1, and the valve V_1 is closed while V_2 is opened. At the same time, alternating current from the generator G is supplied to the winding of the transformer Tr_2. Such an arrangement is used, for example, in certain systems of voice-frequency dialing.

Figure 9 gives a circuit of crossing from a two-wire subscriber line to a four-wire circuit in which the number and the call are transmitted by a current of ultrasonic frequency. The transformer Tr_2 and the valves V_1 and V_2 serve for the conversion of dialed pulses into high-frequency current. The call current of high frequency, arriving from the side of the four-wire line, is separated by the high-frequency receiving filter and fires the thyratron (L), through which alternating current with a frequency of 25 cps begins to flow, Through the transformer Tr_3, this current en-

-ters the subscriber line.

The use of rectifier elements and a thyratron in a dial office of the 7-E type ("Rotary") (Bibl.2) made it possible in a new way to solve the circuit (Fig.10) of

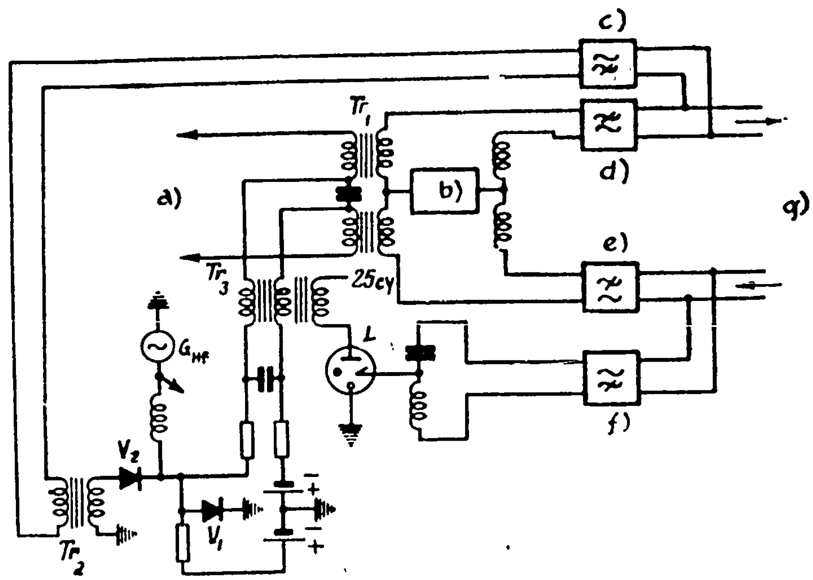

Fig.9

a) To subscriber; b) Balance; c) High-frequency filter - transmission; d) Low-frequency filter - transmission; e) Low-frequency filter - reception; f) High-frequency filter - reception; g) Channel

the subscriber-line equipment and the triggering of the finder switches (IV). The subscriber-line equipment contains two valves V_1 and V_2 and a set of three resistances R_1, R_2, and R_3.

The valve V_1 connects the subscriber-line equipment with the triggering device, common for 50 lines. From this device, a potentiometer feeds a counter voltage of the order of -36 v. When the subscriber circuit is open, the voltage at point A is equal to -48 v, and both valves are closed. When the subscriber circuit is closed, the voltage at point A rises to -24 v, thus opening the valve V_1 in the subscriber-

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line equipment and the valves V_3 and V_4 in the common device. An alternating current with a frequency of 450 cps (voltage, about 2.5 v) will start flowing through the

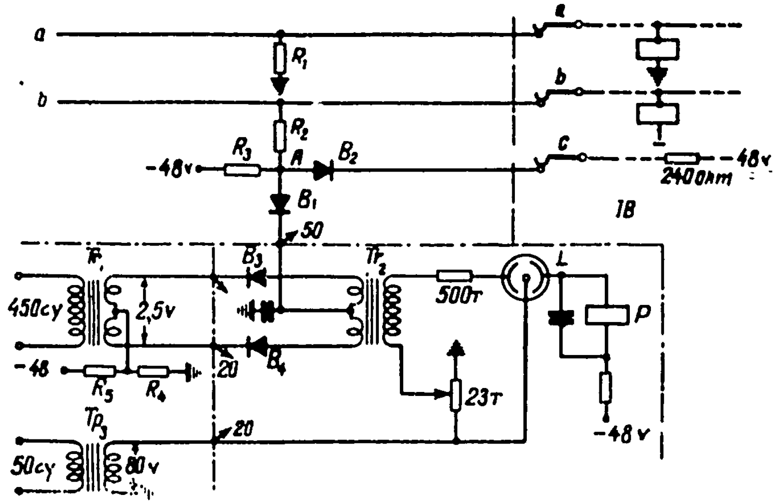

Fig.10

valves V_3 and V_4. At the firing electrode of the thyratron L, this current creates a corresponding potential, the thyratron is ignited and begins to pass ripple current, created by superimposing the battery voltage with an alternating voltage of 80 v of a frequency of 50 cycles. This trips the relay P, closing the trigger circuits of the finder switches (IV). When the brushes of any finder switch contact the line of the calling subscriber through the valve V_2, the test relay in the switch IV operates and the finder stops. In addition, the potential of point A is again lowered to -48v due to voltage being fed from IV across the resistance of 240 ohms. This causes the valves V_1, V_3 and V_4 to close, and the tube L to extinguish at the instant of a negative half-wave of alternating current.

Such a system, at a 10,000 subscriber office, made it possible to reduce the total number of relays by approximately 27%. Structurally, the subscriber-line equipment comprises a small mounting plate 55 × 50 × 8.5 mm, installed in the main switchboard. The resistances R_1 and R_2 of 15 k-ohm each, permanently connected to the speech circuit, have practically no effect on the quality of the call, since the

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attenuation introduced by them amounts to 0.01 - 0.02 neper in all.

In the mechanoelectronic telephone switching system (Bibl.5) the directing transmission inside the office is accomplished by means of pulses shifted in time. Since a detailed discussion of the entire control circuit would go beyond the scope of this article, we will examine only the circuit generating the distinguishing pulse which serves for determining the number of a calling subscriber in a hundred (Fig.11). At the office there is a common pulse generator which creates 14 pulse trains of varied duration and shifted in time, as shown in Fig.12. The pulses of trains N_a and N_c have a duration of 0.2 msec; the pulses N_a follow each other with an interval of 4 pulses, i.e., after 1 msec; and the pulses N_c follow with an interval of 3 pulses, i.e., after 0.8 msec. The pulses of train N_b have a duration of 1 msec.

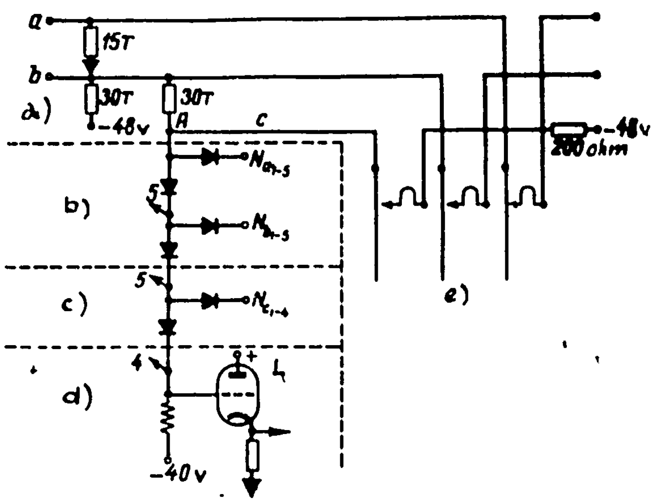

Fig.11

a) Line circuit; b) On 5 lines; c) On 25 lines; d) On 100 lines; e) Selector

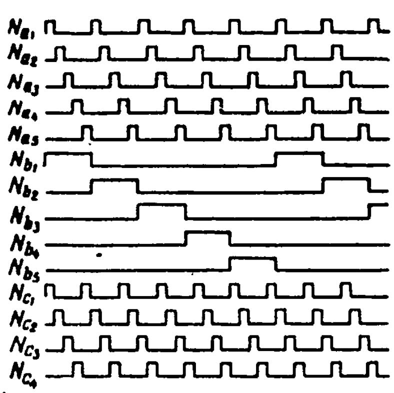

Fig.12

As long as the subscriber circuit is open, the potential of point A (Fig.11) will be -48 v and all valves connected to this point, will be closed. The resistances of the circuit and the voltages are so selected that the grid of the cathode follower L has a trigger voltage only in the case when the subscriber circuit is closed (the potential of point A rises approximately to -16 v) and positive pulses are fed to all points N_a, N_b, and N_c. Therefore, after the calling subscriber has lifted

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the receiver, a pulse appears at the output of the cathode follower at the instant of coincidence in time of the pulses of all three trains. Thus, depending on which of the subscribers lifted the receiver, the created pulse will occupy a definite position in time, appropriated to the given subscriber. Deciphering is done by a special tube circuit, connected with the same pulse generator.

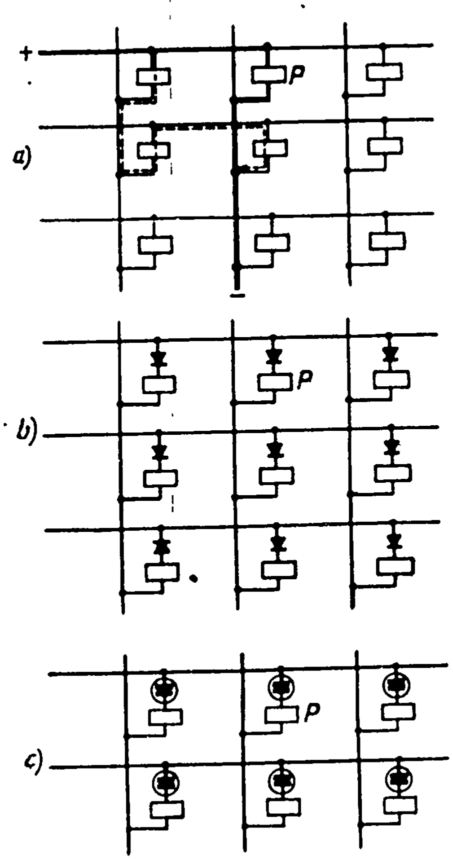

Fig.13

The nonlinearity of certain elements permits creating a decoupling and blocking circuit in relay and crossbar switches. In simplified form, such a selector can be pictured as a crossbar exchange (Fig.13) where relays are placed at the intersections of "horizontals" and "verticals". For establishing connection, a certain relay, located at the intersection of the "horizontal" and the "vertical" to which voltages have been fed must operate. If appropriate measures are not taken, a number of parasitic circuits appear (for example, the broken line in Fig.13a) and a simultaneous operation of several relays becomes possible. Elimination of this defect purely by relay means results in considerable complication of the circuit. But the introduction into the circuit of each relay of a valve or gas-discharge tube (Fig.13b, c) completely eliminates this shortcoming. In particular, such a method of employing a gas-discharge tube was used in the system of selectors of an experimental automatic office with electronic directing, developed by Bell Laboratories (Bibl.6).

The circuit in Fig.14 has the characteristic that, at a definite correlation between the magnitudes of the resistances of relays 0 and the general resistance R, the voltage feed will ignite only one of the gas-discharge tubes and trip only one of the relays O. This property of gas-discharge tubes is used for trunk hunting, in particular in the already mentioned experimental automatic switching system of the Bell Laboratories (Bibl.6). A combination of the characteristics of valves and gas-discharge tubes permits the selection of a free line in the wanted direction through several selection steps.

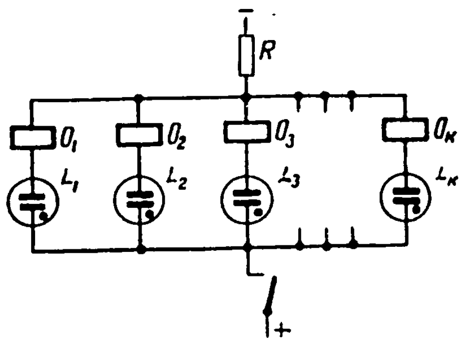

Fig.14

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BALANCING CIRCUITS OF COMMON BATTERY TELEPHONE SETS WITH SEMICONDUCTOR AMPLIFIERS

by

A.S.Sadovskiy

All kinds of variations of balancing circuits of common battery telephone sets with transistors are discussed, formulas are given for design of optimum parameters, and the most rational circuit of such a set is selected. Experimentally derived frequency characteristics of the sidetone of the set are cited.

Equivalent Circuits of Common Battery Telephone Sets with Transistors

In the selection, from a number of possible variations, the rational circuit of a telephone set with transistors and also in determining the basic correlations for computing its parameters, it is convenient to substitute the set circuits by equivalent circuits of transmission and reception. An example of such a substitution for one of the variations of balancing circuits was cited in a preceding article by the author (see Bibl.1, Figs.2b, 2c, 3a and 4).

We will examine circuits of sets connected into a subscriber line of limit length, so that the resistance R_d (see Bibl.1, Fig.3) which regulates the incoming constant current of set feed, can be assumed as equal to zero. With feed through the balance line Z_b, the latter must contain the parallel connection of the resistance R_b and the capacitance C_b (see Bibl.1, Fig.3a); therefore, we will call this line a parallel balance line. To prevent direct current from flowing through the telephone, a capacitor C_T is connected in series with the latter.

On the basis of the foregoing, a series of possible variations of balancing circuits of telephone sets with electromagnetic microphones and single-cycle transis-

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tors can, in transmission and reception, be represented by the equivalent circuits 1 – 6 (Fig.1). Here R_g is the output resistance of the transistor, which plays the role of internal resistance of the equivalent generator with the emf E_n, which replaces the amplifier.

The derivation of the basic correlations and design formulas are cited below for circuit 1. For the remaining circuits (Fig.1, circuits 2 – 6) the similarly found rating formulas are given in Table 1.

For all circuits of Fig.1 we assume an ideal transformer and a voltage ratio equal to

$$n_1 = \frac{w_1}{w_2} \text{ and } n_2 = \frac{w_3}{w_2}. \tag{1}$$

The sidetone-reduction condition of the set at one frequency is the equation:

$$Z_b = MZ_l = M|Z_l|e^{i\varphi_l}, \tag{2}$$

where Z_l is the input resistance of the line, equal to the wave impedance;

φ_l is its angle;

Z_b is the total resistance of the balance line;

M is the numerical factor which determines the voltage ratios of the differential transformer of the set.

Elements of the parallel balance line are determined from the following correlations:

$$R_b = \frac{M|Z_l|}{\cos\varphi_l} \text{ and } C_b = -\frac{1}{\omega M|Z_l|}\sin\varphi_l. \tag{3}$$

In the circuit fed across the balance line Z_b, the active resistance R_b of the balance line determines the input resistance of the set to the direct current r_{in} (here we disregard the active resistances of the transformer windings II and III); therefore the magnitude $R_b \approx r_{in}$ cannot be taken arbitrarily. In subscriber lines of limit length, we have $r_{in} \leq 300$ ohms according to existing standards; however, it

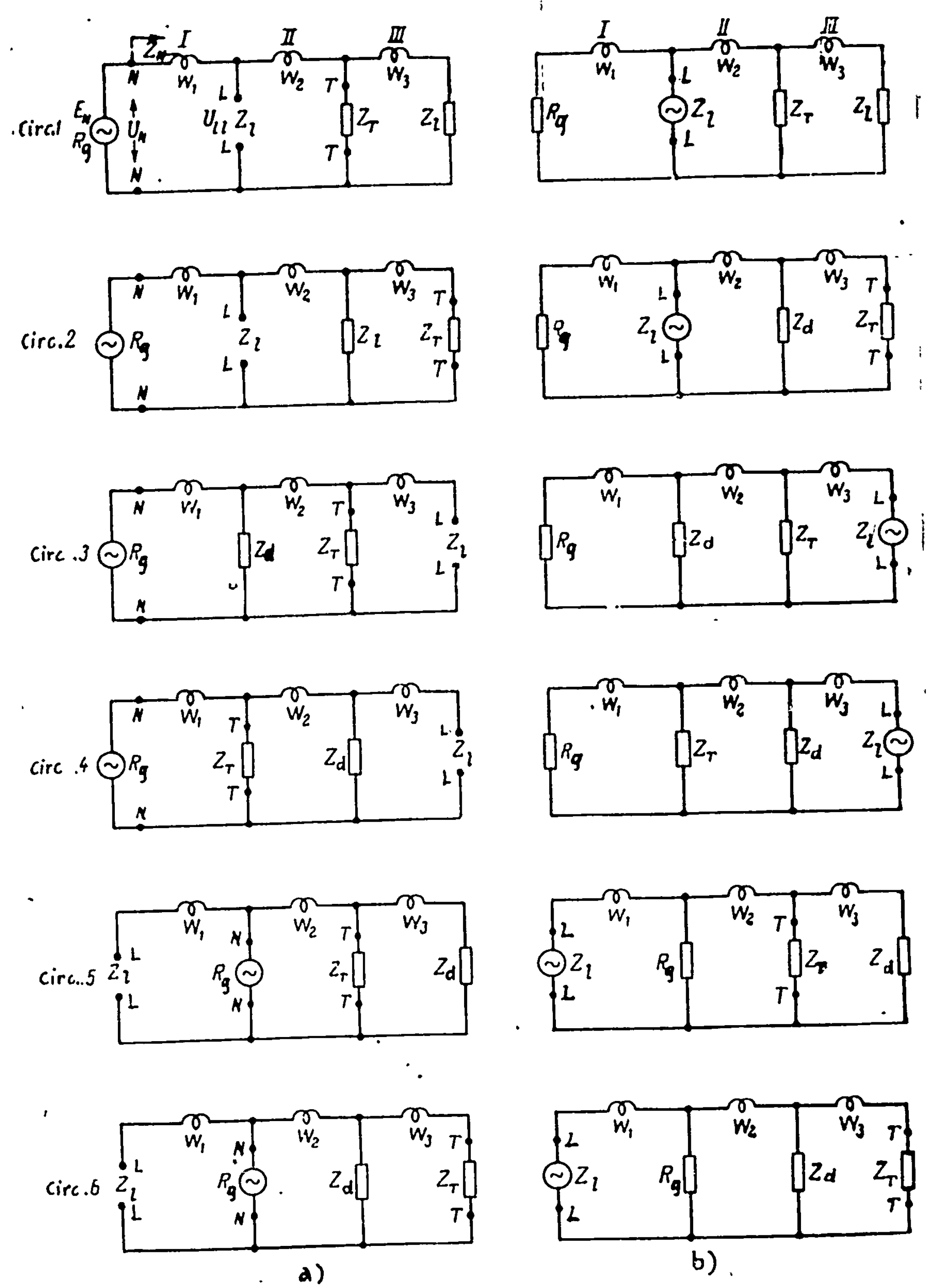

Fig.1

a) Transmission; b) Reception

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is rational to increase it in sets with transistors.

Basic Correlations and Optimum Parameters of Circuit 1

Complete sidetone-reduction at one frequency means that the attenuation of the sidetone effect in transmission must be $b_{st} = \infty$. This leads to the condition

$$\frac{Z_b}{Z_l} = M = \frac{n_2(1+n_1+n_2)}{n_1}. \tag{4}$$

Under this condition, the input resistance of the circuit Z_n at the points n - n during transmission is determined by the expression:

$$\frac{Z_n}{Z_l} = N = \frac{(1+n_1+n_2)(1+n_1)}{1+n_2} \tag{5}$$

and

$$\frac{U_{ll}}{U_n} = \frac{1}{1+n_1}, \tag{6}$$

where U_{ll} and U_n are voltages developed during transmission in the circuit 1a (Fig.1), at the points $l - l$ and n - n.

The operating attenuations during transmission b_{tr} and during reception b_{re} can be found in the following form:

$$b_{tr} = \ln\left|\frac{(1+n_2)\sqrt{\frac{R_r}{|Z_l|}}\, e^{-i\frac{\varphi_l}{2}} + (1+n_1)(1+n_1+n_2)\sqrt{\frac{|Z_l|}{R_r}}\, e^{i\frac{\varphi_l}{2}}}{2(1+n_1+n_2)}\right| \tag{7}$$

and

$$b_{re} = \ln\left|\frac{n_1(1+n_2)\sqrt{\left|\frac{Z_T}{Z_l}\right|}\, e^{i\frac{\varphi_T-\varphi_l}{2}} + n_2(1+n_1)\sqrt{\left|\frac{Z_l}{Z_T}\right|}\, e^{-i\frac{\varphi_T-\varphi_l}{2}}}{2n_1n_2}\right|, \tag{8}$$

where Z_T is the total resistance of the telephone;

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φ_T is its angle.

The greatest allowable increase of the undistorted level* of transmission p_0 of the set with transistors, in comparison with the usual set with carbon microphone, is determined (Bibl.1) by the expression

$$p_0 = \ln\left|\frac{U_{ll}}{U_l}\right| = \ln\left|\frac{U_n}{U_l}\frac{1}{1+n_1}\right| < \ln\left|\frac{U_{0k}}{U_l(1+n_1)}\right|. \qquad (9)$$

Here U_{ll} and U_l are voltages of speaking currents developed in the line during transmission at the points $l - l$, respectively, of the set with transistors (circuit 1a, Fig.1) and of the conventional set with carbon microphone;

U_{ok} is the voltage feed in the collector of the transistor.

The condition for obtaining the lowest possible operating attenuation of the circuit transmission b_{tr} is equality of the internal resistance of the generator R_g and of the input resistance of this quadripole $|Z_n|$ or matching during transmission, i.e.,

$$\frac{R_G}{|Z_l|} = \frac{|Z_n|}{|Z_l|} = N. \qquad (10)$$

Taking into account the value N from eq.(5) and substituting it in eq.(7), will give

$$b_{tr} = \ln\left|\sqrt{\frac{(1+n_1)(1+n_2)}{1+n_1+n_2}}\cos\frac{\varphi_l}{2}\right|. \qquad (11)$$

The optimum magnitude of the modulus of total resistance of the telephone $|Z_T|$, which gives the least attenuation b_{re}, and consequently, also the maximum of current in the telephone during reception, can be found if the derivative of b_{re}, assuming $|Z_T|$ to be variable, is equated to zero. Then,

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*Here, distortions are meant that arise when the alternating voltage at the output of a triode exceeds the direct voltage feed of the triode collector.

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$$\frac{|Z_T|_{opt}}{Z_l} = \frac{n_2 (1 + n_1)}{n_1 (1 + n_2)} = T. \tag{12}$$

Substituting in eq.(8) the derived value $|Z_T|_{opt}$, we find the value of the operating attenuation during reception

$$b_{re} = \ln \left| \sqrt{\frac{(1 + n_1)(1 + n_2)}{n_1 n_2}} \cos \frac{\varphi_T - \varphi_l}{2} \right|. \tag{13}$$

The total value of the operating attenuation of the set during reception and transmission b_{set} characterizes the circuit of the set as a whole. This value is equal to

$$b_{set} = b_{tr} + b_{re} = \ln \left| \frac{(1 + n_1)(1 + n_2)}{\sqrt{(1 + n_1 + n_2) n_1 n_2}} \right| + \ln \left| \cos \frac{\varphi_l}{2} \cos \frac{\varphi_T - \varphi_l}{2} \right| = b_n + b_\varphi. \tag{14}$$

As is evident from the derivation, the first component of the set attenuation b_n depends on the magnitude of the voltage ratios n_1 and n_2 of the transformer and their correlations. The second component b_φ depends on the magnitudes of the angles φ_l and φ_T.

By selecting the optimum ratio of n_1 to n_2, the minimum value of the first attenuation component b_n can be obtained. For determining this ratio, a derivative of b_n according to n_2 should be taken and equated to zero. Then,

$$n_2 = \frac{n_1 + 1}{n_1 - 1} \tag{15}$$

and, consequently

$$b_{set} = b_{n\,min} + b_\varphi = \ln 2 + \ln \left| \cos \frac{\varphi_l}{2} \cos \frac{\varphi_T - \varphi_l}{2} \right|. \tag{16}$$

Substituting the derived value n_2 in eq.(5) of the input resistance Z_n during transmission, we get

$$N = \frac{Z_n}{Z_l} = \frac{1}{2}\,(1 + n_1)^2, \qquad (17)$$

from which we find

$$n_1 = \pm\sqrt{2N} - 1. \qquad (18)$$

Considering the optimum correlation (15) between n_1 and n_2, and also eq.(18), we find the condition of sidetone-reduction, i.e., M from eq.(4), $\frac{U_{ll}}{U_n}$ from eq.(6), and T from eq.(12). Then,

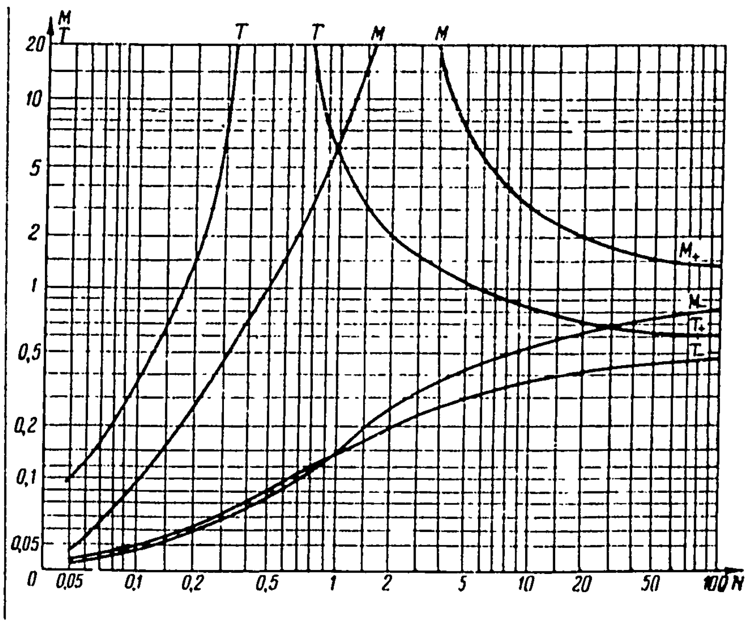

Fig.2

$$M_{\pm} = \frac{N}{(-\sqrt{2} \pm \sqrt{N})^2}, \qquad (19)$$

$$\frac{U_{ll}}{U_n} = \frac{1}{\sqrt{2N}}, \qquad (20)$$

$$T_{\pm} = \frac{N}{(-1 \pm \sqrt{2N})^2}. \qquad (21)$$

Here, M and T each can have two values, depending on the choice of the plus or the minus sign.

Plotted in Fig.2 are curves for the dependences of $M_{\pm}$ and $T_{\pm}$ on N, constructed according to eqs.(19) and (21).

From Fig.2 it is evident that always $N_- < 1$, and $M_+ < 1$ when $N < 0.5$, and $M_+ > 1$ when $N > 0.5$. Since the transistor of the set has an output resistance of $R_g \gg Z$ and since $|Z_n|$ must be equal to or, at any rate, approximate to R_g, we will always have $N \gg 0.5$ in such sets. The quantity M determines the value R_b which determines the input resistance r_{in} of the set to direct current.

When operating in actual automatic telephone networks, the input resistance r_{in}

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of the set with transistors must be approximate to the input resistance of existing sets (300 ohms). However, for increasing the voltage feeding the triode of the set, it is desirable to increase the input resistance to a magnitude of the order of 400 to 600 ohms. For deriving such values r_{in}, the quantity M_-, which has the least of two possible values, should be selected.

The greatest allowable increase of the undistorted level of transmission p_o, with optimum parameters in comparison with the usual sets, can be found from eq.(9):

$$p_0 = \ln\left|\frac{U_n}{U_l}\frac{1}{\sqrt{2N}}\right| < \ln\left|\frac{U_{0\kappa}}{U_l\sqrt{2N}}\right|. \qquad (22)$$

In the case of supplying the set across the balance line, a direct feed current $I_{o\ in}$ flows through the transformer which, when short-circuited, provides for the operation of the telephone office instruments, which also determines the magnitude of this current. The current $I_{o\ in}$ creating the ampere-turns of the constant transformer field, determines the dimensions of the core of the latter.

Figure 3 shows the passage of the feed current for circuit 1 of the set (Fig.1), which consists of two components: I_{ob}, connecting across the resistance R_b of the balance line Z_b; and I_{ok}, feeding a semiconductor triode. The latter creates an inverse magnetic field, so that the ampere-turns AW of the constant excitation are equal to

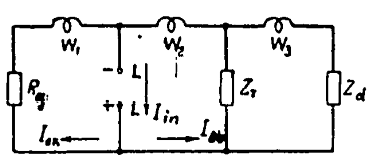

Fig.3

$$AW = I_{0b}(w_2 + w_3) - I_{0\kappa}w_1.$$

For comparison with other circuit variations, we relate this magnitude to the number of turns of the least winding (in circuit 1 of Fig.1, this is w_3).

Using eq.(1), we find

$$aw = \frac{AW}{w_3} = I_{0b}\left(1 + \frac{1}{n_2}\right) - I_{0\kappa}\frac{n_1}{n_2}.$$

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We determine aw at optimum parameters of the set circuit, for which we use eqs.(15), (18), (19), and (21); then,

$$aw = I_{ob}\sqrt{\frac{2}{T}} - I_{0\kappa}\sqrt{\frac{N}{MT}}. \qquad (23)$$

The quantity aw can have two values, determined by the selection of the plus or minus sign in finding M and T.

Selection of a Set Circuit with Feed through the Balance Line

We will discuss the circuits 2 - 6 in Fig.1. Similarly to what was done above for circuit 1, analogous design formulas can be derived for circuits 2 - 6, at optimum parameters of these circuits. From these, the conclusion can be drawn that the magnitude of the minimum total operating attenuation of reception and transmission b_{set}, determined according to eq.(16), and also the magnitudes M, T, p_o, and aw, determined according to eqs.(19), (21), (22), and (23) are identical for all circuits 1 - 6.

The mathematical formulas n_1, n_2, and $\frac{U_{ll}}{U_n}$ for the circuits are given in Table 1.

Table 1

№ of Circuit	1	2	3	4	5	6
n_1	$\pm\sqrt{2N}-1$	$\pm\sqrt{\frac{N}{2}}-1$	$\pm\sqrt{2N}+1$	$\pm\sqrt{2N}-2$	$\frac{1}{\pm\sqrt{2N}}-1$	$\frac{1}{\pm\sqrt{\frac{N}{2}}}-1$
n_2	$\frac{n_1+1}{n_1-1}$	$-\frac{n_1+1}{2n_1+1}$	$\frac{n_1+1}{n_1-1}$	$-\frac{2(n_1+1)}{n_1+2}$	$-\frac{n_1+1}{2n_1+1}$	$\frac{n_1+1}{n_1-1}$
$\frac{U_l}{U_n}$	$\frac{1}{1+n_1}$	$\frac{1}{2(1+n_1)}$	$\frac{1}{1-n_1}$	$\frac{1}{2+n_1}$	$1+n_1$	$\frac{1+n_1}{2}$

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On the basis of the quantity M, the balance-line resistance R_b is calculated, which in case of set being fed across the balance line is found from eq.(3). Consequently, $R_b \approx r_{in}$ also has an identical magnitude for all circuits 1 - 6.

Thus, with an ideal transformer all the circuits are equivalent.

However, under actual conditions, the resistance of the winding W_1 is sufficiently large (300 - 400 ohms), since it must have a large number of turns (order of 2000 - 4000), which is determined by the tendency to approximate $|Z_n|$ to the practically high value R_g. Consequently, in circuits 5 and 6, the resistance of the winding w_1 causes a considerable increase of the set input resistance to direct current. For its reduction, the resistance R_b would have to be reduced.

Moreover, in circuits 5 and 6, in contrast to circuits 1 - 4, the total current $I_{o\ in}$, feeding the set, passes through the winding w_1. The minimum magnitude of this current is equal to 19 - 20 ma so that the voltage drop in it amounts to 6 - 8 v. This reduces the direct voltage feed $U_{ok} \approx I_{o\ in} R_b$ in the collector to an impermissibly low value. Thus, circuits 5 and 6 are not suitable for work with transistors when the feed is through the balance line.

Comparing circuits 1 and 2 with circuits 3 and 4, it will be noted that, in the latter, the collector (i.e., R_g) receives less voltage feed U_{ok} because of the voltage drop in the windings w_2 and w_3, created by the current $I_{o\ in}$ (order of 0.4 to 0.6 v). It should be kept in mind that $I_{o\ in}$ is considerably larger than I_{ok}, which flows in these circuits through the winding w_1.

Circuits 1 and 2 are, therefore, preferable from the viewpoint of obtaining the highest voltage on the collector of the transistor.

A comparison of circuits 1 and 2 shows that the transformer of circuit 1 requires somewhat less copper in the winding. Consequently, circuit 1 should be considered the most rational.

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Frequency Characteristics of the Set during Transmission

The elements of the set circuit were designed on the assumption that it operates on a cable line with a conductor diameter of $d = 0{,}5$ mm, in which, at a frequency of $f = 1000$ cps, the modulus of characteristic impedance would be $Z_l = 775$ ohms, and the angle $\varphi_l = -43^{\circ}40'$. We assume $r_{in} \approx R_b = 600$ ohms, which secures a current strength $I_{o\ in} = 19 - 20$ ma in subscriber lines of more than 6.3 km length, sufficient for automatic telephone office operation. Equation (3) yields the quantity $M = 0.565$.

From the curves of Fig.2 it is obvious that M_- has to be used in order to derive a sufficiently high value of $N = \frac{|Z_n|}{|Z_l|} > 0.5$, as is required for optimum matching of the circuit of the set with a triode. Then, eq.(19), considering the minus sign, it is possible to find $-\sqrt{N} = -4.34$ and $|Z_n| = 14.6$ k-ohm.

From eqs.(18) and (15) we find $n_1 = -7.15$ and $n_2 = 0.75$. In testing a set assembled according to circuit 1 of Fig.1 (Fig.4), a semiconductor triode of the P1D type was used. On the basis of its parameters, measured under operating conditions according to known formulas, its output resistance was calculated, which, with an electromagnetic microphone at its input, exceeded 100 k-ohm. Thus, the resistance of the load at the amplifier output $|Z_n|$ is considerably less than R_g.

Complete matching, i.e., $|Z_n| = R_g$, cannot be achieved, since $|Z_n|$ cannot be made an arbitrary magnitude; consequently, the attenuation of transmission b_{tr} will not be minimal, and the amplification furnished by the amplifier will be less than with matching.

At the same time, a reduction of the load resistance $|Z_n|$ at the triode output, in comparison with its output resistance R_g, increases the maximum permissible undistorted sidetone of the set because of the reduction in the voltage ratio n_1.

We note that for the carbon microphone of the conventional telephone set, according to the latest technical requirements, a factor of nonlinear distortions of

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as high as 15% at 20 b is permitted. It is known that substantially less nonlinear distortions can be achieved in semiconductor amplifiers.

The negative value n_1 indicates an opposition to cutting-in of the winding of the transformer w_1 with respect to the winding of w_2. In the circuit of Fig.4, the beginning and end of the transformer windings are marked by the letters n and k, and their cutting-in corresponds to the derived signs n_1 and n_2.

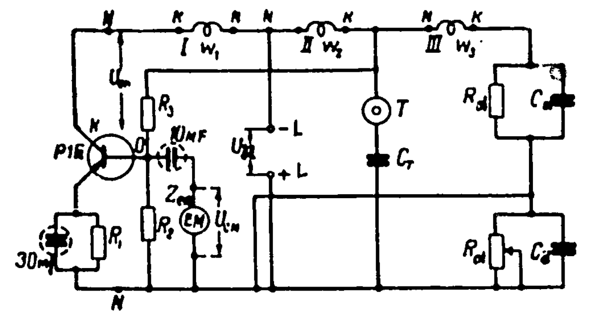

Fig.4

On cut-in, the sets establish an R_d of such magnitude that $I_{o\ in}$ = 19 - 20 ma will result, irrespective of the length of the subscriber line. This maintains the voltage drop in R_b constant and thus keeps the voltage in the collector U_{ok} to the order of 10 v. Thus, the voltage at the operating point will be equal to -12 v, irrespective of the length of the subscriber line.

To obtain the required condition of feed of the triode and stabilization of am-

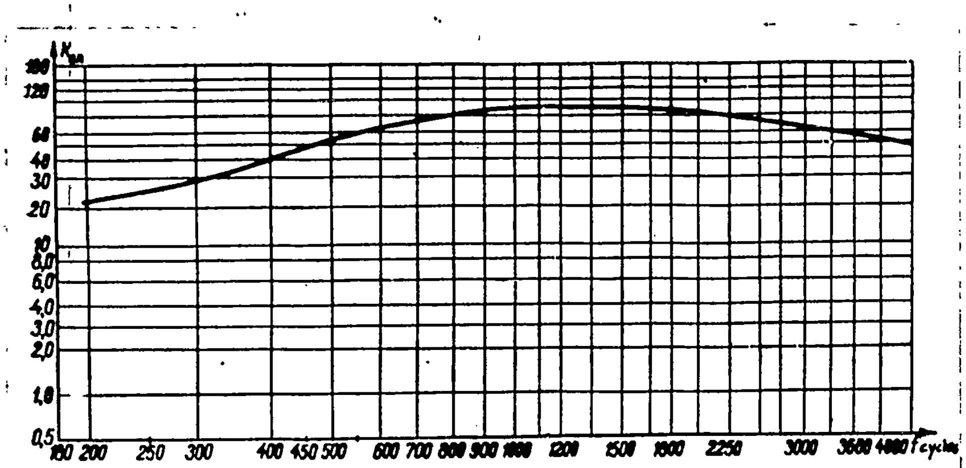

Fig.5

plifier operation on replacement of the triode or a rise in the ambient temperature, the resistances R_1, R_2, and R_3 are used. In order that the reduction in amplifica-

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tion will not be excessive at low frequencies, the capacitance of the emitter circuit is taken sufficiently large (30 μf). A capacitance of 10 μf must provide a sufficiently low resistance to alternating current in the microphone circuit.

The calculation of the operating conditions of the amplifier and of the magnitudes R_1, R_2, and R_3 can be based on methods given in the special literature (Bibl.2).

Figure 5 shows the frequency dependence of the transmission factor of the circuit according to the voltage K_{an}, where (Fig.4)

$$K_{an} = \frac{U_{tl}}{U_{em}},$$

and the resistances are R_1 = 1 k-ohm; R_2 = 5 k-ohm; and R_3 = 90 k-ohm.

The frequency characteristics* are shown in Fig.6:

1. Sidetones of the set with differential electromagnetic microphone from a no-battery set TAK ("Ship Telephone Set"). The average sidetone of the set in the line in the range of 300 - 3500 cps is equal to 51.5 mv/b. The irregularity is 50 db**.

2. Sidetones of the set with differential electromagnetic microphone from an American no-battery set EE-108, with somewhat improved sidetone characteristic***.

The average sidetone of the set in the same range is equal to 30 mv/b. The irregularity is 33.4 db**.

3. Sidetones of the carbon microphone AWZ of "Albiswerk" (Switzerland). The average sidetone of the microphone is equal to 22.4 mv/b. The irregularity is is 9.85 db.

4. Sidetones of the carbon microphone of the Danish set M-52. The average sidetone of the microphone is equal to 35.8 mv/b. The irregularity is 14.8 db.

*In the measurement of the characteristics, A.Ya.Markovich participated.

**The measurement was made with mouthpiece present.

***An improvement of the characteristics was achieved by M.F.Kopp, Ye.A.Samoylenko, and V.M.Romantsov by changing microphone design and arranging suitable air cells.

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When comparing the sidetone curves of Fig.6, it should be kept in mind that the curves 1 and 2 are cited for the set as a whole, while 3 and 4 are for the microphone.

The width of the spectrum reproduced by the set with amplifier and the frequency characteristic of its sidetone are determined by the characteristic of the elec-

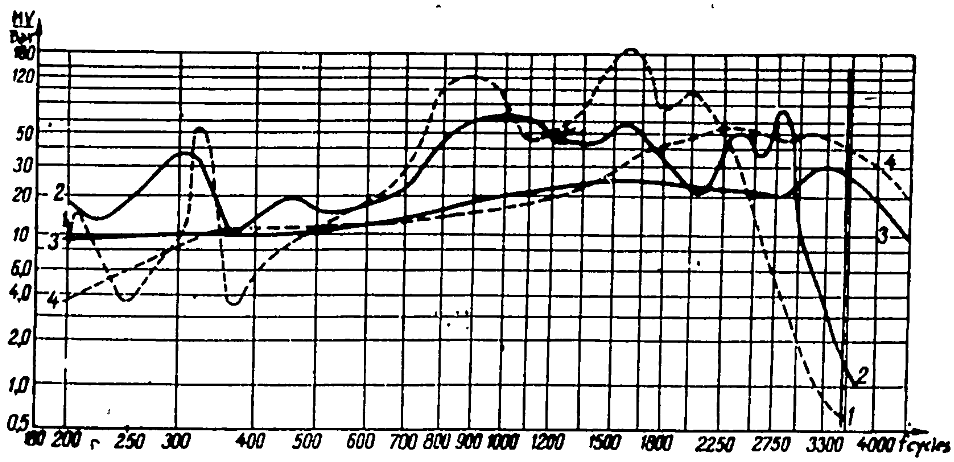

Fig.6

tromagnetic microphone used, which has a resonance character. Such a characteristic is explained by the tendency to get as great a sidetone as possible in the no-battery TAK and EE-108 sets.

A much higher amplification can be obtained by using a dual amplifier. The latter permits lowering the demands made on the sidetone of the electromagnetic microphone, expanding the resultant frequency spectra, and giving the frequency characteristic its necessary slope. This naturally requires the development of a suitable electromagnetic microphone, which is fully feasible. This makes it possible to obtain a telephone set with a preseribed, most rational electroacoustic characteristic (Bibl.3).

The use of small-dimension electrolytic capacitors as well as the small size of semiconductor triodes and resistances permit placing the entire amplifier inside a microtelephone tube of normal dimension.

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CALCULATION OF ACTIVE RESISTANCES OF CONDUCTORS OF TUBULAR FORM

by

G.P.Delektorskiy

The problems of calculating the active resistance of tubular conductors are examined; an analysis is given of the optimum correlations of tube thicknesses.

At certain frequencies, conductors of tubular form (hollow), are found more economical than solid conductors (or those twisted from a number of bare copper wires). Besides, the economy is a result not only of the lower consumption of copper but also of the reduction of active resistance.

We will examine formulas for calculating the active resistance of conductors of tubular form and show the frequency limits of their application.

For calculating the resistance to direct current, the following formula is applicable:

$$R_0 = \frac{1}{\pi \sigma t\,(2r - t)} \text{ Ohm/cm}, \tag{1}$$

where R_0 is the resistance to direct current, in ohm/cm;

r is the outside radius of the tubular conductor, in cm;

t is the tube thickness, in cm;

σ is the specific electric conductivity, in $\frac{1}{\text{ohm - cm}}$.

However, this formula can be successfully used for calculating active resistances of conductors at low frequencies (Bibl.1). The ratio of tube thickness to the equivalent depth of penetration θ or the product of the factor of eddy currents k times the tube thickness t determines the frequency limit of the application of eq.(1). The indicated ratio must not be greater than 1.

Actually, in these cases the equivalent depth of penetration is equal to or greater than the tube thickness and the current can be assumed to be distributed uniformly over the cross section. Consequently, the applicability of eq.(1) will be determined by the inequation

$$0 < \frac{t}{\theta} < 1 \quad \text{or} \quad 0 < kt < \sqrt{2}$$

(for copper, $k = 0.21\sqrt{f}$ and $\theta = \frac{\sqrt{2}}{k}$).

Since, according to the meaning of conductors of tubular form, the inequation $t < r$ is also valid, the following inequation can be recommended as frequency limits of the applicability of eq.(1):

$$0 < kr < \sqrt{2} \quad \text{(if } t \text{ is comparable with } r\text{)}.$$

For the limiting conditions $\sqrt{2} < kr \mu 5$ and $\sqrt{2} < kt < 2\sqrt{2}(1 < \frac{t}{\theta} < 2)$ when calculating active resistances, the full formula should be applied, comprising the Bessel function (Bibl.2). This equation is not derived here since it is too complex for practical computations and, as will be shown below, is practically never used.

For the cases when $\frac{t}{\theta}$ is within the limits of $1 < \frac{t}{\theta} < 2.5$ or $2\sqrt{2} < kt < 3.54$ (boosted frequencies), and also when the product of the factor of eddy currents k and the radius of the conductor r are equal to or somewhat greater than 5, the following formula is valid (Bibl.2) for calculating the active resistance of a tubular conductor

$$R_a = \frac{1}{2\pi r\sigma}\left[\frac{k}{\sqrt{2}}\,\frac{\operatorname{sh} u + \sin u}{\operatorname{ch} u - \cos u} + \frac{4r - t}{8r(r - t)}\right]\frac{\text{ohm}}{\text{cm}}, \qquad (2)$$

where R_a is the active resistance of the conductor, in ohm/cm;

σ is the specific conductance of the material of the conductor, in $\frac{1}{\text{ohm - cm}}$;

r is the outside of the tube, in cm;

t is the tube thickness, in cm;

$u = \sqrt{2kt}$;

$k = 0.21\sqrt{f}\,\frac{1}{\text{cm}}$ (for copper).

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For large values of k, eq.(2) can be simplified. In fact, if an accuracy of 5% is required, the ratio of the first term in the brackets of eq.(2) to the second term, equal to approximately 0.5, must be equal to 95:5. This means that $kr \geq 13.5$ to 14.5 (as will be shown later, the magnitude $\frac{\sinh u + \sin u}{\cosh u - \cos u}$ is within the limits of 0.92 - 1).

Consequently, eq.(2) assumes the form

$$R_a = \frac{k}{2\sqrt{2}\pi r\sigma} \frac{\sinh u + \sin u}{\cosh u - \cos u}. \qquad (3)$$

The second condition for the applicability of eq.(3) is the inequation

$$t \ll r.$$

If this condition is not satisfied, eq.(4) is preferable for the calculations (Bibl.1)

$$\frac{R_a}{R_0} = \frac{kt}{\sqrt{2}} \frac{\sinh u + \sin u}{\cosh u - \cos u}, \qquad (4)$$

where R_0 is the resistance to direct current, determined by eq.(1).

Equation (2) finds application in cases when it is required to calculate the active resistance of tubular conductors of relatively large diameters (0.5 - 2 cm) and operate at relatively low frequencies (2.5 - 10 kc).

When estimating the resistances according to eqs.(2) - (4), it is essential to be able to select the optimum thickness of the tube t. An analysis of eq.(2) permits the conclusion that the minimum resistance (for a conductor of a given radius) will occur at

$$u = \pi.$$

Then, the tube thickness is determined from the correlation

$$t = \frac{\pi}{\sqrt{2}\,k} \text{ cm}, \qquad (5)$$

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and eq.(2) acquires the form

$$R_a = \frac{1}{2\pi r\sigma}\left[\frac{0{,}92k}{\sqrt{2}} + \frac{4r - t}{8r\,(r - t)}\right]\frac{ohm}{cm}. \qquad (6)$$

From the cited calculations, it is obviously possible to draw the conclusion that a reduction in the tube thickness relative to the optimum thickness, will lead to an increase in the active resistance of the conductor. Moreover, the increase in tube thickness over the optimum thickness also causes a certain increase in resistance, which is at first glance somewhat unusual. The described phenomenon has the following physical explanation: For the internal conductor of a concentric cable, the increase in resistance relative to the ohmic resistance in alternating current (especially of high frequencies) is explained by the action of only the skin effect. The latter can be considered a result of attenuation, in a conductor, of an electromagnetic wave which penetrates from a dielectric into the depth of the conductor. Considering that the length of the wave in the conducting material is considerably shorter than the radius of the conductor, it is possible to assume that the wave penetrating into the conductor depth is plane.

The variation of a plane wave in a conducting medium follows (Bibl.3) a wave equation of the form

$$I = \sqrt{i\omega\mu\sigma}\,H_m e^{-\frac{kz}{\sqrt{2}}} e^{\frac{ikz}{\sqrt{2}}}, \qquad (7)$$

where I is the value of the vector of current density;

H_m is the maximum value of the vector of the magnetic field;

σ is the specific conductance $\omega = 2\pi f$, where f is the frequency in cycles;

μ is the magnetic permeability;

k is the factor of eddy currents;

z is a coordinate perpendicular to the axis of the conductor.

A study of eq.(7) indicates that the current amplitude, greatest at the conduc-

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tor surface ($z = 0$), rapidly declines in proportion to the penetration into the depth of the conductor. At the same time, the attenuation proceeds the faster at a higher frequency. Moreover, the curve for the current in the depth of the conductor has a sinusoidal character. This signifies that the current, being directed to one side at the surface, changes to an opposite direction at the instant the quantity $\frac{kz}{\sqrt{2}}$ becomes equal to $\pi/2$.

The change in current direction to the opposite results in an increase of loss (a current, opposite to the operating current, is generated). Further changes in the current direction no longer are of significance, since its amplitude in the second period is negligibly small. Therefore, the quantity z, determined from the correlations

$$\frac{kz}{\sqrt{2}} = \frac{\pi}{2} \text{ and } z = \frac{\pi}{\sqrt{2}k},$$

is the optimum magnitude of the tube. The same expression for this magnitude was derived earlier.

Equations (2), (3), and (6) for calculating the resistances of tubular conductors can be applied also to higher frequencies. Because of the necessity of deriving the maximum value t, however, the quantity u must always be equal to 3.14.

At frequencies of 10 megacycles and higher, the optimum thickness of the tube becomes very small (less than 0.1 mm), and the manufacture of conductors of such thicknesses is difficult. Practically, therefore, the thicknesses of tubular conductors in these cases are greater than required by eq.(5). At the same time, the difference in the resistances of conductors of both designs vanishes (the value kt becomes greater than 3.54). The formula for calculating the resistance is simplified.

The formula for computing active resistances of tubular conductors at high frequencies is applicable when the following conditions are satisfied:

a) $\theta \ll t$, $kt > 3.54$ ($u > 5$),

b) $t \ll r$,

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c) $kr \geq 14.5$

In this case, eqs.(3) and (6) acquire the form

$$R_a = \frac{k}{2\sqrt{2\pi r c}} \text{ ohm/cm,} \quad (8)$$

$$R_a = \frac{4{,}18 \cdot 10^{-3}\sqrt{f}}{r} \text{ ohm/cm (for copper)} \quad (9)$$

(here r is given in cm), i.e., become analogous to the formula for calculating the resistance of solid conductors.

We will discuss the case when the above equation should be applied. The use of internal conductors for concentric cables in the form of tubes is possible in the following types: 1) powerful cables for induction heating of large cross sections; 2) telephone cables; 3) radio cables.

Table 1 lists the sizes of the internal conductors for the indicated cables, the ranges of operating frequencies, and formulas for calculating.

Table 1

Group of Cables	Range of Operating Frequencies	Radii of Internal Conductors (in cm)	Magnitudes of Parameters kr_{min}	Formulas for Calculating Resistance
1	2.5 to 15 kc	0.5 - 1.5	5.2	(6)
2	60 to 300 kc	0.1 - 0.5	5.1	(6)
3	1 to 100 mc	0.15 - 0.6	31.5	(6) and (9)

As is evident, eqs.(6) and (9) are basic for calculating the active resistances of tubular conductors.

We determine the ratio of the resistance of a tubular conductor to alternating current R_a to the resistance of the conductor to direct current R_o for the case when the tube thickness t is optimal, and the resistance is calculated according to eq.(3) with consideration of eq.(5). This ratio becomes equal to

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$$\frac{R_a}{R_0}=\frac{0,92kt}{\sqrt{2}}=\frac{0,92k}{\sqrt{2}}\frac{\pi}{\sqrt{2k}}=1,44.$$

Thus, the ratio of resistances is a constant magnitude.

If the magnitude u becomes greater than 3.14 and increases further, then the ratio of resistances rises with an increase in frequency, since the thickness of the walls in these cases does not change (as was indicated above, it is selected as small as possible, based on the technological possibilities of manufacture). In this case, the ratio of resistances has the form

$$\frac{R_a}{R_0}=\frac{kt\,(2r-t)}{2\sqrt{2r}},$$

or, considering that $t \ll r$,

$$\frac{R_a}{R_0}=\frac{kt}{\sqrt{2}}=0,15t\sqrt{f}.$$

It is also of considerable interest to determine the lower (frequency) limit of the usefulness of conductors of tubular form. For this, it is necessary that the active resistance of a tubular conductor, determined by eq.(6), be adjusted to the resistance of a solid conductor, which is expressed (Bibl.3) by the following formula

$$R_a=R_0\left(\frac{kr}{2\sqrt{2}}+0,25+\frac{3\sqrt{2}}{32kr}\right)$$

(for the case, when $\sqrt{2} < kr < 50\sqrt{2}$).

The solution of the equation

$$\frac{1}{2\pi r\sigma}\left[\frac{0,92k}{\sqrt{2}}+\frac{4r-t}{8r\,(r-t)}\right]=\frac{1}{\pi r^2\sigma}\left(\frac{kr}{2\sqrt{2}}+0,25+\frac{3\sqrt{2}}{32kr}\right)$$

relative to kr, gives kr = 5.

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All of the foregoing permits the following conclusions:

1. Conductors of tubular form (hollow) with direct current and very low frequencies have greater resistance than solid conductors of the same size. With an increase in frequencies, the resistance of solid conductors rises more rapidly than the resistance of tubular conductors, and at a definite instant ($kr = 5$) the resistances are aligned. On a further rise in frequencies, the resistance of tubular conductors becomes less than that of solid conductors because of the reduction in loss from eddy currents. The cited considerations are valid for tubular conductors with optimum values of tube thickness. Finally, at still higher frequencies, when for technological reasons it is impossible to maintain the tube thickness at an optimum value, the resistances of both types of conductors are again made identical. Thus, the application of tubular conductors is useful when $kr > 5$.

2. At optimum tube thickness in the entire frequency range, in which eq.(6) is applicable, the ratio of resistances (active to ohmic) amounts to 1.44.

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FROM FOREIGN JOURNALS

Brief Reports

NEW ELECTRONIC TUBES AND SEMICONDUCTOR DEVICES

At the exhibit of the British Physical Society, which was arranged in May 1956 and devoted to electronics of the Twentieth Century, a number of firms exhibited new electric-vacuum and semiconductor devices.

Special and Transmitting Radio Tubes

The new ceramic radio tubes attracted attention among the various new electric-vacuum radio tubes presented at the exhibit. The use of ceramics instead of glass in the manufacture of tube bulbs is not new; at present, however, the use of ceramics has reached considerable development. The use of ceramics permits designing tubes with greater mechanical strength and smaller dimensions for a given dissipated power. Such tubes can operate well under higher ambient temperatures and allow for more effective degassing, which makes it possible to get greater emission current in impulse conditions. The series of ceramic radio tubes shown by the "Ferranti" Co. included: triodes for generators and amplifiers with dissipated plate power from 15 to 100 watts; triodes with low μ for use in stabilized power units; and half-wave rectifiers of indirect heating. The maximum operating temperature for these tubes reaches 250°C.

Larger ceramic radio tubes were displayed by the "English Electric" Co. Demonstrated in particular was a coaxial transmitting triode CR 1101 with air-cooled anode, able to dissipate 2 kw of power. This tube can operate at a frequency of 900 mc, delivering at the same time useful power of 600 watts.

Shown by this same firm was a new magnetron M 541, designed for operation under pulse conditions at a frequency of about 1200 mc with a peak output of 0.5 megawatt.

The most familiar generator tube with velocity modulation is the klystron; this tube, however, is little known as a device permitting wide retuning of frequencies. At the exhibit, the "E.M.I." Co. demonstrated a new reflection klystron R 5222 with an external cavity resonator, permitting tuning in the range 8400 - 10,100 mc. Besides, the output power increases from 28 to 55 mw in the average frequency range; with further increase in frequency the power again drops to 25 mw.

Receiving Radio Tubes

During multichannel television reception, signals greatly differing in amplitude frequently enter the input of the receiver. This may lead to distortions due to cross modulation between picture signals and audio signals. The "G.E.C." Co. at present produces new pentodes with variable-mu of the W729 type. These pentodes possess considerably better characteristics than the usual tubes of the "vari-mu" type.

For the reception of AM and FM signals, this same firm has developed a new tripartite (triple) diode-triode DH719/EABC80, in which one diode has a separate cathode. The "Mullard" Co. showed a radio-frequency variable-mu pentode with two diodes (EBF 89), which is distinguished by low parasitic grid-plate capacitance (less than 0.002 $\mu\mu f$) and moderate mutual conductance (3.6 ma/v). This pentode gives a high amplification factor without the danger of self-excitation in the IF cascades of receivers with amplitude or frequency modulation.

Electronic Tubes

At the exhibit two new 21-inch (53.3 cm) television tubes with 90° angle of deflection were shown. One of them is of the C21KM type with a tetrode gun developed by the "Brimar" Co.; the other is of the 7501A type with a triode gun developed by the "G.E.C" Co. The screens of both tubes are aluminized. The "Ediswan" Co. showed a 9-inch (22.8ccm) television tube of the CRM93 type. This same firm produces also a 24-inch (60.96 cm) tube with rectanular screen.

Among the oscillogrons shown at the exhibit were tubes of a new type which, instead of the usual series of rings, had a spiral electrode with acceleration after deflection. Such an electrode is obtained by coating glass with a material possessing a certain resistance, and feeding it with an accelerating potential of 10 kv. The recording speed of the tube is 1000 cm/ μ sec. Such a tube of the Cintel G 601-C4 type with acceleration after deflection is capable of making a recording at a frequency as high as 1500 mc.

Transistor Triodes

The "Mullard" Co. demonstrated several examples of the application of transistor triodes operating under "avalanche" conditions, i.e., when operating similarly to a thyratron. The collector, in addition, is under high voltage of a magnitude sufficient to have the flow of holes in the direction of the collector proceed with the velocity required for ionization of the atoms near the collector. The free electrons created in this process flow in the base circuit and thereby improve the triode's amplification action. Such a device can be used for very rapid discharge of a capacitor. At a feed voltage of 70 v, the triode in "avalanche" conditions is capable of delivering a pulse current with a strength of 100 ma at a duration of 0.05 μ sec. The voltage in this case has an amplitude of 60 v and a sawtooth shape. The possible pulse repetition rate amounts to ~ 1 mc.

Exhibited by the "B.T.H." Co. was a new powerful germanium semiconductor triode of the junction type, which can be used in switches designed for 5 amp. A new junction semiconductor triode of the EW 53 type was displayed by the "G.E.C." Co. In this triode, the base is attached to metal. In consequence of this, the semiconductor triode has a lower thermal resistance and a higher mechanical strength.

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The "English Electric" Co. was represented at the exhibit by several types of junction-type semiconductor triodes, designed for use in low-power devices operating at audio frequency.

The "Brimar" Co. showed three new types of semiconductor triodes (TS1, TS2, TS3) of the same assigned purpose. They are well "potted" and replace the previously made TS model semiconductor triodes.

Rectifiers

Widely represented at this same exhibit were transistor diodes which function as rectifiers. Silicon junction-type diodes have gained widest development in this field. The silicon diode is distinguished by small size for a given rated current, a capacity to operate at high temperatures, and small reverse current. The usual small diode, having the size of a 0.25-watt resistance, yields a rectified current of 200 ma and has a reverse current of about 5 μ amp with peak reverse voltage at 100 v at a temperature of 100°C. Samples of such diodes were shown by the "B.T.H.", "G.E.C." and "S.T.C." firms. Also represented at the exhibit were powerful rectifier diodes, capable of passing a current of 30 - 50 amp at a voltage of 100 - 200 v. The diameter of such triodes is only 1.9 cm overall.

New germanium junction rectifiers have a considerably lower resistance in straight direction. For example, the "G.E.C." Co. showed a germanium junction-type diode with air cooling, capable of giving a rectified current of 300 amp.

The "Mullard" Co. displayed a new germanium junction-type diode of the OA10 type which is distinguished by extremely low hole-type conductivity.

A new method of depositing selenium on metal plates by means of vacuum evaporation has led to the creation of metallic rectifiers with somewhat improved characteristic. This process secures a solid bond and makes it possible to deposit a thinner layer of selenium on the plate, which reduces the resistance in the straight direction and increases the magnitude of current, rated for the given size of recti-

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fier, by about 25%. The service life of the rectifier is also increased.

At the exhibit several types of selenium rectifiers which can operate at t = = 85°C were shown. One such rectifier, suitable for operation in TV receivers, can give a rectified current of 300 amp at voltage of 270 v. It contains units with contact cooling at the leads, enclosed in a block of plastic of approximately 2.5 cm^2 in size. It will operate successfully even if the mounting frame, to which it is fixed, is heated to 60°C.

Photoelectric Devices

The photomultiplier of the "Venetian-blind" type possesses a high degree of focusing from cascade to cascade and a large intercascade efficiency. Also distinguished by improved efficiency is the photomultiplier of the "B.T.H." Co. with a cadmium sulfite photoelectric cell. Such a photocell has greater sensitivity than selenium photocells.

The "Mullard" Co. exhibited a new germanium phototransistor of the OCP71 type. It has very small dimensions, operates at a voltage of 10 v and in illumination gives an output current of 5 ma. The current of a "dark" phototransistor amounts to not more than 300 μ amp.

(Wireless World, June 1956, p 281)

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NEW AMERICAN MASS-PRODUCED TV SET

A brief description is given below of a new TV set developed by the General Electric Co. (USA) for mass output in 1956.

The table model TV 14T008 (Fig.1) is a thirteen-tube set with a 14-inch rectangular electron-ray tube. The set is designed for reception of 12 programs within the ranges of 54 - 88 mc, 174 - 216 mc. The sensitivity of the set is 75 - 100 μ v. Picture definition during reception of broadcasts from the Moscow television center is 350 lines. The weight of the set is 14 kg; its dimensions 400 × 342 × × 290 mm; intake is 100 watts.

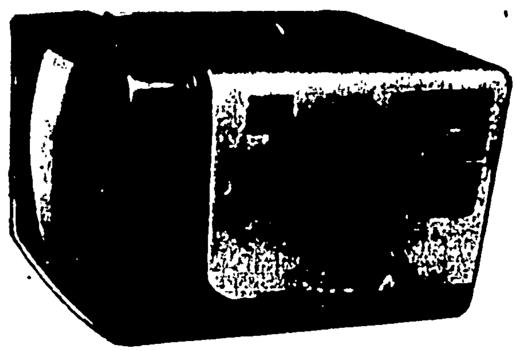

Fig.1

Manufacture of the set is designed for printed assembly and dip soldering. Part of the conductors is reliably connected by twisted joint without soldering. Junction points and parts are fastened lightly and at the same time reliably, by numerous spring nuts stamped from tin, self-threading screws and others. The cable lugs have stamped-out contact points and are fastened in an unshielded conductor by the method of point punchings. Thanks to all this and also to the high level of mechanization of the parts manufacture and of assembly processes, the cost of the set has been substantially reduced.

Widely used in the set circuit are combination tubes (triode-pentodes, dual triodes, diode-triodes), and also ferrocart with high penetrance, which permits compact assembly and lower power consumption.

The set chassis is arranged vertically, which substantially reduces the dimensions and weight of the set (Fig.2). Such a set suffers from a significant defect - complexity of set repair. Thus, for replacement of any resistance or capacitor, it

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is necessary to pull out the chassis and remove the pertinax panels with printed assembly, which makes even minor repair labor-consuming.

The subcarrier and LF amplifiers of the video channel and audio channel and also the generator of lines and frames are assembled in two panels with printed circuits (pertinax foil) and connecting conductors. All volume parts are hinged. The subcarrier frequency amplifier, heterodyne, and mixer are assembled in a separate shielded unit with a pancake-type band switch.

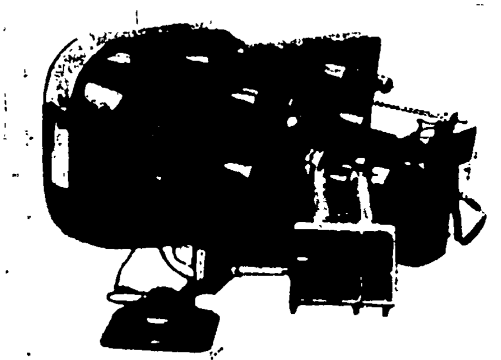

Fig.2

The electric connection of parts and printed circuit is accomplished by dip soldering, twisted joints, and ordinary soldering. The tube panels, manufactured independently, are soldered into the printed circuit by the dip-soldering method. Interunit connections are made by twisted joint.

Widely applied are combination electrolytic capacitors (200, 60, and 30 μf in one frame). Simplified open and small-dimension variable resistances are used. Decoupling chokes are pressed in plastic and assembled as ready-made parts. The contacts of ceramic capacitors are led out to one side, which permits using them at higher frequencies.

The deflecting system and the line transformer are assembled in ferrocart of high penetrance. The deflecting system is attached directly to the tube neck. Nylon insulation is used widely.

The loops of intermediate frequency and the loops of the detector are wound on cardboard frames. The receiver has series feed of the tube channel AC mains current of 115 v. The plate feed of the circuit is accomplished from a half-wave selenium rectifier of 130 v, without reducing transformer. The vertical sweep is fed by a voltage of 260 v, generated in the circuit by the line scanning. The dynamic loudspeaker is of the core type, without outside magnetic field.

In conclusion, it should be noted that, at extremely reliable and stable operation of the set, the picture definition and quality of its sound leave much to be desired and cannot serve as a basis for development of a Russian mass-produced TV set.

RADIO-RELAY SYSTEM WITH PULSE-PHASE MODULATION

In Austria, a microwave radio-relay system has been developed by the "Telefunken" Co. The system has a length of 850 km, uses pulse-phase modulation, and is designed for transmission of ultrashort-wave radiobroadcast programs and, subsequently, for transmission of television. The system provides three radio channels with a pass-band width of about 15 kc in each direction, and six telephone channels. The system is divided into 17 sections. The length of each section is 50 km. Parabolic antennas of 3 m diameter are employed, two antennas in each section. In sections of greater length, depending on local conditions, parabolic antennas of 4 m diameter with exciter and frame are used.

The line runs from Dornbirn to Klagenfurt through Innsbruck, Salzburg, Linz, Vienna, and Graz. The most complex section, from a technical viewpoint, runs between the Zugspitze (station at an elevation of 2805 m) and Pfaender (1064 m), where mountain ranges have to be surmounted and where no points of direct visibility between stations are available in the district. Here direct communication is secured beyond the limits of the horizon by application of the technology of passive radio relaying. For this purpose two radiation-coupled reflectors of rhombic form have been installed on Mt. Walluga (2811 m).

By one reflector, the beam coming from the Zugspitze is deflected to a station located lower at Ulmer-Huette where installation of active radio-relay stations is possible. From Ulmer-Huette, the beam is reflected by the second reflector on

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Mt. Walluga, and then is refracted in the direction of the Pfaender.

The relay station is built at an elevation of 2285 m at the level of Ulmer-Huette. To avoid intense fadings in the repeater sections, the relay stations Zugspitze and Ulmer-Huette were equipped with parabolic reflectors with an aperture area of 6.4×4 m^2. These reflectors are truncated parabolas, so that an almost rectangular surface is obtained. The rhombic reflectors installed on Mt. Walluga have 12 axes, 8 and 5 meters, which creates an area of radiation-coupled reflection of 32 m^2. The focal length of the parabolic antennas was so selected that the aperture angle amounts to approximately 160°. The body of the parabola is made of light metal. The frame permits rotating the parabolic part of the antenna in the vertical through ±5° and in the horizontal, through ±15°. Due to this shifting, sharply directional reception is secured. The dipole is held and fed by means of a concentric supply, which in its turn, is held by a cone mounted to the top of the parabola on a cast plate. The dipole, together with its supply lead and conic holder, can be easily unscrewed and replaced by another. For convenience in transport, the antenna is dismantled in three sections. All the antennas with their holders can withstand a wind velocity of 220 km/hr.

Because of these antennas, the attenuation in the radio sections Pfaender - Ulmer-Huette and Ulmer-Huette - Zugspitze amounts to 75.8 and 77.1 db as compared with the attenuation of 66 db in the sections of 50 km in length with antenna-reflectors 3 m in diameter. These values are 10.5 to 14 times higher than the values of corresponding standard radio sections.

(Radio Mentor, No.10, 1956; Telefunken-Zeitung, No.3, 1956)

MEASURING INSTRUMENTS IN LONG-DISTANCE COMMUNICATIONS OF THE FEDERAL REPUBLIC OF GERMANY

An instrument has been developed in the German Federal Republic for measuring

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level diagrams of 60 and 120 channel long-distance systems of communications without disrupting their operation. As is known, for a single telephone channel, a frequency band of 4000-cycle width is shunted in each direction, whereas only the spectrum with a width of 3100 cps (which corresponds to a speech spectrum of 0.3 to 3.4 kc) is used for transmission. This leaves intervals of 900-cycle width between the spectra of individual channels.

By means of the described instrument, currents of reference frequencies are introduced into the line, their values being arranged in the mentioned intervals between channels. The 22-spot test frequencies differ from the virtual carrier frequencies by -80 cycles in the range of 12 - 252 kc, and by +80 cycles in the range of 312 - 552 kc. The minimum frequency is equal to 12,000 - 80 cps. The transmission level of the current of the test frequency amounts to - 1.5 neper. The level indicator, forming part of the instrument assembly, has two conversion steps and is characterized by high selectivity.

The selection of sent and received frequencies is done by a key pulser. A possibility is provided for automatic sending and receiving test currents of various frequencies. A galvanometer placed in the transmitting part of the instrument, indicates the magnitude of the input resistance of the line at a given frequency.

An oscillographic level meter, described in the German literature, is of interest. The screen of the cathode tube of this instrument shows the frequency characteristics of the line. Divisions of the scale of the instrument are etched on the mask of the screen, within limits of -2 to +3 nepers. The limits of measurement can vary by nine steps, from -6 to +2.3 nepers. The accuracy of frequency measurement amounts to 5% of ±25 cps; and the accuracy of level measurement at the zero mark is not inferior to ±0.02 neper, at a frequency of 800 cps. The instrument can operate with any oscillator, without special units for synchronization. The device can also be used for measuring the magnitudes of input resistances and of balance attenuations. Instruments of this type are manufactured for the frequency range of 30 -

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to 20,000 cycles and 0.3 - 1200 kc. The input resistance of the instrument can be fixed at 600 ohms, 950 ohms, and 25 k-ohm. The input of the instrument is symmetrical.

(Electro-Technik, Vol.38, No.21, 1956)

SUPPRESSION OF RADIO INTERFERENCE

The Fifth Plenary Session of the International Special Committee on Radio Interference (C.I.S.P.R.) was held in Brussels in July 1956. About 80 delegates took part in the work of the session, representing 17 countries (Austria, Belgium, Great Britain, Holland, Denmark, Italy, Canada, Norway, USSR, USA, Finland,France, Federal Republic of Germany, Czechoslovakia, Sweden, Switzerland, Japan) and five international organizations. The transactions of the session were done by three subcommittees which studied the problems:

a) Standards of interference and methods of control;

b) Measurement of radio interference;

c) Problems of safety measures in suppression of radio interference.

At present, the drafting of specifications is nearing completion for equipment which measures radio interference in the frequencies from 150 kc to 30 mc and work is being conducted on creating equipment for frequencies in the range of 25 - 300 mc.

The following limit standards for interference voltages in the terminals of instruments are recommended:

Frequency Band to be Shielded

150 - 200 kc

200 - 285 kc

525 - 1605 kc

Standards for Interference Voltages at Terminals

500 - 1500 μv

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500 - 1000 μv

500 - 1000 μv

The measurements must be made according to the method of the C.I.S.P.R. For mass-produced electric devices, the standards must be fixed on a statistical basis. These are not absolute maximum values which cannot be exceeded. The C.I.S.P.R. solicited proposals from the International Electrotechnical Commission with respect to measuring radiations from radio and television receivers and will then attempt to formulate standards for interference from frequency converters and generators of line scanning.

Inasmuch as the method of interpretation and application of interference standards, caused by mass-produced electric devices, is of great importance, it was recommended that all session participants study the British method (described in Standard BS 800) and submit their observations.

Many administrations regularly publish statistics of complaints on radio interference, but these data have no common base for comparison. In consequence of this, the C.I.S.P.R. recommends that statistics classify interference as a function of the source compiled separately by various ranges:

a) electric devices of low power, including motors, thermostats, contacts, etc. used in households, stores, offices, and small shops;

b) radiobroadcast and television receivers;

c) gas-discharge tubes;

d) industrial, scientific, or medical high-frequency equipment;

e) high-voltage lines.

In many countries, it is prohibited to connect capacitors to external metal structures of electric devices with double insulation, which hampers getting satisfactory suppression of radio interference. It is, therefore, proposed to carry out further investigations on problems of safety, which might arise when capacitors are connected in this manner.

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The C.I.S.P.R. recommends that attention be directed to the importance of leaks of high-frequency currents which can pass through interference-killing capacitors, and that in developing methods of measurement of current leaks, 2000-ohm a resistance should be assumed as necessary for connection in the ground wire.

The ground leakage currents relative to electric devices of various types for appropriate suppression of interference, are determined in the following manner:

Type	Current Leakage
Fixed, constant grounding	5 ma
Double insulation and full insulation	3.5 ma
Grounded portable device	0.5 ma

(Journ.Brit. of IRE, Vol.16, No.10, 1956, p.554)

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AUTHOR CERTIFICATES

Class 21a[1], 7_{05}, No.103862. P.A.Kotov, L.P.Purtov, and N.P.Sokolov. Device for Recording Distortions of Telegraph Pulses.

In the device for recording distortions of telegraph pulses, fixed on paper tape, with the object of simplifying the recording device and securing high accuracy of recording the magnitude of pulse distortions, it is proposed to employ, as the stylus, a rotating drum with a spiral pen fixed on its surface.

With the purpose of increasing the resolving power of the device and excluding the effect of characteristic distortions, it is proposed, for recording of telegraph pulses on tape, to use a polarized electromagnet connected in the circuit of the receiving relay reed, which will operate only during the time of the transition of the reed of this relay from one contact to another.

Class 21a[1], 13, No.104038. P.A.Kotov. Device for Correcting the Phase of Oscillations of Contact Tuning-Fork Generator.

Class 21a[1], 33_{40}, No.104084. A.S.Fomin. Method for Correcting Distortions of a Transmitted Television Picture.

Class 21a[1], 34_{11}, No.10408 - D.A.Tarapets. Method for Realizing Combined Television Transmissions.

Class 21a[3], 63_{30}, No.103746. S.D.Shul'ga and N.M.Radeyev. Device for Frequency Multiplication.

In the device for frequency multiplication, realized in the circuit of a magnetic harmonic generator, it is proposed, with the object of increasing the power of the useful signal and suppression of adjacent frequencies, to provide the harmonic generator with two series-connected oscillatory circuits, tuned to the frequency of the useful signal, with the tank inductances represented by transformers with magnetic coupling through a common plate of a ferromagnetic material.

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Class 21a^4, 14_{01}. No.103743. V.P.Mokshantsev. Modulator.

With the object of getting a modulating signal proportional to the sum of the input signal and its first derivative, the modulator is provided with an additional tube with differentiating iterative circuits and summarizing of the signal and of its first derivative directly in the grids of the modulating tube.

Class 21a^4, 22_{05}. No.103974. G.I.Silkin and L.M.Shur. Method for Effective Countercoupling through the Envelopes in Radiotelephone Transmitters.

To increase the effectiveness of operation of the detector in a circuit of countercoupling through the envelopes of amplitude-modulated oscillations and to eliminate uncompensated additional distortions during overmodulation, it is proposed to provide the detector circuit with an additional carrier-frequency voltage, supplied from the exciter of the radio transmitter across a decoupling stage.

Class 21a^4, 71. No.103742. L. Ye.Leykhter. Method of Measurement of Pulse Amplitudes.

Class 21a^4, 71. No.104035. D.Ye.Rotenberg and Ye.B.Zaslavskaya. Device for Measuring Power in the Ultrahigh-Frequency Band.

Class 21c, 19_{07}. No.103482. Ye.P.Os'makov. Device for Packing Soil in Trenches made by the Knife Blade of a Cable-Laying Machine.

Class 21e, 36. No.103415. V.N.Itskhakin. Method of Measuring Kilometric Attenuation of Coaxial Cables.

Class 21e, 36. No.103552. Ye.I.Rozenfel'd, A.V.Naumov, L.Yu.Razmaninov, and G.I.Uryamzon. Device for Measuring Mean Values of Voltage of any Form.

Class 21e, 36. No.104002. Ye.A.Bodrov and O.V.Barablin. Method for Measuring the Peak Value of Alternating Voltages and Pulses in Oscillographs.

Class 42b, 14. No.103984. A.N.Zaslavskiy. Device for Drawing Time Marks on Oscillograph Tape.

Class 42k, 20_{03}. No.103602. A.I.Parfent'yev. Method for Measuring Magnetic Materials Used for Magnetic Sound Recording.

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FOREIGN PATENTS

From the Editor: In publishing brief data on certain foreign patents in the field of television technology, the editorial office informs readers that photostats of the patents (description in the original language, drawings) can be obtained in person c.o.d. from the All-Union Patent Technical Library of the Committee on Affairs of Invention and Discovery attached to the Council of Ministers USSR.

Address of the Library: Moscow-Center, Proyezd Serova 4, pod'yezd 7a. Telephone for information B8-64-52.

Photostats of the patents can be obtained also through the Mail Order Office of the All-Union Institute of Scientific and Technical Information of the Academy of Sciences USSR and Gostekhnika (State Technology).

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

Patent of Federal Republic of Germany, Class $21a^4$, 71, No.929740, 04.07.55. H.Wahl. Device for Visual Observation of the Spectrum of a High-Frequency Signal (H.Wahl - Telefunken G.m.b.H. - Einrichtung zur Sichtbarmachung der Frequenzverteilung eines hochfrequenten Frequenzgemisches.).

A device is proposed for visual observation of a high-frequency signal in a limited frequency band. In the device, one pair of deflecting plates of the cathode-ray tube is fed with the rectified voltage of a signal separated by means of a resonance circuit from the examined signal during smooth and periodic variation of the frequency of circuit tuning. Another pair of plates is fed with a variable DC voltage, tapped from a potentiometer, whose wiper rotates synchronously with the rotor of the element of resonance circuit tuning.

Patent of USA, Class 250-20, No.2711477, 21.06.55 Bassard. Input Unit for Television Receivers. (E.J.H.Bassard - Avco Manufacturing Corporation - Tuner for

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Television Receiver).

An input unit is proposed with smooth tuning in television channels in the 54 - 216 megacycle range. The input circuit, the circuit of the HF amplifier, and the heterodyne are tuned by means of variable inductances of the general tuning knob.

Patent of German Democratic Republic, Class $21a^4$, 69, No.10416, 22.09.55. Schnabel. A device for Noncontact Switching of Channels or Bands (Schnabel - SABA G.m.b.H. - Anordnung zur Kontaktfreien Kanal-oder Bereichumschaltung).

A method is proposed for noncontact switching of received channels or ultrashort-wave bands in a television receiver, by means of which the relative width of the transmission band can be varied simultaneously with the variation in the operating frequency range. Resonance circuits in tuned stages of the receiver remain resonant at a certain average frequency of the covered frequency range. Variation of the resonance frequency of each of the circuits is achieved by introducing reactance by the other circuit, inductively connected with it, which is detuned relative to its resonance frequency. The reactance introduced into the primary circuit may assume an inductive or capacitive character. The selection of the damping of the coupled circuit permits varying the width of the transmission band of the primary circuit.

Patent of Federal Republic of Germany, Class $21a^1$, 35_{20}, No.930761, 25.07.55. W.Schroeder. Device for Beam Deflection in Electron-Ray Tube (W.Schroeder - Telefunken G.m.b.H. - Schaltungsanordnung zur Ablenkung des Elektronenstrahls in einer Kathodenstrahlroehre).

An electromagnetic deflecting system for a kinescope is proposed, which is realized in the form of a bridge, whose four arms consist of identical deflecting coils. The generators of line scanning and vertical sweep are connected in the diagonals of the bridge. At the same time, the current flowing through the coils forms a horizontal and vertical deflecting field. Such a deflecting system can be realized in the form of a coil with toroidal windings, or in the form of a deflect-

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ing system of the "rotor" type.

Patent of Federal Republic of Germany, Class $21a^1$, 33_{70}, No.934050, 13.10.55.

W.Bruch,A.Scholz. Circuit for Television Set Operating by the Method of Beat Between Carriers (W.Bruch, A.Scholz - Telefunken G.m.b.H. - Schaltung fur Fernsehempfaenger nach dem Differenztraegerverfahren).

A circuit is proposed for joint amplification of the picture signals and the difference frequency in the receiver with sound reception by the method of heterodyning between the carriers of sound and picture. The plate circuit of the amplifier contains a series-connected parallel oscillatory circuit, connected directly to the anode, and an active load resistance shunted by a series oscillatory circuit. Both circuits are tuned to a difference frequency. The signal in the sound channel is taken from the anode of the tube, and the video signal from the point of junction of the parallel circuit and the load resistance. When using a pentode, the amplification factor in the difference frequency is approximately equal to 60. For the video channel without additional correcting elements, a uniform frequency characteristic to 5 mc is obtained at a load of about 500 ohm, with an amplification factor of 3 and a tenfold attenuation of the audio signals by the series circuit.

Patent of Federal Republic of Germany, $21a^1$, 34_{04}, No.940407, 15.03.56.

E.Legler. Method for Measuring the Magnitude of Sync Pulses in a Television Signal (E.Legler - Fernseh G.m.b.H. - Verfahren zur Messung der Groesse des Synchronpegels von Fernsehsignalen).

A circuit is proposed for measuring the magnitude of sync pulses in a video signal, which consists of two tubes connected in parallel, whose control grids are fed with the full video signal. By means of the usual locking circuits, the potential of the grid of the first tube is locked to the level of the peak of sync pulses, and the potential of the grid of the second tube, to the level of the peak of quenching pulses (the base of sync pulses). Between the anodes of the tubes a voltmeter is inserted. The tube operating conditions are so selected that, in the ab-

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sence of an input signal, the plate voltages of both tubes are equal, and the tubes operate in the linear sections of the characteristics. When a signal is present, the plate voltages of the tubes will be proportional to the levels of the grid locks, and the difference of these voltages, read from the voltmeter, will be proportional to the magnitude of the sync pulses.

Patent of Federal Republic of Germany, Class $21a^1$, 35_{10}, No.755948, 28.07.55. K.Schlesinger. Amplitude Filter for Separation of Picture and Synchronization Signals in Television (K.Schlesinger - Opta Radio A.G. - Amplitudensiebschaltung fuer die Trennung von Bildinhaltssignalen und Synchronisierzeichen beim Fernsehen).

A bridge circuit is proposed for the valve in the plate circuit of a video amplifier for amplitude separation of synchronization and picture signals, a circuit which neutralizes the effect of capacitance between plate and cathode of the valve. One arm of the bridge is formed by the valve connected to the anode of the amplifier and by one half of the primary of the transformer, the second half of which together with neutralization capacitor form the second arm. The threshold voltage on the valve is fed to the center tap of the primary from part of the amplifier load resistance across an RC filter with a time constant equal to the duration of a field, owing to which the voltage during a field remains constant. From the secondary winding of the transformer the horizontal sync pulses are tapped.

Patent of USA, Class 178 - 69.5, No.2726282, 06.12.55. J.Bigelow. System of Color Television Synchronization. (J.Bigelow - International Telephone and Telegraph Corp. - Television Synchronizing System).

A circuit is proposed for separation of color synchronization signals (flashes), which are transmitted in line quenching pulses after horizontal sync pulses, with the circuit containing a key tube. The circuit of the control grid of the key tube contains a loop, collision-excited by the horizontal sync pulses. The parameters of the loop are so designed that the second maximum of an intensely quenching oscillation coincides in time with the instant of passage of a flash in the signal, sup-

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plied to the grid of the key tube from the video amplifier. Under (normal condition, the key tube is closed. For its opening, apart from the impact oscillation, also the pulses of the line return from the line transformer can be used. This prevents passage of random pulses during the time between flashes.

Patent of USA, Class 178-5.2, No.2729697, 03.01.56. J.Chatten. System of Color Television (J.Chatten - Philco Corporation - Color Television System).

Selection is proposed of the composition of the transmitted color signal in systems of color television, with transmission of the brightness signal and two color-difference signals by means of quadrature modulation of the color subcarrier. Color-difference signals, P - Y and Q - Y, are proposed, selected with consideration of color vision and the maximum possible values of the signal of the color subcarrier in the transmission of a typical range of colors. With the new signals, the ellipses of color errors due to noises in the color-difference signals (arising in the transmission and receiver channels) coincide with the ellipses of the sensitivity of the average eye. This is achieved by proper selection of the axes of signals and their relative amplifications. The axes of new signals coincide also with the main axes of the ellipse of the maximum possible values of the subcarrier. The primary colors of the transmission P and Q, corresponding to the new signals, are equal in the system MKO: $x_p = 1.3$; $y_p = 0$; $x_a = 0.189$; $y_a = 0$.

Patent of Federal Republic of Germany, Class $21a^1$, 32_o, No.931052, 01.08.55. V.Duke. Device for Recording and Reproducing Television Signals (V.Duke - Radio Corporation of America - Einrichtung zur Aufzeichnung und Wiedergabe von Bildsignalen).

A device is proposed for recording television signals on motion-picture film from the screen of a kinescope. The film moves with a constant speed in the direction of line scanning on the kinescope screen, so that the lines of the picture are recorded separately. The picture of a complete frame is recorded in the form of two rhombic groups of lines, which represent the odd and even lines of the frame in

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alternate-line scanning. The reproduction is accomplished by means of a unit with scanning beam, during analogous movement of film. For elimination of possible mismatching of the kinescope lines of the scanning beam with the lines of the recording of the picture on the film, swinging of the beam of the kinescope in a vertical direction is applied.

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A brief report is given of the 1957 publishing program of technical literature which, in the Editor's opinion, might be of interest to the readers of "Elektrosvyazi".

Gosudarstvennoye Energeticheskoye Izdatel'stvo

(State Publishing House for Engineering)

K.M.Polivanov. Ferromagnetic Bodies in Constant and Variable Fields. 30 quires. Circulation 7000. Price 16 rubles 50 kopeks. Publication date - First Quarter.

L.A.Bessonov. Self-Excited Oscillations in Electric Circuits. 18 quires. Circulation 5000. Price 10 r 50 k. Publication date - First Quarter.

G.P.Tartakovskiy. Dynamics of Systems of Automatic Control of Amplification. 10 quires. Circulation 10,000. Price 6 r. Publication date - Fourth Quarter.

Van Der Ziel. Electric Noise. (Translation from English). 30 quires. Circulation 10,000. Price 16 r 50 k. Publication date - Second Quarter.

Kaden. Electromagnetic Screening in Technology of Wire and Radio Communications. (Translated from German). 19 quires. Circulation 10,000. Price 10 r 50 k. Publication date - Second Quarter.

Johns. Modern Phototelegraphic Equipment. (Translated from English). 15 quires. Circulation 5000. Price 8 r 50 k. Publication date - Third Quarter.

I.L.Belopolskiy. Power Supply for Radio Units. 12 quires. Circulation 10,000. Price 5 r 20 k. Publication date - Third Quarter.

Yu.A.Bulanov, S.A.Usov. Radio Receiving Units. 25 quires. Circulation 10,000. Price 9 r 75 k. Publication date - Fourth Quarter.

D.D.Sachkov, Ye.K.Eydlin. Design of Radio Equipment. 20 quires. Circulation 15,000. Price 8 r. Publication date - Fourth Quarter.

K.A.Shutskoy. Course and Diploma Planning of Radio Receiving Units.

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10 quires. Circulation 10,000. Price 4 r 50 k. Publication date - Fourth Quarter.

V.S.Bukler and others. Assembly of Radio Equipment.

20 quires. Circulation 15,000. Price 8 r. Publication date - Third Quarter.

A.A.Kulikovskiy. Linear Cascades of Radio Receivers.

27 quires. Circulation 10,000. Price 14 r 50 k. Publication date - Second Quarter.

I.I.Levenstern. Receiver of FM Signals.

5 quires. Circulation 7000. Price 2 r 50 k. Publication date - Second Quarter.

A.L.Kharinskiy. Foundations of Designing Elements of Radio Equipment.

35 quires. Circulation 10,000. Price 19 r. Publication date - Second Quarter.

I.G.Mamonkin. Pulse Amplifiers.

21 quires. Circulation 10,000. Price 12 r. Publication date - Third Quarter.

D.P.Linde. Elements of Calculating Generators of Ultrahigh Frequencies.

20 quires. Circulation 10,000. Price 11 r 50 k. Publication date - First Quarter.

N.P.Yermolin, A.P.Vaganov. Calculation of Low-power Transformers.

8 quires. Circulation 10,000. Price 4 r. Publication date - Third Quarter.

V.Yu.Roginskiy. Electric Supply of Radio-Technical Devices.

20 quires. Circulation 10,000. Price 11 r. Publication date - Second Quarter.

Ye.G.Mamot. Radio-Technical Measurements.

17 quires. Circulation 10,000. Price 8 r 50 k. Publication date - Third Quarter.

Ye.S.Antseliovich. Radio-Technical Measurements.

22 quires. Circulation 10,000. Price 12 r. Publication date - First Quarter.

N.N.Solov'yev. Bases of Measurement Technology in Wire Communications, Part 2.

25 quires. Circulation 10,000. Price 13 r 50 k. Publication date - Fourth Quarter.

Ya.I.Efrussi. New Circuits of Intermediate-Frequency Amplifiers in Television.

6 quires. Circulation 10,000. Price 3 r. Publication date - First Quarter.

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