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THE POSSIBLE TRANSMISSION BAND IN THE CASE OF LONG-DISTANCE TROPOSPHERIC PROPAGATION

by

V.N.Troitskiy

The question of the distortion of transmission in the case of long-distance tropospheric propagation is considered under the assumption that the atmosphere is anisotropic and that the horizontal inhomogeneities of the dielectric constant are smaller than the vertical inhomogeneities. Formulas are given for determining the band of the spectrum of communications that can be transmitted without distortion. The influence of antenna directivity on the possible transmission band is analyzed.

Determination of the band width of the communications spectrum that can be transmitted without distortion in the case of long distance tropospheric propagation is a problem of great practical importance for practice.

The fundamental cause of such distortions is the formation by the field at the place of reception of a large number of waves with unequal propagation times, owing to the different length of the path of each such wave. If we bear in mind that the troposphere is in fact anisotropic, and that the horizontal inhomogeneities of the dielectric constant are smaller than the vertical inhomogeneities, then, as has been shown in my previous paper (Bibl.1), the field of a tropospheric wave may be represented as the superposition of the waves separately reflected from inhomogeneities of different dimensions. In this case there will be a continuous spectrum of these inhomogeneities, as defined by the Kolmogorov-Obukhov law.

It has been shown (Bibl.1) that under this assumption the field strength of a tropospheric wave is defined by the equation

$$\overline{E^2} = C\int_0^\infty p^4 l^{\frac{1}{3}} e^{-4\pi p \sin\varphi_0} \int_0^\infty e^{-\frac{8\pi p}{R}x} dx, \qquad (1)$$

where C = const, $p = \frac{l}{\lambda}$, l = dimension of the inhomogeneity, λ = the wavelength, R = distance between stations, $\varphi_0 = \frac{\alpha}{2}$, α = geocentric angle between points of contact between the rays and the earth (Fig.1), x = height of point of reflection above the point at the limit of direct visibility on the side of the receiving antenna and on that of the transmitting antenna.

The time lag of the individual rays will be

$$t = \frac{2x \sin\varphi_0}{c}.$$

If originally the trailing front of the wave had an ideal form (Fig.2a), then this front will become blurred (Fig.2a) at the place of reception, owing to the arrival of a number of reflected waves with different lags. The value of the amplitude of the received signal at any point τ of the frailing front will be

$$\overline{E_\tau^2} = C\int_0^\infty p^4 l^{\frac{1}{3}} e^{-4\pi p \sin\varphi_0} \int_{x_\tau}^\infty e^{-\frac{8\pi p}{R}x} dx, \qquad (2)$$

where

$$x_\tau = \frac{c\tau}{2\sin\varphi_0}. \qquad (3)$$

Integrating, by analogy to our earlier procedure (Bibl.1), we get

$$\overline{E}_\tau = 0{,}19\sqrt{F_0}\,\frac{\sqrt{R}\,\lambda^{\frac{1}{3}}}{\sin^{\frac{11}{6}}\varphi_0}\,\frac{E_0}{\left[1+\frac{c\tau}{R\sin^2\varphi_0}\right]^{\frac{11}{6}}}, \qquad (4)$$

where E_0 = field in free space, and F_0 = quantity characterizing the intensity of the inhomogeneity.

Thus, by eq.(4), the front of the received signal of the tropospheric wave will be of the form defined by the equation:

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$$\bar{E}_\tau = \frac{\bar{E}}{\left(1+\frac{c\tau}{R\sin^3\varphi_0}\right)^{\frac{11}{6}}}, \tag{5}$$

where $\bar{E}$ = amplitude of tropospheric wave under stationary conditions.

Consequently the medium in which long-distance tropospheric propagation occurs acts on the signal in about the same way as a certain band filter. It is well known that a band filter with a finite pass band also causes blurring of the wave front.

Fig.1

Fig.2

This blurring is usually characterized by the time of variation of the amplitude of the output signal from 0.1 to 0.9 of its steady-state value. In the case of an ideal filter with a rectangular frequency characteristic, it is commonly known (Bibl.2) that the lag time τ_{max} is inversely proportional to the band width of the filter

$$\tau_{max} = \frac{0,86}{\Delta f}. \tag{6}$$

In our case the lag time may be easily determined from the equation

$$\tau_{max} \approx 2,5\frac{R\sin^3\varphi_0}{c}. \tag{7}$$

On determining the value of Δf from eq.(6), and substitution of the value of τ_{max} from eq.(7), we get the "pass band of the medium" in the case of tropospheric propagation:

$$\Delta f_{max} = 0,343\frac{c}{R\sin^3\varphi_0}. \tag{8}$$

It is easy to determine the angle φ_0 from geometrical considerations, if we know the distance and height of the antennas:

$$\sin\varphi_0 = \frac{R}{2a_e}(1-q), \tag{9}$$

where

a_e = equivalent radius of the earth,

$$q = \frac{R_1 + R_2}{R}$$

R_1, R_2 = distance from horizon to transmitting and receiving antennas respectively (Fig.1).

On substituting the value of $\sin\varphi_0$ from eq.(9) in eq.(8), we obtain:

$$\Delta f_{max} = 1,37\frac{ca_e^2}{R^2(1-q)^3}. \tag{10}$$

We have until now been considering the distortion of a signal for the case of not very strongly directional antennas, when the angle of directivity is markedly greater than the angle φ_0, so that the directivity of the antennas could be neglected. Let us now consider the general case of directional antennas, assuming that $f_1(\alpha_1)$ and $f_2(\alpha_2)$ are the characteristics of directivity of the transmitting and receiving antenna (α_1 and α_2 are angles measured from the principal direction).

Then eq.(2) takes the form:

$$\bar{E}^2 = C\int_0^\infty p^4 l^{-\frac{1}{3}} e^{-4\varepsilon p\sin\varphi_0}\,dl \int_{x_\tau}^\infty f_1^2(\alpha_1)\, f_2^2(\alpha_2)\, e^{-\frac{8\varepsilon p}{R}x}\,dx. \tag{11}$$

By geometrical reasoning we get

$$\alpha_1 \approx \alpha_2 \approx \frac{2x}{R}. \tag{12}$$

For an approximate evaluation of the influence of the antennas on the pass

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band, we shall approximate the directional characteristics of the antenna in the form of the right triangles of width α_0, where α_0 = angle of directivity of antennas.

In that case the integral of eq.(11) will be of the form

$$\overline{E_\tau^2} = C\int_0^\infty p^4 l^{-\frac{1}{3}} e^{-4\pi p \sin\varphi_0} dl \int_{x_\tau}^{x_{\alpha_0}} e^{-\frac{8\pi p}{R}x} dx, \qquad (13)$$

where

$$x_{\alpha_0} = \frac{\alpha_0 R}{4}.$$

Integration is performed in the same way as before.

As a result we get

$$\overline{E}_\tau = 0,19\sqrt{F_0}\,\frac{\sqrt{R}\,\lambda^{\frac{1}{3}}}{\sin^{\frac{11}{6}}\varphi_0}\sqrt{\frac{1}{\left(1+\frac{c\tau}{R\sin^2\varphi_0}\right)^{\frac{11}{3}}} - \frac{1}{\left(1+\frac{\alpha_0}{2\sin\varphi_0}\right)^{\frac{11}{3}}}}. \qquad (14)$$

The form of the wave front is defined by the equation:

$$\frac{\overline{E}_\tau}{\overline{E}} = \sqrt{\frac{\left(1+\frac{\alpha_0}{2\sin\varphi_0}\right)^{\frac{11}{3}} - \left(1+\frac{c\tau}{R\sin^2\varphi_0}\right)^{\frac{11}{3}}}{\left(1+\frac{c\tau}{R\sin^2\varphi_0}\right)^{\frac{11}{3}}\left[\left(1+\frac{\alpha_0}{2\sin\varphi_0}\right)^{\frac{11}{3}}-1\right]}}. \qquad (15)$$

From the condition $\frac{\overline{E}_\tau}{\overline{E}} = 0.1$, we find τ_{max}:

$$\tau_{max} = \frac{R\sin^2\varphi_0}{c}\left\{\left(1+\frac{\alpha_0}{2\sin\varphi_0}\right)\left[10^{-2}\left[\left(1+\frac{\alpha_0}{2\sin\varphi_0}\right)^{\frac{11}{3}}-1\right]+1\right]^{-\frac{3}{11}}-1\right\}. \qquad (16)$$

Consequently,

$$\Delta f_\alpha = \Delta f\, Q\left(\frac{\alpha_0}{\sin\varphi_0}\right), \qquad (17)$$

Under such conditions the band Δf varies with time. The value of these variations depends on how strongly the actual spectrum of inhomogeneities at the individual instants of time differs from the Kolmogorov spectrum. If there are gaps in the spectrum, then selective fading is possible, leading to additional distortion.

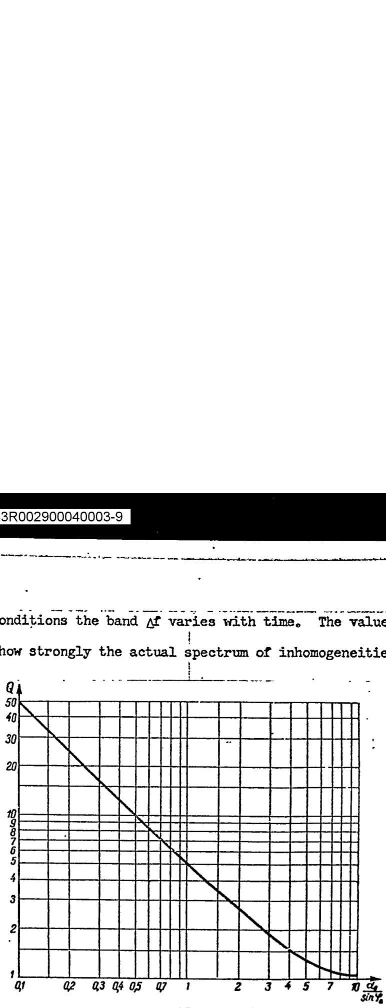

Fig.3

Paper received by editors 2 April 1956

BIBLIOGRAPHY

1. Troitskiy,V.N. - This Journal. 11 (5) (1956)

2. Siforov,V.I. - Radio Receiving Equipment, Chapter 16. 1954

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where

Δf_α = pass band in the case of directional antennas,

Δf = pass band in the case of nondirectional antennas.

$$Q\left(\frac{\alpha_0}{\sin\varphi_0}\right)=2,5\left\{\left(1+\frac{\alpha_0}{2\sin\varphi_0}\right)\left[10^{-2}\left[\left(1+\frac{\alpha_0}{2\sin\varphi_0}\right)^{\frac{11}{3}}-1\right]+1\right]^{-\frac{3}{11}}-1\right\}^{-1}. \quad (18)$$

The factor $Q\left(\frac{\alpha_0}{2\sin\varphi_0}\right)$ defines the influence of antenna directivity.

For $\frac{\alpha_0}{\sin\varphi_0} \gg 1$,

$$Q\left(\frac{\alpha}{\sin\varphi_0}\right)=1.$$

Figure 3 shows the relation between the value of Q and the parameter $\frac{\alpha_0}{\sin\varphi_0}$. Under the condition that $\frac{\alpha_0}{\sin\varphi_0}$ 1, we may obtain from eq.(18):

$$Q=5\,\frac{\sin\varphi_0}{\alpha_0}. \quad (19)$$

Consequently, in this case

$$\Delta f_\alpha=1,77\,\frac{c}{\alpha_0 R\sin\varphi_0}=3,54\,\frac{ca_e}{R^2(1-q)}. \quad (20)$$

As will be seen from eqs.(10) and (17), the pass band of the medium depends sharply on the distance. In the case of actual traffic lines 200-300 km long, this band is rather wide. Thus a communication can be transmitted with a rather wide spectrum by tropospheric propagation without appreciable distortions.

It must, however, be noted that all these conclusions are true only for average conditions, since the assumption that the spectrum of inhomogeneities is of Kolmogorov character is evidently true only on the average. At individual instants of time the spectrum of inhomogeneities differs significantly from the Kolmogorov spectrum.

Under such conditions the band Δf varies with time. The value of these variations depends on how strongly the actual spectrum of inhomogeneities at the individual instants of time differs from the Kolmogorov spectrum. If there are gaps in the spectrum, then selective fading is possible, leading to additional distortion.

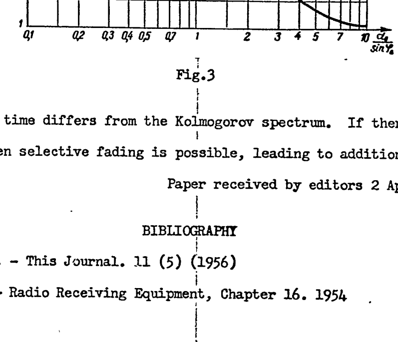

Fig.3

Paper received by editors 2 April 1956

BIBLIOGRAPHY

1. Troitskiy,V.N. - This Journal. 11 (5) (1956)
2. Siforov,V.I. - Radio Receiving Equipment, Chapter 16. 1954

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A METHOD OF CALCULATING PROPAGATION CONSTANTS IN WAVEGUIDES WITH NON-IDEALLY CONDUCTING WALLS

by

L.N.Loshakov,

Full Member of the Society

The author describes an approximate method of calculating propagation constants in waveguides with non-ideally conducting walls, based on the application of the conjugate lemma, gives a confirmation of the correctness of the method, and considers the results of its application.

1. It is well known that the rigorous theory of the propagation of waves in waveguides with non-ideally conducting walls is based on the solution of the Maxwell equations and the application of the Shchukin-Leontovich boundary condition, which is true for sufficiently high conductivity of the material of the wall. Such theory, however, cannot be extended to the case of a waveguide of rectangular cross section. The method of approximate calculation of damping in waveguides, which is termed the skin-effect method (Bibl.1), and is less rigorous, but more general, has therefore enjoyed widespread use in engineering practice. This method yields the damping constant for operation of a waveguide at frequencies higher than the critical, but does not permit calculation of that constant for the critical frequency, nor at all of the propagation constant in a waveguide.

In an earlier paper (Bibl.2) I have pointed out the possibility of approximate calculation of the propagation constant for a waveguide by the aid of a method based on the conjugate lemma. The proposed method possesses considerable simplicity and generality, and involves no restrictions on the working frequencies.

In the belief that this method might find practical application, and in view of the absence of any discussion of the possibilities of that method in my earlier

paper (Bibl.2), I present below a brief verification of the accuracy of the method and give certain results obtained by its use.

We note that the results given below may also be found by means of the Umov-Poynting complex theorem; and that method of calculation has been discussed in a work by G.V.Kisun'ko (Bibl.3); but the "lemma method" described below is more convenient.

2. It follows from my earlier paper (Bibl.2) that by applying the conjugate lemma (the Lorentz lemma, written on the replacement of one of two independent fields by a complex conjugate field) to the element of length dz of the transmission line we may obtain the starting relation:

$$\frac{\partial}{\partial z}\int_{s}([\overline{E}_{I}\overline{H}_{II}^{*}]+[\overline{E}_{II}^{*}\overline{H}_{I}])\overline{a}_{z}\,ds=-\int_{L}[\overline{E}_{I}\overline{H}_{II}^{*}]\overline{n}\,dl. \qquad (1)$$

The following notation was used in writing this relation:

$\overline{E}_1$, $\overline{H}_1$ = electric and magnetic fields in the line, taking account of the non-ideal conductivity of the walls (the field sought), $\overline{E}_{11}$, $\overline{H}_{11}$ = electric and magnetic fields in the same line with ideal conductivity of the walls (the auxiliary field), s = cross section of the line, L = contour of the walls, $\overline{a}_z$ = unit vector of longitudinal axis (z axis), and $\overline{n}$ = unit vector of internal normal to wall.

The sign * denotes complex conjugate quantities.

In the case under consideration, the transmission line is a waveguide, so that the auxiliary field consists of waves of type E or H.

If we postulate that the non-ideal conductivity has no influence on the transverse structure of the fields, and assume the components of the required and auxiliary fields to differ only in the dependence on the coordinate z, then we may obtain from eq.(1), by using the transformations described in my earlier paper (Bibl.2), an equation for the propagation constant γ_0 of the required field (of one of the possible waves) in the form

$$\gamma_0^2+\beta_0^2=\frac{\beta_0}{2P_0}\sqrt{\frac{\omega\mu_1}{2\sigma_1}}\,(i-1)\int_{L}|\overline{H}_0|^2\,dl. \qquad (2)$$

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Here β_0, P_0 are respectively the phase constant and the mean power flux with time, in the wave guide, with ideal conductivity of the walls, ω = angular frequency of field (the relation between the time t and the coordinate z is given by the factor $e^{-i(\omega t-\beta_0 t)}$), $|\overline{H}_0|$ = amplitude of magnetic field, in the absence of losses, μ_1, σ_1 = magnetic permeability and electric conductivity of the material of the wall.

In writing eq.(2) and hereafter, the rational practical system of units is used.

The correctness of the equation so obtained may be confirmed by comparison with the known results of the rigorous theory.

For the wave E_{on} in a circular waveguide, solution of the Maxwell equation, using the Shchukin-Leontovich boundary condition, yields, in first approximation (cf., for example, eq.(13) of Reference (Bibl.4)):

$$\gamma_0^2 + \beta_0^2 = \frac{2\omega\varepsilon}{a}\sqrt{\frac{\omega\mu_1}{2\sigma_1}}(i-1), \qquad (3)$$

where a = radius of cross section of waveguide.

Bearing in mind that in this case

$$\int_L |\overline{H}_0|^2\, dl = \frac{2\pi a\,\omega^2\varepsilon^2}{\kappa^2 - \beta_0^2} A^2 J_1^2(\nu_{on}),$$

where $\kappa^2 = \omega^2\varepsilon\mu$, ε, μ = parameters of the medium filling the waveguide, A = amplitude coefficient, and $J_1(\nu_{on})$ = Bessel function whose argument is a root of the equation $J_0(\nu_{on}) = 0$, while

$$P_0 = \frac{\pi a^2\beta_0\omega\varepsilon}{2(\kappa^2-\beta_0^2)} A^2 J_1^2(\nu_{on}),$$

we find that eqs.(2) and (3) are in complete agreement.

When eq.(2) is to be used for practical purposes, we put

$$\gamma_0 = -\alpha - i\beta.$$

Then we get, instead of eq.(2), the following equations:

$$\beta^2 - \alpha^2 = \beta_0^2 + \frac{\beta_0}{2P_0}\sqrt{\frac{\omega\mu_1}{2\sigma_1}}\int_L |\overline{H}_0|^2\, dl = \beta_0^2 + N_1^2 = N_2^2, \qquad (4)$$

$$2\alpha\beta = \frac{\beta_0}{2P_0}\sqrt{\frac{\omega\mu_1}{2\sigma_1}}\cdot\int_L |\overline{H}_0|^2\, dl = N_1^2, \qquad (5)$$

where β^2 and $-\alpha^2$ are the roots of the quadratic equation

$$x^2 - N_2^2 x - \frac{N_1^4}{4} = 0. \qquad (6)$$

By solving this equation, we find

$$\alpha = \pm\sqrt{\frac{-N_2^2 + \sqrt{N_2^4 + N_1^4}}{2}}, \qquad (7)$$

$$\beta = \pm\sqrt{\frac{N_2^2 + \sqrt{N_2^4 + N_1^4}}{2}}. \qquad (8)$$

The two signs correspond to the direct and reflected wave, and for each wave $\sigma\beta > 0$ should hold true.

It is easy to establish that the damping constant α_0 obtained by the skin-effect method is connected with the quantity N_1^2 by the relation

$$N_1^2 = 2\alpha_0\beta_0, \qquad (9)$$

whence

$$\alpha\beta = \alpha_0\beta_0. \qquad (10)$$

For a frequency above the critical value, $\beta_0 \gg \alpha_0$ and $N_2^2 \gg N_1^2$, and therefore, when the formula for approximate calculation

$$(1+q)^{\frac{1}{2}} = 1 + \frac{1}{2}q - \frac{1}{8}q^2;\ q \ll 1$$

is used, we obtain, from eqs.(7) and (8),

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$$\alpha = \alpha_0; \ \beta = \beta_0 + \alpha_0. \qquad (11)$$

At the critical frequency, $\beta_0 = 0$, $N_1^2 = N_2^2$, and eqs.(7) and (8) give us:

$$\alpha = \sqrt{\frac{\sqrt{2}-1}{2} N_1^2}; \qquad \beta = \sqrt{\frac{\sqrt{2}+1}{2} N_1^2}. \qquad (12)$$

Finally, for frequencies less than the critical, $\beta_0^2 < 0$. For this reason $N_2^2 < 0$, if $|\beta_0^2| > N_1^2$. In the case where $N_2^2 \gg N_1^2$, eqs.(7) and (8) pass over into the expressions

$$\alpha = |N_2| \approx |\beta_0|; \ \beta = \frac{N_1^2}{2\alpha} = \frac{N_1^2}{2|N_2|}. \qquad (13)$$

For the air dielectric,

$$\beta_0^2 = \frac{4\pi^2}{\lambda_{cr}^2}\left[\left(\frac{\lambda_{cr}}{\lambda}\right)^2 - 1\right], \qquad (14)$$

where λ = excitation wavelength, and λ_{cr} = critical wavelength of the waveguide.

Thus the "lemma method" allows us to determine the damping constant α and the phase constant β without restrictions with respect to the frequency, which may be either less than the critical or greater than it. It follows from eqs.(13)-(14) that for $\lambda \to \infty$ the damping constant α approaches a constant value equal to $\frac{2\pi}{\lambda_{cr}}$.

To give a quantitative illustration of the results, a numerical calculation of α and β as functions of λ was worked out for a specific case. As an example we took a rectangular waveguide operating on the wave H_{10}, having the cross section a = 72 mm and b = 34 mm. It was assumed that the waveguide had an air dielectric and copper walls (conductivity $\sigma_1 = 5.8 \times 10^7 \ \text{ohm}^{-1} \ \text{m}^{-1}$).

For this case the auxiliary quantities N_1^2 and B_0^2 were calculated by eqs.(15) and (16):

$$\frac{N_1^2}{2} = \sigma_0 \beta_0 = \frac{0.169 \cdot 10^{-4}\pi}{a^2\sqrt{a}}\left(\frac{a}{2b} \cdot \frac{\lambda_{cr}}{\lambda}\sqrt{\frac{\lambda_{cr}}{\lambda}} + \sqrt{\frac{\lambda}{\lambda_{cr}}}\right), \qquad (15)$$

$$\beta_0^2 = \frac{\pi^2}{a^2}\left[\left(\frac{\lambda_{cr}}{\lambda}\right)^2 - 1\right]; \ \lambda_{cr} = 2a, \qquad (16)$$

where a is expressed in meters.

The results of the calculation are given in the Table.

$\frac{\lambda_{cr}}{\lambda}$	$\frac{1}{2}N_1^2$	β_0^2	$\lvert\beta_0\rvert$	$\sigma_0 = \frac{N_1^2}{2\lvert\beta_0\rvert}$	α	β
1,10	0,083	399,81	19,995	0,00415	α_0	19,999
1,05	0,081	195,14	13,969	0,00 78	α_0	13,975
1,02	0,079	76,92	8,770	0,00 06	0,00903	8,779
1,01	0,079	38,27	6,186	0,01277	0,01277	6,191
1,00	0,078	0	0	∞	0,1804	0,4356
0,98	0,078	− 75,39	8,683	0,00896	8,674	0,00897
0 95	0,077	−185,63	13,624	0,00562	13,619	α_0
0 75	0,070	− 832,94	28,861	0,00244	28,858	α_0
0 50	0,068	−1427,9	37,788	0,00181	37,786	α_0
0,20	0,089	− 1827,7	42,7 2	0,00208	42,749	α_0
0,10	0,122	−1884,8	43,414	0,00281	43,412	α_0

A glance at the Table will show that eqs.(11) and (13) may be used over a wide range of variation of λ, and only in the immediate vicinity of λ to λ_{cr} should the general formulas eqs.(7) - (8) be used,

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BIBLIOGRAPHY

1. Vvedenskiy,B.A. and Arenberg,A.G. - Radio Wave Guides. ONTI, Moscow, 1946
2. Loshakov,L.N. - Zhur.Tekh.Fiz. 26(4) (1956)
3. Kisun'ko,G.V. - Electrodynamics of Hollow Systems. 1949
4. Loshakov,L.N. - Zhur.Tekh.Fiz. 23(10) (1953)

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THE EFFECT OF PRECIPITATION ON THE ELECTRICAL PROPERTIES OF WIRE ANTENNA ARRAY SURFACES

by

V.K.Paramonov

The effect of icing on the reflecting properties of the wire surfaces of antenna arrays is analyzed. Formulas are derived by which the transmission factor through an antenna array, each wire of which is covered with a uniform layer of ice, can be calculated. The results of experimental studies are given, confirming the correctness of the formulas obtained.

1. Introduction

In ultrashort-wave antenna engineering, wide use is made of noncontinuous reflecting surfaces designed in the form of wire arrays or of perforated metal surfaces. The use of noncontinuous reflectors is usually due to the effort to reduce the weight or wind resistance of the reflector.

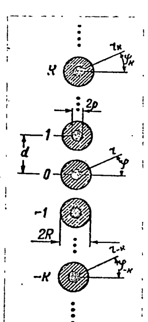

Fig.1

In winter, reflectors, like other equipment and structures, are subject to the action of atmospheric precipitation (sleet and frost). The presence of such rainfall on a reflector modifies its electrical properties.

The present work is devoted to a study of the question of the influence of rainfall on the electrical properties of a single-line antenna array of wires of circular cross section; and each wire of the array under consideration is assumed to be covered with a uniform layer of ice, i.e., the wire and the ice layer form coaxial cylinders. In spite of a certain idealization, this case is very close to an actually occurring case, and is of definite practical interest.

2. Normal Incidence

Let a plane wave impinge on a single-line wire array, each wire of which is covered with a coaxial layer of ice. We assume the array to consist of an infinite number of equidistant wires of infinitely long extension. We number the wires of the array from $-\infty$ to $+\infty$ and attach the cylindrical coordinate system r_k, φ_k, z_k to each wire (Fig.1).

The secondary field excited by the k-th wire, may be written in the following way (Bibl.1):

$$E_1^{(\kappa)} = E_0 \sum_{m=-\infty}^{\infty} Q_m H_m^{(2)}(\alpha r_\kappa)\, e^{i m \varphi_\kappa} \quad \text{at} \quad r_\kappa > R, \tag{1}$$

where E_0 = amplitude of incident field, $\alpha = \frac{2\pi}{\lambda}$ = phase constant, R = radius of ice layer, $H_m^{(2)}(z)$ = cylindrical Hankel function, Q_m = coefficients determined from the boundary conditions and not depending on the number of wires.

The total secondary field is defined by the radiation of all the wires

$$E_{vm} = \sum_{\kappa=-\infty}^{\infty} E_1^{(\kappa)}. \tag{2}$$

By summating the field of all the wires in the distant zone, in the direction of propagation of the incident wave (for instance, as is done in the References of (Bibl.2) or (Bibl.3), we may obtain:

$$E_{vm} = E_0 \frac{\lambda}{\pi d} \sum_{m=-\infty}^{\infty} (-i)^m Q_m. \tag{3}$$

The transmission factor (for power) may therefore be written in the form

$$\delta = \left| 1 + \frac{\lambda}{\pi d} \sum_{m=-\infty}^{\infty} (-i)^m Q_m \right|^2. \tag{4}$$

To find the coefficients Q_m, let us make use of the boundary conditions: co

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tinuity of the tangential components of the intensity vectors of the magnetic and electric fields on the surface of the wire, and vanishing of the tangential component of the vector of electric field intensity on the surface of the wire:

$$E_{inc} + E_{vm} = E_2^{(\kappa)} \quad at \quad r_\kappa = R, \qquad (5)$$

$$H_{inc} + H_{vm} = H_2^{(\kappa)} \quad at \quad r_\kappa = R, \qquad (6)$$

$$E_2^{(\kappa)} = 0 \quad at \quad r_\kappa = \rho, \qquad (7)$$

where E_{inc} and H_{inc} are the respective intensities of the electric and magnetic fields of the incident wave, ρ = radius of the wire, $E_2^{(k)}$ and correspondingly $H_2^{(k)}$ = field intensity inside the ice layer, which may be written as follows:

$$E_2^{(\kappa)} = E_0 \sum_{m=-\infty}^{\infty} [R_m J_m(\alpha' r_\kappa) + i S_m N_m(\alpha' r_\kappa)] e^{i m \varphi_\kappa} \qquad (8)$$

$$at \quad \rho < r_\kappa < R,$$

where $\alpha' = \alpha\sqrt{\varepsilon_l}$ = phase constant for the propagation of the wave in the ice; ε_l = 3.2 = dielectric constant of ice, $J_m(z)$ and $N_m(z)$ = cylindrical Bessel and Neumann constants, R_m and S_m = coefficients determined from the boundary conditions and not depending on the number of wires.

The condition of eq.(6) may obviously be written as follows:

$$\frac{\partial E_{inc}}{\partial r_\kappa} + \frac{\partial E_{vm}}{\partial r_\kappa} = \frac{\partial E_2^{(\kappa)}}{\partial r_\kappa}. \qquad (6a)$$

By the aid of the theorem of composition of cylindrical functions (Bibl.4), the field excited by the k-th wire ($E_1^{(k)}$) in the coordinate system of the zero-th wire (r and φ) may be written in the form:

$$E_1^{(\kappa)} = \sum_{m=-\infty}^{\infty} Q_m \sum_{n=-\infty}^{\infty} H_n^{(2)}(\alpha\kappa d) J_{m+n}(\alpha r) \cos\left(n\frac{\pi}{2}\right) e^{i(m+n)\varphi}, \qquad (9)$$

Using eq.(9), the boundary condition, eq.(5), may be written in the expanded form:

$$\sum_{m=-\infty}^{\infty} i^m J_m(\alpha R) e^{i m \varphi} + \sum_{m=-\infty}^{\infty} Q_m H_m^{(2)}(\alpha R) e^{i m \varphi} +$$

$$+ 2 \sum_{\kappa=1}^{\infty} \sum_{m=-\infty}^{\infty} Q_m \sum_{n=-\infty}^{\infty} H_n^{(2)}(\alpha\kappa d) J_{m+n}(\alpha R) \cos\left(n\frac{\pi}{2}\right) e^{i(m+n)\varphi} =$$

$$= \sum_{m=-\infty}^{\infty} [R_m J_m(\alpha' R) + i S_m N_m(\alpha' R)] e^{i m \varphi}. \qquad (10)$$

The first series on the left side of eq.(10) is the expansion of the incident plane wave in cylindrical functions (Bibl.4).

The conditions eqs.(6a) and (7) are similarly written in the expanded form as follows:

$$\sum_{m=-\infty}^{\infty} i^m J'_m(\alpha R) e^{i m \varphi} + \sum_{m=-\infty}^{\infty} Q_m H_m^{(2)\prime}(\alpha R) e^{i m \varphi} +$$

$$+ 2 \sum_{\kappa=1}^{\infty} \sum_{m=-\infty}^{\infty} Q_m \sum_{n=-\infty}^{\infty} H_n^{(2)}(\alpha\kappa d) J'_{m+n}(\alpha R) \cos\left(n\frac{\pi}{2}\right) e^{i(m+n)\varphi} =$$

$$= \eta \sum_{m=-\infty}^{\infty} [R_m J'_m(\alpha' R) + i S_m N'_m(\alpha' R)] e^{i m \varphi}, \qquad (11)$$

$$\sum_{m=-\infty}^{\infty} [R_m J_m(\alpha' \rho) + i S_m N_m(\alpha' \rho)] e^{i m \varphi} = 0, \qquad (12)$$

where

$$\eta = \frac{\alpha'}{\alpha} = \sqrt{\varepsilon_l}. \qquad (13)$$

On equating the terms in eqs.(10) and (11) with the same exponents, we get two infinite systems of equations. Thus, from eq.(10) we obtain a system of equations of the form:

$$i^m J_m(\alpha R) + Q_m H_m^{(2)}(\alpha R) + 2 J_m(\alpha R) \sum_{n=-\infty}^{\infty} Q_{m-n} \cos\left(n\frac{}{2}\right) \sum_{\kappa=1}^{\infty} H_n^{(2)}(\alpha\kappa d) =$$

$$= R_m J_m(\alpha' R) + i S_m N_m(\alpha' R), \qquad (14)$$

where m = ... -2, -1, 0, 1, 2, ...

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and from eq.(11):

$$i^m J'_m(\alpha R) + Q_m H_m^{(2)'}(\alpha R) + 2 J'_m(\alpha R) \sum_{n=-\infty}^{\infty} Q_{m-n} \cos\left(n \frac{\pi}{2}\right) \sum_{\kappa=1}^{\infty} H_n^{(2)}(\alpha \kappa d) = \eta \left[R_m J'_m(\alpha' R) + i S_m N'_m(\alpha' R)\right], \tag{15}$$

where m = ... -2, -1, 0, 1, 2, ...

Making use of the well-known property of the Hankel function

$$H_{-m}^{(2)}(z) = (-1)^m H_m^{(2)}(z),$$

we can show that for

$$\left.\begin{array}{ll} m - even & Q_{-m} = Q_m \\ m - odd & Q_{-m} = -Q_m \end{array}\right\} \tag{16}$$

Taking eqs.(16) into account, an infinite system of inhomogeneous equations is obtained from eq.(14). The solution of this system may be obtained by the method of successive approximations, provided that the process of approximation converges.

If for simplicity we confine ourselves to only the first approximation, then the following value may be obtained for Q_m:

for m ≠ 0

$$Q_m = -i^m \frac{J_m(\alpha R) - R_m J_m(\alpha' R) - i S_m N_m(\alpha' R)}{H_m^{(2)}(\alpha R) + 2 J_m(\alpha R) \sum_{\kappa=1}^{\infty} \left[H_0^{(2)}(\alpha \kappa d) + H_{2m}^{(2)}(\alpha \kappa d)\right]}, \tag{17}$$

for m = 0

$$Q_0 = -\frac{J_0(\alpha R) - R_0 J_0(\alpha' R) - i S_0 N_0(\alpha' R)}{H_0^{(2)}(\alpha R) + 2 J_0(\alpha R) \sum_{\kappa=1}^{\infty} H_0^{(2)}(\alpha \kappa d)}. \tag{18}$$

In exactly the same way the following expression for Q_m may be obtained from the system (15):

for m ≠ 0,

$$Q_m = -i^m \frac{J'_m(\alpha R) - \eta \left[R_m J'_m(\alpha' R) + i S_m N'_m(\alpha' R)\right]}{H_m^{(2)'}(\alpha R) + 2 J'_m(\alpha R) \sum_{\kappa=1}^{\infty} \left[H_0^{(2)}(\alpha \kappa d) + H_{2m}^{(2)}(\alpha \kappa d)\right]}, \tag{19}$$

for m = 0,

$$Q_0 = -\frac{J'_0(\alpha R) - \eta \left[R_0 J'_0(\alpha' R) + i S_0 N'_0(\alpha' R)\right]}{H_0^{(2)'}(\alpha R) + 2 J'_0(\alpha R) \sum_{\kappa=1}^{\infty} H_0^{(2)}(\alpha \kappa d)}. \tag{20}$$

In addition, we also have from eq.(12):

$$R_m J_m(\alpha' \rho) + i S_m N_m(\alpha' \rho) = 0. \tag{21}$$

Eqs.(17), (19) and (21) for a system of equations in the unknowns Q_m, R_m and S_m. By solving this system for Q_m, the following value for it may be obtained:

$$Q_m = -i^m \frac{J'_m(\alpha R) U_m - J_m(\alpha R) V_m}{K_m U_m - L_m V_m}, \tag{22}$$

where

$$\left.\begin{array}{l} U_m = N_m(\alpha' R) J_m(\alpha' \rho) - N_m(\alpha' \rho) J_m(\alpha' R) \\ V_m = \eta \left[N'_m(\alpha' R) J_m(\alpha' \rho) - N_m(\alpha' \rho) J'_m(\alpha' R)\right] \end{array}\right\}. \tag{23}$$

The coefficients K_m and L_m, for m ≠ 0, are equal to:

$$\left.\begin{array}{l} K_m = H_m^{(2)'}(\alpha R) + 2 J'_m(\alpha R) \sum_{\kappa=1}^{\infty} \left[H_0^{(2)}(\alpha \kappa d) + H_{2m}^{(2)}(\alpha \kappa d)\right] \\ L_m = H_m^{(2)}(\alpha R) + 2 J_m(\alpha R) \sum_{\kappa=1}^{\infty} \left[H_0^{(2)}(\alpha \kappa d) + H_{2m}^{(2)}(\alpha \kappa d)\right] \end{array}\right\}, \tag{24}$$

and for m = 0 to

$$\left.\begin{array}{l} K_0 = H_0^{(2)'}(\alpha R) + 2 J'_0(\alpha R) \sum_{\kappa=1}^{\infty} H_0^{(2)}(\alpha \kappa d) \\ L_0 = H_0^{(2)}(\alpha R) + 2 J_0(\alpha R) \sum_{\kappa=1}^{\infty} H_0^{(2)}(\alpha \kappa d) \end{array}\right\} \tag{25}$$

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As shown by calculations, for those wavelengths, wire diameters and coating diameters, and also for those distances between the wires, that are usually found in actual situations, the value of Q_m decreases sharply with increasing index of summation m. In most practical cases, this fact allows us, in calculating the transmission factor, to confine ourselves to only the first term of the infinite series (Q_o) and thereby considerably to simplify all the calculations. Thus

$$\delta = \left|1 + \frac{\lambda}{\pi d} Q_0\right|^2. \qquad (26)$$

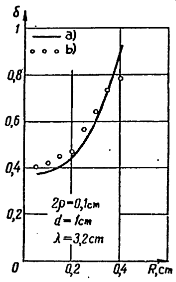

Fig.2

a) Calculated; b) Experimental

Figures 2 and 3 show graphs, calculated by eq.(26), of the relation between the transmission factor and the radius of the cylinder of ice coating. The calculation has been made for two cases characterized by different wire diameters and different distances between the wires of the array.

As will be seen from the graphs, when ice appears on the wires of the array, and with further increase in the thickness of the ice layer, within certain limits (up to about R = 0.1), the transmission factor increases, and very significantly at that.

As shown by calculations, further increase in the thickness of the ice coating leads to a fall in the transmission factor.

This variation of the transmission factor may be explained as follows. The increase in the thickness of the ice coating has two consequences. On the one hand, part of the space between the wires of the array is filled by a dielectric, which is equivalent to a certain shortening of the wavelength of the wave incident on the array. In this case the transmission factor increases. On the other hand, the reflecting surface of the array is also increased, leading to a reduction in the transmission factor. The relative predominance of either of these two tendencies determines the character of the variation of the transmission factor within the given limits of variation of the thickness of the ice coating.

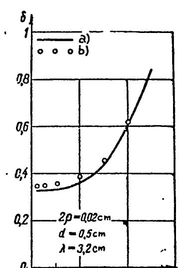

Fig.3

a) Calculated; b) Experimental

Thus the presence of a uniform ice layer on the wires of a reflector array adversely affects its reflecting properties. At certain values of the ice layer thickness the transmission factor increases so greatly, even for strongly reflecting arrays, that the array almost completely loses its reflecting properties.

3. Oblique Incidence of Wave on Array

Let us now consider the cases which are of the greatest practical importance: those of the oblique incidence of a homogeneous plane wave on an array coated with a uniform layer of ice.

We shall term the first case case E. Here:

1) the plane of incidence is normal to the wires of the array;
2) the direction of propagation of the wave makes the angle β with the normal to the plane of the array;
3) the vector $\overline{E}$ is normal to the plane of incidence and parallel to the wires of the array.

We shall term the second case case H. Here:

1) the plane of incidence is normal to the plane of the array and parallel to the wires of the array;
2) the direction of propagation of the wave makes the angle β with the normal to the plane of the array;
3) the vector $\overline{E}$ is normal to the plane of incidence.

We shall hereafter denote the quantities relating to the cases E and H by the respective superscripts E and H.

a) Case E. The intensity of the total secondary field may be expressed in the system of coordinates of the zero-th wire as follows:

$$E^E_{sm} = E_0 \sum_{m=-\infty}^{\infty} Q^E_m H^{(2)}_m(\alpha r) e^{i m (\varphi+\beta)} + $$
$$+ 2E_0 \sum_{\kappa=1}^{\infty} \sum_{m=-\infty}^{\infty} Q^E_m \sum_{n=-\infty}^{\infty} H^{(2)}_n(\alpha \kappa d) J_{m+n}(\alpha r) \cos\left(\alpha \kappa d \sin\beta + n\frac{\pi}{2}\right) \times$$
$$\times e^{i(m+n)(\varphi+\beta)}. \qquad (27)$$

If the law of variation of field intensity in the ice layer is written in the form

$$E^E_2 = E_0 \sum_{m=-\infty}^{\infty} \left[R^E_m J_m(\alpha' r) + i S^E_m N_m(\alpha' r)\right] e^{i m (\varphi+\beta)}, \qquad (28)$$

then, by performing the entire derivation as in the case of normal incidence, the following approximate expression for the coefficient Q^E_m may be obtained:

$$Q^E_m = -i^m \frac{J'_m(\alpha R) U_m - J_m(\alpha R) V_m}{K^E_m U_m - L^E_m V_m}. \qquad (29)$$

The expression so obtained is analogous in form to eq.(22). The difference consists in the values of the coefficients K^E_m and L^E_m.

For $m \neq 0$,

$$\left.\begin{aligned} K^E_m &= H^{(2)'}_m(\alpha R) + 2J'_m(\alpha R) \sum_{\kappa=1}^{\infty} \left[H^{(2)}_0(\alpha \kappa d) + H^{(2)}_{2m}(\alpha \kappa d)\right] \cos(\alpha \kappa d \sin\beta) \\ L^E_m &= H^{(2)}_m(\alpha R) + 2J_m(\alpha R) \sum_{\kappa=1}^{\infty} \left[H^{(2)}_0(\alpha \kappa d) + H^{(2)}_{2m}(\alpha \kappa d)\right] \cos(\alpha \kappa d \sin\beta) \end{aligned}\right\}. \qquad (30)$$

For $m = 0$,

$$\left.\begin{aligned} K^E_0 &= H^{(2)'}_0(\alpha R) + 2J'_0(\alpha R) \sum_{\kappa=1}^{\infty} H^{(2)}_0(\alpha \kappa d) \cos(\alpha \kappa d \sin\beta) \\ L^E_0 &= H^{(2)}_0(\alpha R) + 2J_0(\alpha R) \sum_{\kappa=1}^{\infty} H^{(2)}_0(\alpha \kappa d) \cos(\alpha \kappa d \sin\beta) \end{aligned}\right\}. \qquad (31)$$

In the case under consideration, that of oblique incidence of the wave on the array, the expression for the transmission factor will be of the form:

$$\delta_E = \left|1 + \frac{\lambda}{\pi d \cos\beta} \sum_{m=-\infty}^{\infty} (-i)^m Q^E_m\right|^2.$$

As in the case of normal incidence, the terms of the sum with indices different from zero may be neglected in calculating the transmission factor δ_E without introducing great error, i.e.:

$$\delta_E = \left|1 + \frac{\lambda}{\pi d \cos\beta} Q^E_0\right|^2. \qquad (32)$$

The principal difficulty in calculating the transmission factor is in summating the series

$$\sum_{\kappa=1}^{\infty} H^{(2)}_0(\alpha \kappa d) \cos(\alpha \kappa d \sin\beta).$$

With an accuracy unquestionably sufficient, in our case, for calculating this sum, the following approximate formula (Bibl.5) may be used:

$$2 \sum_{\kappa=1}^{\infty} H^{(2)}_0(\alpha \kappa d) \cos(\alpha \kappa d \sin\beta) \approx$$
$$\approx \sqrt{\frac{2}{\pi \alpha d}}\, e^{i\frac{\pi}{4}} \left\{L\left(-4\frac{\pi d}{\lambda}\cos^2\frac{\xi}{2}\right) + L\left(-4\frac{\pi d}{\lambda}\sin^2\frac{\xi}{2}\right)\right\}, \qquad (33)$$

where $\xi = \frac{\pi}{2} - \beta$.

The function L(x) has a period of 2π. The value of this function may be determined by the aid of the graph given in Fig.4. We note that $L(-x) = L^*(x)$, where $L^*(x)$ is the complex number conjugate to L(x).

Figure 5 gives a graph of the relation between the transmission factor δ_E and the radius of the ice layer. The calculation has been performed for the case when $\beta = 78.5^\circ$; $\frac{2\rho}{\lambda} = 0.0438$; and $\frac{d}{\lambda} = 1.25$. As will be seen from the figure, there is a

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sharp rise, in this case as well, of the transmission factor with increasing thickness of the ice coating on the wires of the array.

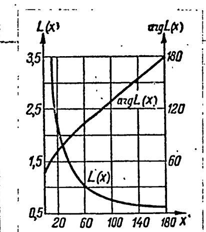

Fig.4

b) Case H. It can be shown that the expression for the coefficient Q_m^H is of the following form:

$$Q_m^H = -i^m \frac{J'_m(\alpha R\cos\psi)\, U_m^H - J_m(\alpha R\cos\psi)\, V_m^H}{K_m^H U_m^H - L_m^H V_m^H}, \qquad (34)$$

where the coefficients U_m^H, V_m^H, K_m^H and L_m^H differ from the corresponding coefficients in normal incidence by the fact that $\alpha \cos\psi$ must be substituted for α in them.

The transmission factor may be determined by the formula

$$\delta_H = \left|1 + \frac{\lambda}{\pi d\cos\psi} Q_0^H\right|^2. \qquad (35)$$

Thus the case under consideration, of the incidence of a wave at the angle ψ on an ice-coated array, is in all respects equivalent to the case of normal incidence, on the same array, of a wave with the phase constant $\alpha_1 = \alpha\cos\psi$. In other words, for the case when a wave of wavelength λ is incident on an array at the angle ψ, all the conclusions and results of the calculation for the case of normal incidence, on the same array, of a wave with the wavelength $\frac{\lambda}{\cos\psi}$, still hold.

4. Experimental Check of the Results Obtained

The energy transmission factor through an array of wires coated with a uniform layer of ice was measured. The necessary uniform ice coating was obtained by an artificial method. For this purpose I made use of the fact that if water is poured from above in a thin jet at a temperature of 10-15°C, on a vertical wire, the water in freezing will form an almost uniform layer of ice on the wire. This same method was also used for the subsequent growth of the ice layer.

To measure the transmission factor I used a set-up consisting of a transmitting and a receiving antenna, between which the array under study was placed.

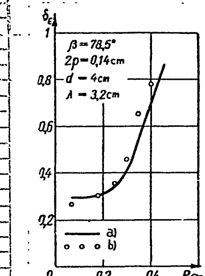

Fig.5

a) Calculated; b) Experimental

Figures 2, 3, and 5 give the results of the measurements of the transmission factor for normal and oblique (case E) incidence of the wave on an array of wires coated with a uniform layer of ice. As will be seen from the figures, over a relatively great range of variation of the thickness of the ice layer (up to about 0.1 λ), the results of the calculation are found to be in rather good agreement with the experimental data. The less satisfactory agreement on further thickening of the ice layer may be explained by the increasing relative importance, in this range of variation of the ice thickness, of the higher harmonics, which we neglected in calculating the transmission factor, and by the greater effect of the inaccuracies due to the assumptions and approximations made to simplify the analysis.

5. Conclusions

On the basis of the above, the following conclusions may be drawn.

The presence of precipitation on the wires of an array exerts a substantial influence on its electrical properties. In this case the influence of precipitation reduces down, on the whole, to an increase in the portion of the energy passing through the array, and, consequently, to the worsening of the reflecting properties of the array. In the most unfavorable cases, even strongly reflecting arrays may lose their reflecting properties almost completely.

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BIBLIOGRAPHY

1. Potekhin, A.N. - Some Problems in the Diffraction of Electromagnetic Waves. Publishing House "Sovetskoye Radio", Moscow 1948

2. Wessel, - - Hochfrequenz und Elektroakkustik (HF and Electro Acc.) 9, v.54, 1939

3. Kozhevnikov, V.A. - Dissertation for Degree of Candidate. MAI. 1952

4. Vatson [Watson], G.N. - Theory of Bessel Functions. Part 1, Publishing House for Foreign Literature. Moscow 1949

5. Yampol'skiy, V.G. - This Journal. No.9 (1955)

RECEIVING ANTENNAS AND INDUSTRIAL RADIO INTERFERENCE

by

V.V.Roditi and M.S.Gartsenshteyn

This paper discusses the question of the effective heights of room antennas and the noise transfer factors as the basic parameters determining broadcast reception quality under urban conditions, in the presence of industrial radio interference. Data on the measurement of these quantities in several USSR cities are given, and the results obtained are analyzed by the methods of mathematical statistics.

1. Introduction

The conditions of broadcast reception in cities are determined primarily by the field intensities produced by the broadcasting stations and by the level of industrial noise. The intensity of the field of the radio stations is usually measured at frequency control points far from the cities, where there is no distorting influence from buildings, overhead wires, metal structures, etc. Since room antennas are widely used, the field intensities of radio stations must be determined directly at the places where the receiving equipment is installed.

Measurements in the middle wavelength range made by various investigators have shown that the brick walls of dwelling houses, which possess good electrical conductivity, absorb the energy of the electromagnetic field, and that the electric component of that field undergoes considerable attenuation, while the magnetic field merely changes its direction, but suffers almost no decrease. In multi-story buildings the field intensity of the station usually decreases as the ground is approached. For instance, according to German data, if the field intensity on the roof of a three-story building is taken as 100%, it will be 70 - 80% in the attic, 50% on the third floor, 20% on the second floor, 5 - 10% on the first floor, and only 3 – 5% in the cellar.

In practice, however, deviations from such a distribution of field intensity by floors are often observed. This is explained, in most cases, by the influence of secondary radiators. The steel skeletons of the buildings, drain pipes and water pipes, iron staircases, cables, and other objects may act as such secondary radiators, which may attenuate the field, or, in some cases, may intensify it, depending on the natural frequency of these radiators. Owing to the screening effect of the iron roof, the field strength of the station may prove to be lower on the attic than on the floor below it.

The task of the present paper is to determine the station field strength for reception on small room antennas, as well as the noise voltage that can be induced in them by coupling with interfering networks. The measurements were made on a gamma-type antenna of copper wire 1.5 mm in diameter, with a horizontal part 3 m long and a vertical part 2 m long, suspended as far as possible from the lighting wires. A steam heat pipe or water pipe was used as the ground.

If the antenna is located in free space, its effective height may be calculated by the formula

$$h_e = \frac{2\sin m\left(l_{\alpha}-\frac{h_1}{2}\right)\sin\frac{m h_1}{2}}{m \sin m l_{\alpha}}, \qquad (1)$$

where $m = \frac{2\pi}{\lambda}$; h_1 = length of vertical part of antenna; l_{oe} = equivalent antenna length, determined as the sum of the lengths of the descent and the reduced length of the horizontal part.

Calculations for the frequency band 0.16 – 1.5 megacycles showed that the effective height of an antenna of the usually adopted dimensions should range from 1.45 to 1.65 m. It is difficult to calculate the possible attenuation of the field owing to the walls of the house, the surrounding structures, and other objects, since local conditions vary greatly.

2. Measurement of Actual Effective Heights of Receiving Antennas

Since the surrounding environment affects the value of the effective antenna height, it will be advisable to introduce the concept of the actual effective height of an antenna, i.e., allowing for the attenuation of the field due to surrounding objects. We shall denote it by the letter h.

Let us consider the method of measuring h. At first we measured the electric component of the field intensity of some powerful radio station at an open place, remote from structures and metal masses, at a height of not less than 2.5 m. Then, by means of an interference meter of type IP-12 M, we measured the EMF, e, in microwatts induced in the room antenna by the radio station in question. The meter was connected to the antenna through its equivalent antenna (a 10 μμf capacitor) and to the ground wire. The effective height was determined as the ratio of this EMF e to the field intensity E in free space:

$$h = \frac{e}{E}\, m. \qquad (2)$$

This procedure was used in 1954-1955 to measure the effective heights of receiving room antennas at Moscow, and also at Kalinin, Orel, Chkalov, Molotov, and Ivanovo. The measurements were made only in the longwave and medium-wave bands, since the fading on shortwave makes the study difficult. In all, over 320 measurements of effective antenna height were made in multi-story and low dwelling houses of various types (brick, frame and cinder-block).

The results of the measurements were worked up separately by the methods of mathematical statistics for the longwave and medium-wave bands. The values of the effective antenna heights were expressed in decibels with respect to the value at 1m:

$$H = 20\lg\frac{e}{E}\ \text{db}, \qquad (3)$$

and the class-interval of the distributions was taken at 4 or 6 db. It was found that the distribution of the values of the effective antenna heights obeys the nor-

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mal law with a rather good degree of approximation.

From the histograms of the distributions so obtained, by the pattern method, we next selected the values of the statistical parameters of the normal curves ($\overline{H}$ = mean probable values and σ_n = mean square deviations) for which the divergence from the experimental data was found to be slight. A check of the approach of the experimental distribution curves to the normal curves so selected, made by the Kolmogorov-Smirnov criterion of goodness of fit, showed the discrepancies to be insubstantial. This check was made in the following way. For the boundaries of the class-intervals of the distribution, the differences between the experimental and theoretical values of the distribution function, $N_i - N_i'$, were found, the theoretical values being determined by the aid of tables of the probability integral for assigned values: of the mean probable value (or mathematical expectation) $\overline{H}$ and of the mean square deviation σ_n. For this purpose, the values of the normed deviation t were first computed from the relation

$$t = \frac{H - \overline{H}}{\sigma_n}, \qquad (4)$$

after which the values of the probability integral were found from the tables (Bibl.1) for the ends of the class-intervals

$$\Phi(t) = \frac{1}{\sqrt{2\pi}} \int_0^t e^{-\frac{t^2}{2}} dt. \qquad (5)$$

Since it was necessary to have the values of these integrals over the range from $-\infty$ to t, then

$$F(t) = 0{,}5 + \Phi(t). \qquad (6)$$

For negative values of t, Φ (t) must be taken with a minus sign. By multiplying the values of F(t) so obtained by the total number of measurements n, we get the values of the integral distribution function N_i', i.e., the number of measurements

with values not exceeding the values of the i-th class-interval:

$$N_i' = F(t)\, n. \qquad (7)$$

For each computation point we find the modulus of the difference $|N_i - N_i'|$. Of the values so obtained we select the greatest. For instance, in calculating from the data of longwave measurements in the city of Kalinin, we obtained:

$$|N_i - N_i'|_{max} = 1{,}52.$$

Further, we determine the argument λ_0

$$\lambda_0 = \frac{|N_i - N_i'|_{max}}{\sqrt{n}}. \qquad (8)$$

In the case under consideration,

$$\lambda_0 = \frac{1{,}25}{\sqrt{25}} = 0{,}3.$$

For a given argument we find the value of the function:

$$K(\lambda_0) = \sum_{\kappa=-\infty}^{\infty} (-1)^{\kappa} e^{-2\kappa^2 \lambda_0^2}. \qquad (9)$$

This function is tabulated and given in several manuals on probability theory (Bibl.1). In our case, K = 0.3814. If the difference

$$P(\lambda_0) = 1 - K(\lambda_0) \qquad (10)$$

is great (in practice, over 0.05), then the discrepancy between the empirical and normal curves is not substantial. In the example we are calculating:

$$P(\lambda_0) = 1 - 0{,}535 = 0{,}465 \gg 0{,}05$$

and the discrepancy is not substantial.

Table 1 gives the values of the statistical parameters of the distribution of the effective heights of receiving antennas for individual cities and for all the measurements as a whole. In addition, Fig.1 shows the cumulative distribution curve

Table 1

a)	0,16÷0,5			0,5÷1,5		
b)	$\overline{H}$, db	$\overline{h}$, m	σ_n, db	$\overline{H}$, db	$\overline{h}$, m	σ_n, db
c)	—13,7	0,21	9,0	—15,4	0,17	4,6
d)	—9,1	0,35	7,6	—15,0	0,18	9,4
e)	—17,0	0,14	9,7	—15,3	0,18	6,2
f)	—11,5	0,27	9,5	—	—	—
g)	—25,3	0,054	7,7	—18,6	0,12	—
h)	—21,9	0,08	8,3	—20,5	0,095	8,8
i)	—15,7	0,164	11,0	—16,4	0,151	8,4

a) Frequency Band, megacycles; b) Parameters; c) Moscow; d) Kalinin; e) Orel; f) Chkalov; g) Ivanovo; h) Molotov; i) All these cities

of the effective heights of receiving antennas for all the measurements in all six cities. The normal curve, plotted by the dashed line, is very close to the curve

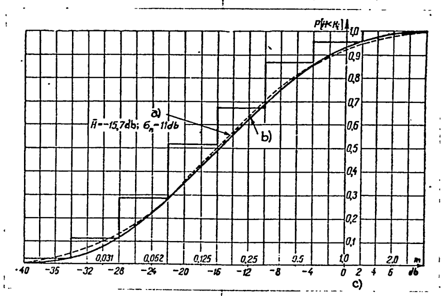

Fig.1

a) Normal curve; b) Experimental curve; c) Effective height H

constructed from the experimental data. The frequency dependence of the mean probable value of the effective heights is shown in Fig.2.

It is interesting to compare these data with those published earlier. In view of the absence of such information from the USSR literature, only foreign data

Table 2

a)	0,16÷0,20	0,7÷0,9	1,1÷1,2
b)	—8,2	—1,9	—3,4
c)	8,5	8,9	8,4

a) Frequency band, megacycles; b) Mean probable values of effective angenna heights $\overline{H}$, db; c) Mean square deviations σ_n, db

dating rather far back is available for comparison (Bibl.2,3). A statistical work-up of data given by SMKPR* in 1935, based on the results of over 500 measurements of the effective heights of receiving antennas in various countries gave the results presented in Table 2.

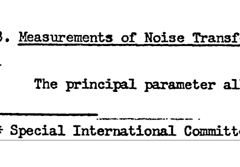

Fig.2

Thus the mean probable value of the effective heights of receiving antennas on long wave was found to be about 7 db less, and on medium wave, about 13 db less, then according to the SMKPR data. The discrepancy is probably to be explained by the differences in house construction and room size, and in the smaller antennas that were used.

3. Measurements of Noise Transfer Ratios

The principal parameter allowing determination of the noise voltage at the re-

* Special International Committee for Control of Broadcast Interference.

ceiver input is the noise transfer ratio K_p, i.e., the ratio of the noise voltage U across the terminals of the source of noise connected to the network, and the noise voltage U_a across the antenna-ground terminals of the receiver:

$$K_p = 20 \lg \frac{U}{U_a} \, db. \quad (11)$$

The value of the noise transfer ratio depends on the distance of the source of noise from the receiver, on the design of the wiring, the dimensions and mutual position of the room antenna and the noise-carrying networks. The noise transfer ratio is a random quantity, so that statistical measurements are required to determine its law of distribution and its principal parameters, the mean probable value $\overline{K}_p$ and the mean square deviation σ_k. It is assumed that the use of power-supply filters in the feed circuits prevents the direct penetration of noise into the receivers.

The noise transfer ratio in a building is measured as follows. A high-frequency voltage of the order of a few volts is impressed on the electric lighting system or other noise-carrying network from the main switch or distribution switch-

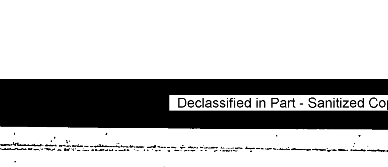

Fig.3

a) HF band generator (50 kc. to 40 mc.); b) Power-supply unit; c) Modulator, 1000 cycles; d) Cathode voltmeter (0.1 to 150 volts)

board, and the voltage at this same frequency is measured between the antenna and ground terminals in several rooms on different stories, using an interference meter, for instance one of type IP-12M. The high-frequency voltage is supplied from the output of the noise transfer ratio meter, type IKP-3M, developed by the TsLIRMEP,

containing a high-frequency generator with asymmetric output and a vacuum-tube voltmeter at the output.

Figure 3 is a schematic diagram of the type IKP-3M instrument. The filter choke at the input of the cathode voltmeter is necessary to prevent the instrument from measuring the voltage of the industrial 50-cycle current, so that it will measure only the high-frequency voltage introduced into the network. The IKP-3M is usually placed on a fishbone antenna ladder-type cell near the main switchboard. The high-frequency voltage is successively imposed on all three wires of the lighting system, and also on the two wires of the wire-broadcasting system which is connected with the lighting wire and is also a good carrier of noise. The level of the high frequency voltage introduced into the system (of the order of 5 - 10 volts) is considerably higher than the level of all the outside noises in the system, so that these noises do not interfere with the measurement procedure.

The measurements were made in one or two apartments of each floor at the frequencies recommended in the standards for the maximum allowable industrial radio interference. Such measurements of the noise transfer ratios were made in several houses at Moscow and Kalinin. The total number of measurements of the noise transfer ratios, at frequencies from 0.16 to 150 megacycles, was over 1500.

The results of the measurements were worked up statistically for the individual houses in various frequency ranges. The values of the noise transfer ratios in decibels were distributed into class-intervals 5 db wide, and the distribution curves were constructed. Normal curves were selected for the distribution curves so obtained, and a check of the possibility of representing the experimental curves by normal ones, performed by the Kolmogorov-Smirnov criterion of goodness of fit, analogous to the check made for the effective antenna heights, showed that the discrepancies were not substantial.

Table 3 gives the values of the statistical parameters of distribution for separate houses at Moscow and Kalinin.

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They can be regarded only as a first approximation to the elucidation of the problem, since the number of houses in which measurements were made was small. On

Table 3

a)	c)	b)			
		d)		e)	
		f)	g)	h)	i)
0,16÷0,4	$\overline{K}_p$	43,2	42,7	24,2	29,3
	σ_K	6,4	9,4	8,1	5,5
0,5÷1,5	$\overline{K}_p$	45,8	43,4	38,6	43,9
	σ_K	8,5	10,2	12,5	16,6
3÷20	$\overline{K}_p$	51,3	68,0	51,6	63,0
	σ_K	9,6	16,8	21,9	11,3

a) Frequency band, megacycles; b) Mean Probable Values of Interference Transfer Ratios $\overline{K}$ and of Mean Square Deviations σ_k, db; c) Parameters of Distribution; d) Moscow; e) Kalinin; f) Four-Story House; g) Six-Story House; h) Three-Story House; i) Four-Story House

comparing the results of measurements in different houses, it will be seen that in the medium wave and short wave range the mean probable values of the interference transfer ratio measured at Moscow and Kalinin differ only slightly from each other.

Table 4

a)	b)		c)	
	$\overline{K}$ db	σ_k db	$\overline{K}$ db	σ_k db
50	68,6	19,4	86,7	10,2
80	75,0	14,1	84,2	5,0
150	67,4	20,4	88,2	9,2

a) Frequency, megacycles; b) Four-Story House, $\overline{K}$,db, σ_k,db; c) Six-Story House, $\overline{K}$,db, σ_k,db

In the ultrashort wave range, the noise transfer ratios are considerably higher than on the lower frequencies. The statistical parameters of the distributions of

the noise transfer ratios from the results of measurements in two houses are given in Table 4 for frequencies 50, 80, and 150 megacycles. A tuned horizontal half-wave dipole was used as the receiving antenna. The lack of detailed data in the USSR and foreign literature makes it impossible to compare our data with those of others.

The data on the statistical parameters of distribution of the effective heights of the receiving antennas of radio kisteners and of the noise transfer ratios will allow the solution of important questions connected with the conditions of broadcast reception: (1) determination of the probability of the disturbing effect of interference in radio receivers at known levels of the noise voltage across the terminals of equipment connected to the line voltage; (2) determination of the probability of the disturbing effect of interference if the current standards of maximum allowable industrial radio noise are observed, according to the field intensity of the radio station in free space; (3) determination of the necessary standards of noise voltage across the terminals of noise sources in reception from a radio station in order for an assigned signal-noise ratio to obtain for an assigned value of the field intensity.

A consideration of these questions would be beyond the scope of the present paper.

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BIBLIOGRAPHY.

1. Gnedenko,B.V. - Course in the Theory of Probability. 2nd Ed. GITTL. 1954

2. - Session of Special International Committee on Radio Broadcast Interference. Journal des Telecommunications (5), pp.123-125 (1935)

3. Seeman,F. - Suppression of Radio Interference, Darmstadt-Berlin. (1954), pp.307-308

MUTUAL CORRELATION OF FLUCTUATION NOISES AT THE OUTPUT OF FREQUENCY DISCRIMINATORS

by

M.V.Maksimov

The function of mutual correlation $K(\tau)$ is determined for fluctuation noises at the output of line filters constituting the load of radio-frequency equipment with and without a detector. On the basis of general considerations, the function $K(\tau)$ is found with respect to resonant oscillatory circuits. The parameters of the system under consideration are analyzed.

1. Introduction

In a number of radio engineering devices the signals, after passing the radio-frequency stage Φ_o (Fig.1) and the detector D of the receiver, are fed to narrow-band frequency discriminators Φ_1 and Φ_2, designed to separate the channels. The oscillations received at the outputs of these filters, after certain transformations in the devices Pr_1 and Pr_2, enter the comparator, and then the final apparatus.

One of the numerous examples of radio receiving installations of this type is the system in which the numerical values of the transmitted signal are characterized by the phase of the sinusoidal carrier voltage. In such a system, the filters Φ_1 and Φ_2 separate, respectively, the sinusoidal oscillations of the alternating phase (mismatch voltage) and the direct current phase (reference voltage). The converters Pr_1 and Pr_2 produce signals of the same frequency, which act on the comparator, or phase detector.

It is also possible to construct systems in which the filters Φ_1 and Φ_2 are located immediately after the filter Φ_o and the signals are detected separately in each channel. We shall henceforth term an installation of the latter type a system

with detector.

We note that the number of narrow-band filters and comparators may be considerably greater than 2.

In considering the question of the action of noise on a comparator, the fact that the two output circuits have a common source of noise, the radio receiver, must be taken into account.

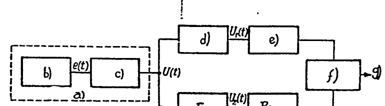

Fig.1

a) Radio receiver; b) F_o; c) D; d) F_1; e) Pr_1; f) Comparator; g) Output

In this connection it is important to know the function of mutual correlation $K(\tau)$ between the voltages $U_1(t)$ and $U_2(t)$ (Fig.1), and also the parameters of the system at which $K(\tau)$ becomes insignificant and can be neglected. On the bases of the known function $K(\tau)$ not only can we solve the question of the action of fluctuation noises on the comparator, but also, as a consequence, the problem of the interference of complex signals, which has been considered by another author (Bibl.2,3). In these works essentially what was done was to find the dispersion of the product of $U_1(t)$ and $U_2(t)$ under the condition of the absence of a detector ahead of the filters Φ_1 and Φ_2 when white noise acts on these quadripoles.

The present paper is devoted to the determination of the function of mutual correlation $K(\tau)$ of the voltages $U_1(t)$ and $U_2(t)$ obtained at the outputs F_1 and F_2 (Fig.1), which the noise reaches from the resonant system of the radio receiver through the detector D. The solution of this problem for a system without a detector is also briefly discussed, and examples are given of the determination of $K(\tau)$ when single resonant oscillatory circuits, which are most often used in actual radio engineering, are used as the filters Φ_1 and Φ_2.

2. The Function of Mutual Correlation of the Noises in a System with a Detector

Assume that the receiver uses the linear detector D and employs radiofrequency stages having the resonant system Φ_o with a rather narrow and symmetrical frequency characteristic. Assume further that only fluctuation noise is present, while a useful signal is absent. The following expression may then be written for e(t) (Fig.1):

$$e(t) = E(t) \cos [\omega_0 t + \theta(t)], \qquad (1)$$

where E(t) is the noise envelope, varying stochastically with time, $\alpha(t)$ is the phase, varying stochastically with time, and ω_o the resonant frequency of the filter Φ_o, representing the radiofrequency stages of the radio receiver.

The voltage of the low-frequency components of the noise at the output of Φ_1 and Φ_2, provided the linear detector D is used, and that the constant component of the noises $\overline{E}(t)$ does not pass, is equal to

$$U(t) = cE(t) - c\overline{E(t)}, \qquad (2)$$

where c = factor of proportionality.

The values of the steady voltages at the output of Φ_1 and Φ_2, in accordance with the Duhamel integral, are determined as

$$U_1(t_1) = \int_0^\infty U(t_1 - x)\, h_1(x)\, dx = c\int_0^\infty [E(t_1 - x) - \overline{E(t)}]\, h_1(x)\, dx, \qquad (3)$$

$$U_2(t_2) = \int_0^\infty U(t_2 - y)\, h_2(y)\, dy = c\int_0^\infty [E(t_2 - y) - \overline{E(t)}]\, h_2(y)\, dy. \qquad (4)$$

In these expressions $h_1(x)$ and $h_1(y)$ = reaction of the filters Φ_1 and Φ_2 to a unit pulse.

On the basis of eqs.(3) and (4), we find the function of mutual correlation $K(t_2 - t_1)$ of the noises $U_1(t)$ and $U_1(t)$:

$$K(t_2 - t_1) = \overline{U_1(t_1)\, U_2(t_2)} =$$
$$= c^2 \int_0^\infty \int_0^\infty \overline{[E(t_1 - x) - \overline{E(t)}]\, [E(t_2 - y) - \overline{E(t)}]}\, h_1(x)\, h_2(y)\, dx dy.$$

Here $\overline{U_1(t_1)U_2(t_2)}$ = mean statistical product of $U_1(t_1)$ and $U_2(t_2)$.

By expanding the product of the first two cofactors in the integrand expression, and denoting

$$\overline{E(t)} = \overline{E},$$

we get

$$K(t_2 - t_1) = c^2 \int_0^\infty \int_0^\infty [\overline{E(t_1 - x)\, E(t_2 - y)} - \overline{E}^2]\, h_1(x)\, h_2(y)\, dx dy. \qquad (5)$$

The quantity

$$K_0(t_2 - t_1 + x - y) = K_0(\tau + x - y) = \overline{E(t_1 - x)\, E(t_2 - y)}$$

represents the function of autocorrelation of the envelope of voltage across the output of Φ_o.

From eq.(5) we find

$$K(\tau) = c^2 \int_0^\infty \int_0^\infty [K_0(x + \tau - y) - \overline{E}^2]\, h_1(x)\, h_2(y)\, dx dy. \qquad (6)$$

In the calculations it is more convenient to use the expression for the coefficient of mutual correlation:

$$R(\tau) = \frac{K(\tau)}{\sigma_1 \sigma_2}, \qquad (7)$$

where σ_1^2 and σ_2^2 = mean values of the squares of the voltages $U_1(t)$ and $U_2(t)$. These quantities are easily determined if we know the autocorrelation functions for $U_1(t)$ and $U_2(t)$. Using the above described method, we get:

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$$\sigma_1^2 = c^2 \int_0^\infty \int_0^\infty [K_0(x - z) - \overline{E}^2]\, h_1(x)\, h_1(z)\, dx dz, \qquad (8)$$

$$\sigma_2^2 = c^2 \int_0^\infty \int_0^\infty [K_0(x - z) - \overline{E}^2]\, h_2(x)\, h_2(z)\, dx dz. \qquad (9)$$

On the basis of eqs.(6), (7), (8) and (9), we find that

$$= \frac{\int_0^\infty \int_0^\infty [K_0(x + \tau - y) - \overline{E}^2]\, h_1(x)\, h_2(y)\, dx dy}{\left\{\int_0^\infty \int_0^\infty [K_0(x - z) - \overline{E}^2]\, h_1(x)\, h_1(z)\, dx dz \int_0^\infty \int_0^\infty [K_0(x - z) - \overline{E}^2]\, h_2(x)\, h_2(z)\, dx dz\right\}^{\frac{1}{2}}}. \qquad (10)$$

It is commonly known (Bibl.1) that

$$\overline{E} = \sigma \sqrt{\frac{\pi}{2}}, \qquad (11)$$

$$K_0(x + \tau - y) = \frac{\pi\sigma^2}{2}\left[1 + \left(\frac{1}{2}\right)^2 p^2(x + \tau - y) + \left(\frac{1}{2 \cdot 4}\right)^2 p^4(x + \tau - y) + \ldots\right], \qquad (12)$$

where

$$p^2(x + \tau - y) = r^2(x + \tau - y) + s^2(x + \tau - y).$$

The quantities $r(x + \tau - y)$ and $s(x + \tau - y)$ are related in the following way to the coefficient of correlation of the noises, $R_o(x + \tau - y)$ at the output of Φ_o:

$$R_0(x + \tau - y) = r(x + \tau - y)\cos\omega_0(\tau + \tau - y) + s(x + \tau - y)\sin\omega_0(x + \tau - y).$$

With a sufficiently narrow and symmetrical frequency characteristic of the resonant system of Φ_o of the receiver,

$$s(x + \tau - y) \approx 0. \qquad (13)$$

The series (12) converges very rapidly. For this reason, as shown in Bibl.1, in calculating $K_o(x + \tau - y)$, it is sufficient to confine oneself to the first two terms. Taking this value into account, and also eqs.(11), (12) and (13), we get

$$R(\tau) = \frac{\int_0^\infty \int_0^\infty r^2(x + \tau - y)\, h_1(x)\, h_2(y)\, dx dy}{\left[\int_0^\infty \int_0^\infty r^2(x - z)\, h_1(x)\, h_1(z)\, dx dz \int_0^\infty \int_0^\infty r^2(x - z)\, h_2(x)\, h_2(z)\, dx dz\right]^{\frac{1}{2}}}. \qquad (14)$$

This last expression allows $R(\tau)$ to be calculated as a function of the parameters of the filters Φ_o, Φ_1, and Φ_2.

3. The Mutual Correlation Function of the Noises in a System without Detector

If the filters Φ_1 and Φ_2 are placed immediately after the filter Φ_o, then the problem under consideration may be solved with respect to the instantaneous values of the noise voltages arriving from Φ_o.

Assuming that the instantaneous voltage $e(t)$ acts in the system without a detector, and making use of the Duhamel integral to determine the voltages $U_1(t_1)$ and $U_2(t_2)$ at the outputs of Φ_1 and Φ_2, we get the following expression for the mutual correlation function of the noises $K_1(\tau)$:

$$K_1(\tau) = \overline{U_1(t_1)\, U_2(t_2)} = \int_0^\infty \int_0^\infty \overline{e(t_1 - x)\, e(t_2 - y)}\; h_1(x)\, h_2(y)\, dx dy. \qquad (15)$$

But, bearing in mind that the mean value of the noise at the output of Φ_o equals zero, while the effective value equals σ, we find that $\overline{e(t_1 - x)\, e(t_2 - y)} = \sigma^2 R(x + \tau - y)$, and that, therefore,

$$K_1(\tau) = \int_0^\infty \int_0^\infty \sigma^2 R(x + \tau - y)\, h_1(x)\, h_2(y)\, dx dy.$$

In these last formulas $R(x + \tau - y) = r(x + \tau - y)\cos\omega_o(x + \tau - y)$ is the coefficient of correlation of the noises at the output of Φ_o. Similarly, we obtain for the quantities σ_1 and σ_2:

$$\sigma_1^2 = \sigma^2 \int_0^\infty \int_0^\infty R(x - z)\, h_1(x)\, h_1(z)\, dx dz, \qquad (16)$$

$$\sigma_2^2 = \sigma^2 \int_0^\infty \int_0^\infty R(x - z)\, h_2(x)\, h_2(z)\, dx dz. \qquad (17)$$

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On the basis of eqs.(15), (16) and (17), we find the following expression for the coefficient of mutual correlation of the instantaneous values of the noises in the system without detector:

$$R_1(\tau)=\frac{\int_0^\infty\int_0^\infty R(x+\tau-y)\,h_1(x)\,h_2(y)\,dxdy}{\left[\int_0^\infty\int_0^\infty R(x-z)\,h_1(x)\,h_1(z)\,dxdz\int_0^\infty\int_0^\infty R(x-z)\,h_2(x)\,h_2(z)\,dxdz\right]^{\frac{1}{2}}}. \qquad (18)$$

Equation (18) enables us to find the function of mutual correlation of the noises at the outputs of Φ_1 and Φ_2 for known parameters of these devices, and also of the filter Φ_0.

4. Determination of the Mutual Correlation of the Noises when Resonant Circuits are Used as Filters

If resonant circuits are used as the filters Φ_1 and Φ_2, then the corresponding coefficients of mutual correlation may be calculated from the generally known expressions for $R(x+\tau-y)$, $h_1(x)$ and $h_2(y)$. It must be noted, in this case, that resonant circuits are rather frequently used in practice as separation filters. The approximation of the high-frequency stages of a radio receiver to a single oscillatory circuit is permissible, since the function of mutual correlation of the noises at the output, $K(\tau)$, with which we are concerned, has roughly the same character for frequency filters of various types (Bibl.1).

a) Determination of the Coefficient of Mutual Correlation in a System with Detector

It is commonly known (Bibl.1) that for single resonant oscillatory circuits:

$$R(x+\tau-y)\approx r(x+\tau-y)\cos\omega_0(x+\tau-y)=$$
$$=e^{-\alpha|x+\tau-y|}\cos\omega_0(x+\tau-y),$$
$$h_1(x)=2\beta\,e^{-\beta x}\cos\omega_1 x,$$
$$h_2(y)=2\gamma\,e^{-\gamma y}\cos\omega_2 y,$$

where $\alpha=\frac{\omega_0\delta_0}{2}$; $\beta=\frac{\omega_1\delta_1}{2}$; $\gamma=\frac{\omega_2\delta_2}{2}$ δ_0, δ_1 and δ_2 = the damping of the circuits of Φ_0, Φ_1 and Φ_2.

Under these conditions, the numerator in eq.(14) takes the form:

$$I_1=4\beta\gamma\int_0^\infty e^{-\gamma y}\cos\omega_2 y\,dy\int_0^\infty e^{-2\alpha(x+\tau-y)}e^{-\beta x}\cos\omega_1 x\,dx. \qquad (19)$$

We shall calculate the integral of I_1 separately for $\tau<0$ and $\tau>0$. If $\tau<0$, then, expressing $\cos\omega_1 x$ in terms of exponential functions, the inner integral may be represented in the form:

$$I_2=\frac{1}{2}\int_0^\infty e^{-2\alpha|x-(\tau+y)|}\left[e^{(-\beta+i\omega_1)x}+e^{(-\beta-i\omega_1)x}\right]dx.$$

Dividing the entire domain of integration into two, we get

$$I_2=\frac{1}{2}\int_0^{\tau+y} e^{-2\alpha(\tau+y-x)}\left[e^{(-\beta+i\omega_1)x}+e^{(-\beta-i\omega_1)x}\right]dx+$$
$$+\frac{1}{2}\int_{\tau+y}^\infty e^{-2\alpha[x-(\tau+y)]}\left[e^{(-\beta+i\omega_1)x}+e^{(-\beta-i\omega_1)x}\right]dx.$$

Taking account of the expression for I_2, after performing the integration of eq.(19) and the corresponding transformations of the relation so obtained, we find

$$(I_1)_{\tau<0}=4\beta\gamma\Big\{e^{-\beta\tau}\left[A\cos\omega_1\tau+B\sin\omega_1\tau\right]-$$
$$-e^{-2\alpha\tau}\frac{(2\alpha-\beta)(2\alpha+\gamma)}{[(2\alpha-\beta)^2+\omega_1^2]\,[(2\alpha+\gamma)^2+\omega_2^2]}\Big\}, \qquad (20)$$

where

$$A=\frac{2\alpha(\beta+\gamma)[2(\beta+\gamma)^2+\Delta\omega^2+(2\omega_1+\Delta\omega)^2][4\alpha^2+\omega_1^2-\beta^2]+8\alpha\beta\omega_1^2[(\beta+\gamma)^2-2\omega_1\Delta\omega-\Delta\omega^2]}{[(\beta+\gamma)^2+(2\omega_1+\Delta\omega)^2][(\beta+\gamma)^2+\Delta\omega^2][(4\alpha^2-\beta^2)^2+2\omega_1^2(4\alpha^2+\beta^2)+\omega_1^4]}$$

$$B=\frac{4\alpha\omega_1[(\beta+\gamma)^2-2\Delta\omega\omega_1-\Delta\omega^2][4\alpha^2+\omega_1^2-\beta^2)+}{[(\beta+[illegible])^2+(2[illegible]+\Delta\omega)^2]\,[(\beta+\gamma)^2+\Delta\omega^2][(4\alpha^2-\beta^2)^2+}\rightarrow$$

$$\rightarrow \frac{+4\alpha\beta\omega_1(\beta+\gamma)\,[2\,\beta+\gamma)^2+\Delta\omega^2+(2\omega_1+\Delta\omega)^2]}{+2\omega_1^2(4\alpha^2+\beta^2)+\omega_1^4]},$$

$$\Delta\omega = \omega_2 - \omega_1,$$

$(I_1)_{\tau<0}$ = values of I_1 for $\tau < 0$.

Knowing I_1, it is easy to find $K(\tau)_{\tau<0}$ for $\tau < 0$:

$$K(\tau)_{\tau<0} = \frac{c^2\pi\sigma^2}{2}(I_1)_{\tau<0}. \tag{21}$$

For the case $\tau > 0$, the integral of eq.(19) may be found as follows: First of all, let us pass to the new constant of integration $z = x + \tau$, and let us divide the entire domain of integration with respect to y into two parts: a) $0 \leqslant y \leqslant \tau$; b) $\tau \leqslant y \leqslant \infty$.

Then we obtain

$$(I_1)_{\tau>0} = 4\beta\gamma\int_0^{\tau} e^{-\gamma y}\cos\omega_2 y\,dy\int_{\tau}^{\infty} e^{-2\alpha|z-y|}e^{-\beta(z-\tau)}\cos\omega_1(z-\tau)\,dz +$$

$$+ 4\beta\gamma\int_{\tau}^{\infty} e^{-\gamma y}\cos\omega_2 y\,dy\int_{\tau}^{\infty} e^{-2\alpha|z-y|}e^{-\beta(z-\tau)}\cos\omega_1(z-\tau)\,dz.$$

By division into separate regions of integration with respect to y, this expression may be freed from modulus signs. We get:

$$(I_1)_{\tau>0} = 4\beta\gamma\int_0^{\tau} e^{-\gamma y}\cos\omega_2 y\,dy\int_{\tau}^{\infty} e^{-2\alpha(z-y)}e^{-\beta(z-\tau)}\cos\omega_1(z-\tau)\,dz +$$

$$+ 4\beta\gamma\int_{\tau}^{\infty} e^{-\gamma y}\cos\omega_2 y\,dy\int_{\tau}^{y} e^{-2\alpha(y-z)}e^{-\beta(z-\tau)}\cos\omega_1(z-\tau)\,dz +$$

$$+ 4\beta\gamma\int_{\tau}^{\infty} e^{-\gamma y}\cos\omega_2 y\,dy\int_{y}^{\infty} e^{-2\alpha(z-y)}e^{-\beta(z-\tau)}\cos\omega_1(z-\tau)\,dz.$$

After performing integration and the corresponding transformations, we find:

$$(I_1)_{\tau>0} = 4\beta\gamma\left\{e^{-\gamma\tau}\left[C\cos\omega_2\tau - D\sin\omega_2\tau\right] -\right.$$

$$\left. - e^{-2\alpha\tau}\frac{(2\alpha+\beta)(2\alpha-\gamma)}{\left[(2\alpha+\beta)^2+\omega_1^2\right]\left[(2\alpha-\gamma)^2+\omega_2^2\right]}\right\}, \tag{22}$$

where

$$C = A + \frac{(2\alpha+\beta)(2\alpha-\gamma)}{\left[(2\alpha+\beta)^2+\omega_1^2\right]\left[(2\alpha-\gamma)^2+\omega_2^2\right]} - \frac{(2\alpha-\beta)(2\alpha+\gamma)}{\left[(2\alpha-\beta)^2+\omega_1^2\right]\left[(2\alpha+\gamma)^2+\omega_2^2\right]},$$

$$D = \frac{4\alpha\left[(\omega_1+\Delta\omega)(\beta+\gamma)^2+3\,\omega_1\Delta\omega\,(\omega_1+\Delta\omega)+\Delta\omega^3\right]\left[4\alpha^2+\right.}{\left[(\beta+\gamma)^2+(2\omega_1+\Delta\omega)^2\right]\left[(\beta+\gamma)^2+\Delta\omega^2\right]\left[(4\alpha^2-\beta^2)^2+\right.} \rightarrow$$

$$\rightarrow \frac{\left.+\omega_1^2-\beta^2\right] - 16\,\omega_1^2\alpha\beta\,(\beta+\gamma)(\omega_1+\Delta\omega)}{\left.+2\omega_1^2(4\alpha^2+\beta^2)+\omega_1^4\right]} +$$

$$+ \frac{(2\alpha+\beta)\,\omega_2}{\left[(2\alpha+\beta)^2+\omega_1^2\right]\left[(2\alpha-\gamma)^2+\omega_2^2\right]} - \frac{(2\alpha-\beta)\,\omega_2}{\left[(2\alpha-\beta)^2+\omega_1^2\right]\left[(2\alpha+\gamma)^2+\omega_2^2\right]}.$$

Correspondingly,

$$K(\tau)_{\tau>0}\,\frac{c^2\pi\sigma^2}{2}(I_1)_{\tau>0}. \tag{23}$$

We note that (as it should be) for $\tau = 0$ the values of the functions of mutual correlation eqs.(21) and (23) coincide. By analyzing the expressions for $(I_1)_{\tau>0}$ and $(I_1)_{\tau<0}$, it is easy to see that $K(\tau)$ depends both on the time $\tau = t_2 - t_1$, and on the parameters of the filters Φ_0, Φ_1 and Φ_2. The maximum value of $K(\tau)$ corresponds to $\tau = 0$. To ascertain the relationship between the coefficient of mutual correlation and the parameters of Φ_0, Φ_1 and Φ_2, let us determine $R(\tau) = R(0)$ for $\tau = 0$. For this purpose let us substitute, in eq.(20), the value $\tau = 0$, and in accordance with eqs.(8) and (9) let us find σ_1^2 and σ_2^2:

$$\sigma_1^2 = 4\,\frac{\alpha\beta\,(2\beta^2+\omega_1^2)(4\alpha^2+\omega_1^2-\beta^2)+2\alpha\beta^3\omega_1^2-\beta^2(\beta^2+\omega_1^2)(4\alpha^2-\beta^2)}{(\beta^2+\omega_1^2)\left[(4\alpha^2-\beta^2)^2+2\omega_1^2(4\alpha^2+\beta^2)+\omega_1^4\right]},$$

$$\sigma_2^2 = 4\,\frac{\alpha\gamma\left[(2\gamma^2+(\omega_1+\Delta\omega)^2\right]\left[4\alpha^2+(\omega_1+\Delta\omega)^2-\gamma^2\right]+}{\left[\gamma^2+(\omega_1+\Delta\omega)^2\right]\left[4\alpha^2-\gamma^2)^2+\right.} \rightarrow$$

$$\rightarrow \frac{+2\alpha\gamma^3(\omega_1+\Delta\omega)^2-\gamma^2\left[\gamma^2+(\omega_1+\Delta\omega)^2\right]\left[4\alpha^2-\gamma^2\right]}{+2(\omega_1+\Delta\omega)^2(4\alpha^2+\gamma^2)+(\omega_1+\Delta\omega)^4]}.$$

By making use of the values so found for σ_1^2 and σ_2^2, as well as the formula for $(I_1)_{\tau=0}$ for $\tau = 0$, we may, on the basis of eq.(14) write the expression for R(0). It is easily shown that this is obtained in a very unwieldy form, little suited for technical calculations. It must, however, be borne in mind that usually, in practice, $\delta_1 \ll 1$, $\delta_0 \ll 1$, while $m = \frac{\omega_0}{\omega_1}$, may lie in the range from several hundred to several thousand. The latter conclusion is easy to draw, in view of the fact that the frequencies ω_0, which are essentially intermediate frequencies of the radio receiver, have values from a few hundred kilocycles to a few tens of megacycles; while the resonance frequencies of the filters Φ_1 and Φ_2 lie, as a rule, in the audio-frequency range. Therefore $m\delta_0$ is always more than unity, and in most cases the condition $m\delta_0 \gg 1$ is satisfied.

Bearing all this in mind, we may obtain the following approximate expression for R(0) under the condition that the pass bands of the filters are the same:

$$R(0) = \left[\frac{\delta_1^2[\varepsilon^2 + (2+\varepsilon)^2]}{(2+\varepsilon)^2(\delta_1^2+\varepsilon^2)} - \frac{m\delta_0\delta_1}{m^2\delta_0^2 + (1+\varepsilon)^2}\right] \times$$

$$\times \frac{m^2\delta_0^2 + (1+\varepsilon)^2}{\sqrt{(m^2\delta_0^2 + 1 - m\delta_0\delta_1)[m^2\delta_0^2 + (1+\varepsilon)^2 - m\delta_0\delta_1(1+\varepsilon)]}}, \qquad (24)$$

where $\varepsilon = \Delta\omega/\omega_1$. It will be clear from this expression that with increasing ε (especially for $\varepsilon < 1$), the value of R(0) declines very rapidly. This confirms our calculations of the dependence of R(0) on k, $m\delta_0$, and δ_1. The data of the calculations are given in the graphs of Fig.2.

It follows from these graphs that the quantity R(0) is a function not only of the relative detuning $k = \frac{\Delta\omega}{\Delta\omega_n} = \frac{\varepsilon}{\delta_1}$ of the filters Φ_1 and Φ_2, but also depends on the ratio between the pass band of the filter Φ_0, which has the pass band $\Delta\omega_{n0}$, and the pass bands of the filters Φ_1 and Φ_2, which have the pass band $\Delta\omega_n$. It follows from Fig.2 that with increasing ratio $\frac{\Delta\omega_{no}}{\Delta\omega_n}$, the values of R(0) for the same values of k increase, but insignificantly.

The increase in R(0) may be explained by the fact that at one and the same values of k, with increasing value of $\frac{\Delta\omega_{no}}{\Delta\omega_n}$, noises closer together in intensity arrive in the band of Φ_1 and Φ_2.

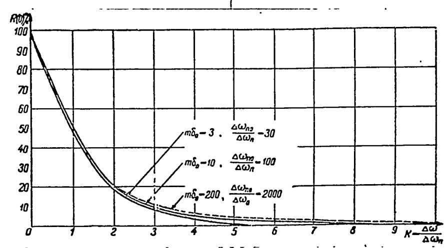

Fig.2

As a result of the analysis of the graphs given in Fig.2, the general conclusion may be drawn that in most practical cases the voltages across the outputs of separating filters may be considered uncorrelated (coefficient of correlation less than 2%) if their resonant frequencies are separated by

$$k = \frac{\Delta\omega}{\Delta\omega_n} > 5.$$

b) Determination of Coefficient of Mutual Correlation of Noises in System without Detector

By substituting in eq.(18) the values of $R(x + \tau - y)$, $h_1(x)$ and $h_2(y)$, an expression may be obtained for the coefficient of mutual correlation $R_1(\tau)$ in a system without a detector, using single circuits as the filters Φ_1 and Φ_2. This expression is very unwieldy. The maximum value of $R_1(\tau)$, as was to be expected, will occur at $\tau = 0$. The expression for $R_1(0)$ may be somewhat simplified if we take into account the fact that usually $\delta_0 \ll 1$, $\delta_1 \ll 1$, and $\delta_2 \ll 1$. Moreover, the frequencies ω_1 and ω_2 should lie in the pass band of the filter Φ_0, and therefore ω_1 and ω_2 are close to ω_0. Finally, so that several separating filters may be placed in the pass band of the filter Φ_0, it is necessary that the pass band of Φ_0 shall be considerably wider than the pass band of each of the separating filters. This condition leads to the existence of the inequality $\delta_0 \gg \delta_1$.

Taking account of all these remarks, and also assuming that the pass bands of the filters Φ_1 and Φ_2 are equal, the following approximate expression for $R_1(0)$ may be obtained:

$$R_1(0) = \left\{ \delta_1 \frac{1}{\delta_1^2 + \varepsilon^2} \cdot \frac{1}{m^2\delta_0^2 + 4(1-m)} - \frac{4m\delta_0 (1+m^2)(1+m^2+2\varepsilon+\varepsilon^2)}{(1+m)^2(1+m+\varepsilon)^2[m^2\delta_0^2+4(1-m)^2][\delta_0^2+4(1-m+\varepsilon)^2]} \right\} :$$

$$: \sqrt{\left\{ \frac{1}{\delta_1} \frac{1}{m^2\delta_0^2 + 4(1-m)^2} - \frac{4m\delta_0(1+m^2)^2}{(1+m)^4[m^2\delta_0^2+4(1-m)^2]^2} \right\} \times \left\{ \frac{1}{\delta_1} \frac{1}{m^2\delta_0^2+4(1+\varepsilon-m)^2} - \frac{4m\delta_0[(1+\varepsilon)^2+m^2]^2}{(1+m+\varepsilon)^4[m^2\delta_0^2+4(1-m+\varepsilon^2)]^2} \right\}}. \quad (25)$$

It must be noted that eq.(25), which rather accurately represents the dependence of $R_1(0)$ on the system parameters, allows us to draw the conclusion that with the growth of ε, the coefficient of mutual correlation $R_1(0)$ declines very rapidly. This conclusion follows from the fact that if the bands of the filters Φ_1 and Φ_2 lie in the pass band of Φ_0, then the values of ε are substantially less than unity. If, however, ε is small, then when it increases, the expression under the radical sign in eq.(25) remains practically constant, while the summands of the first cofactor decrease.

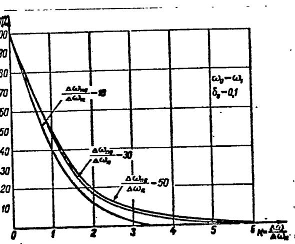

Fig.3

The results of calculations by eq.(25) for m = 1 and various values of δ_1 are represented by the graphs of Fig.3. These graphs confirm the conclusions that have just been drawn on the behavior of the function $R_1(0)$.

By comparing the graphs given in Figs.2 and 3, the conclusion may also be drawn that the values of the coefficients of correlation R(0) and $R_1(0)$ in a system with detector and in one without detector are determined primarily by the ratio of the pass band $\Delta\omega_n$ of the filter Φ_1 (or respectively Φ_2) to the pass band $\Delta\omega_{no}$ of the filter Φ_0. In this case, when the ratio $\frac{\Delta\omega_{no}}{\Delta\omega_n}$ is the same, the values of R(0) and $R_1(0)$ roughly coincide.

5. Conclusions

The expressions obtained for R90) and $R_1(0)$ speak in favor of the view that, with increasing separation of the frequencies ω_1 and ω_2, the coefficients of mutual correlation of the noises decline rather rapidly and by about the same law in a system with a linear detector and in a system with no detector.

It should be noted that about the same results should be expected when a square-law detector is used in a radio receiver, since the function of correlation of the low-frequency bands in this case is also proportional to $r^2(x + \tau - y)$.

The calculations presented for relations R(0) = F(k) and $R_1(0) = F_1(k)$ permit the conclusion that for radio equipment the voltages $U_1(t)$ and $U_2(t)$ may be considered uncorrelated if $k \geq 4 - 5$ (i.e., if the separation between the resonant frequencies of the filters Φ_1 and Φ_2 exceeds the pass band of one of them by a factor of not less than four or five).

As a rule, in a considerable part of the radio devices met in practice, there is always such a separation of the resonant frequencies of the separation filters.

In conclusion I may state that the parameters of the resonant system of a radio receiver likewise affect the function of mutual correlation of the noises at the output of the separating filters.

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BIBLIOGRAPHY

1. Bunimovich,V.I. - Fluctuation Processes in Radio Receiving Equipment. Publishing House "Sovetskoye Radio", 1951

2. Karnovskiy,M.I. - "Contribution to the Question of Energetic Summation." Collection of Papers. Institute of Motion Picture Engineers, No.II,1954

3. Karnovskiy,M.I. - Contribution to the Question of the Interference of Complex Signals. Trudy komm.po akust, No.8 (1955)

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STUDY OF SHOCK EXCITATION AND FORCED QUENCHING OF OSCILLATIONS OF QUARTZ

by

T.N.Yastrebtseva and I.G.Akopyan

A method of shock excitation of the oscillations of quartz by the aid of a balanced circuit is considered. Several different methods of forced quenching of the oscillations so obtained were studied. A circuit is given which allows rapid excitation and quenching of the oscillations of quartz, and is suitable for practical use.

1. Introduction

Modern pulse engineering makes widespread use of devices creating an electrical time scale. The accuracy of time measurement here is determined by the stability of the marker generator.

It is well known that quartz resonators have highly stable oscillatory properties. When a quartz resonator is used to measure the duration of single, nonperiodic and unsynchronized processes it is necessary to have the oscillations of the quartz crystal occur with constant initial phase and amplitude at an arbitrary instant of time coinciding with the beginning of the process under investigation. It is obviously possible to satisfy this requirement by using shock excitation of quartz, followed by its damping to complete rest, by the arrival of the following exciting pulse.

Papers (Bibl.1,2) have been devoted to the production of marker pulses by the aid of the quartz resonator, but the circuits proposed in them have a number of shortcomings.

It is therefore of interest to make a further study of various methods of exciting and quenching the oscillations of a quartz resonator, which may be utilized to build a reliably operating pulse generator of high-precision marker signals.

The present paper presents a theoretical and experimental investigation of shock excitation of quartz oscillations, and considers several methods for their forced quenching.

2. Shock Excitation of a Quartz Resonator

To study the shock excitation of quartz, the balanced circuit shown in Fig.1 was assembled. The exciting signal, applied to the input of a paraphase amplifier, through the capacitance C_o, equal to the capacitance of the crystal-holder, is fed to the quartz crystal Kv. At the instant of shock, the quartz may be considered in first approximation to be of capacitance C_o. At the output (points a,b) the exciting signal, when the circuit is in exact balance, may be considered to be practically absent; at the same time a signal taken from the crystal appears between these points. Thus only the oscillations of the crystal will act on the input of the amplifier, connected to the points a,b.

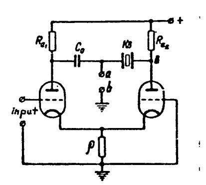

Fig.1

a) Input

In calculating the amplitude and phase of the oscillations being excited, the ordinary equivalent circuit of the quartz is used. The equivalent circuit of Fig.1, used in the calculation, is presented in Figs.2. Here $r = \frac{2R_aR_i}{R_a + R_i}$; $K = \frac{\mu R_a}{R_a + R_i}$; L, C, R, C_o = equivalent parameters of the crystal under study.

The computation was performed by the operational method. Two cases were investigated:

a) Excitation of the quartz by a voltage drop.

Denoting the voltage on the plates of the quartz by u (t), we find the transfer function h (t) for an input voltage of the form E (t) = $\sigma_o(t)$, where $\sigma_o(t)$ is a unit function,

$$h(t) = A + Be^{p_1 t} + e^{\theta t} G \sin(\omega t + \varphi). \qquad (1)$$

The coefficients A, B, G, θ , φ , p_1 are expressed in terms of the constants of the system, and, as shown by research, $A \approx 1$, $B \approx -1$, $p_1 < 0$, $G > 0$, $\theta < 0$.

The first two terms in eq.(1) give an increasing exponent, the third gives damped harmonic oscillations.

b) Excitation of the quartz by a rectangular pulse of amplitude E_o and duration τ.

In this case we have, for $t < \tau$, $u(t) = E_o h(t)$.

For $t > \tau$,

$$u(t) = E_o B e^{p_1 t}(1 - e^{-p_1 \tau}) + E_o G \{ e^{\theta t} \sin(\omega t + \varphi) - e^{\theta(t-\tau)} \sin[\omega(t - \tau) + \varphi]\}. \qquad (2)$$

Let us investigate the relation between the oscillating part, $u_\sim(t)$ and the parameter τ.

For $\theta\tau \ll 1$,

$$u_\sim(t) = 2E_o G e^{\theta t} \sin \frac{\omega\tau}{2} \cos(\omega t + \alpha), \qquad (3)$$

where $\alpha = \varphi - \frac{\omega\tau}{2}$.

Thus the amplitude of the useful signal depends on τ and may vary from 0 at $\tau = kT$ to $2E_oG$ for $\tau = (k + \frac{1}{2})T$, where $T = \frac{2\pi}{\omega}$, k = 0, 1, 2, 3 ...

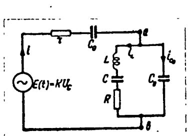

Fig.2

To compare the theoretical results with the experimental data, the formulas obtained were calculated for a quartz generator, type K-11 (cut X + 5°), of frequency 100 KC. The calculations showed that the coefficient G in eq.(3) equals 7.65×10^{-4}, i.e., that the amplitude of the useful signal is about 1 [illegible] the amplitude of the exciting pulse. The results of the calculation agreed with the experimental data within the limits of accuracy of measurement.

The relation between the amplitude of the oscillations and the duration of the exciting pulse that had been found from the calculation was also checked experimentally. Figure 3 gives the results of the measurements.

Oscillograms of the oscillations of the quartz crystal were taken at various durations of the exciting pulse (Fig.4). As will be seen from them, at $\tau = \frac{T}{2} = 5$ microsec (Fig.4b), the oscillations excited are close to sinusoidal and have a considerable amplitude. At $\tau = T = 10$ microsec, (Fig.4c), when according to the calculations no oscillations should be excited, they still do exist, but have a very small amplitude and a complex form, differing in frequency from the fundamental oscillations. At intermediate values of τ (for instance, Fig.4a shows an oscillogram at $\tau = 0.3$ microsec) the form of the oscillations excited is also complex, indicating the presence of oscillations of various frequencies.

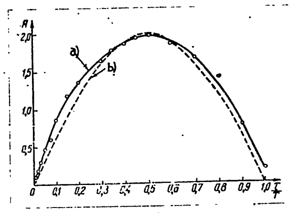

Fig.3

a) Experimental; b) Theoretical

These data tend to support the view that it is desirable to use particularly monofrequent quartz crystals for producing time marks.

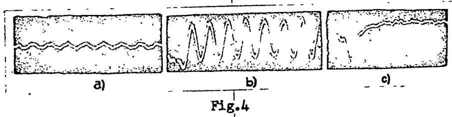

Fig.4

Studies were made on quartz plates vibrating in flexure. The plates, figured for vibration in thickness, gave poor results.

3. Quenching the Oscillations of a Quartz Resonator

The forced quenching of a quartz resonator was studied by various methods:

a) Quenching the Oscillations of the Quartz Crystal by Shunting of an Active Resistance

The simplest method of quenching the oscillations in ordinary resonant circuits is by shunting the circuit with an active resistance. A check of this method for quartz is of interest. In an earlier reference (Bibl.3) the impossibility of effective quenching of the oscillations by this method has been demonstrated. In the present paper the influence of the value of the shunting resistance on the damping of the quartz crystal is calculated and experimentally checked. Figure 5 gives an equivalent circuit for a quartz crystal shunted by an active resistance. The equation of this system is of the form:

$$\frac{d^3 i_L}{dt^3} + (\gamma + \alpha)\frac{d^2 i_L}{dt^2} + \left(\omega_0^2 + \frac{1}{LC_0} + \alpha\gamma\right)\frac{di_L}{dt} + \omega_0^2 \gamma i_L = 0, \quad (4)$$

where: $\gamma = \frac{1}{rC_0}$, $\alpha = \frac{R}{L}$, $\omega_0^2 = \frac{1}{LC}$.

The character of the damping of the quartz crystal may be established by studying the roots of the characteristic equation. It will be easily seen that at $r = 0$ the characteristic equation takes the form $p^2 + \alpha p + \omega_0^2 = 0$, and the damping decrement of the system equals the natural damping decrement of the free quartz crystal. The influence of the intermediate values of p was investigated by numerical solution of the characteristic equation.

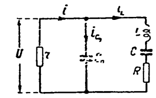

Fig.5

The calculations showed the damping to be maximum at a definite value of r about $\frac{1}{\omega_0 C_0}$, but to increase by only one order of magnitude over the value for the free damping of the crystal. Figure 6 gives a comparison of the calculation with the experimental values (θ = damping factor of the system).

b) Damping of Oscillations in a Circuit with Negative Feedback

Figure 7 shows the simplest circuit for imposing a negative feedback on the quartz crystal. Its equivalent circuit is analogous to the circuit given in Fig.2. The equation of the system is of the form:

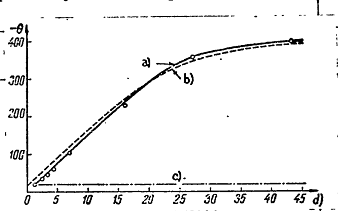

Fig.6

a) Experimental; b) Theoretical; c) Natural damping; d) r, kiloohms

$$\frac{d^3 i_L}{dt^3} + \left[\frac{R}{L} + \frac{1+K}{rC_0}\right] \times \frac{d^2 i_L}{dt^2} + \left[\frac{1}{LC} + \frac{1}{LC_0} + \frac{R(1+K)}{LrC_0}\right]\frac{d i_L}{dt} + \frac{1+K}{LCrC_0}\, i_L = 0, \qquad (5)$$

where K – overall amplification factor of the system.

If we put $r^* = \frac{r}{1+K}$, then we arrive at the same eq.(4), which describes the behavior of a quartz crystal shunted by the resistance r*.

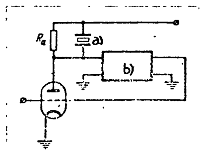

Fig.7

a) Quartz crystal;

b) Amplifier

Thus the variation of K is equivalent to the variation of r. The case r → 0 corresponds to the case K → ∞. For K = 0, r* = r. Consequently such a method cannot provide a greater damping than that given by shunting with an active resistance of optimum value.

c) Quenching the Oscillations of a Quartz Crystal in a Balanced Circuit

The circuit shown in Fig.1 was now studied. To effect the quenching, the oscillations of the crystal from the points (a, b) must be fed through an amplifier and phase converter back to the input of the circuit. Figure 8 gives an equivalent circuit for this connection.

The equation of the system is of the form:

$$\dddot{u} + (\alpha+\gamma)\ddot{u} + \left(\omega_0^2 + \frac{1}{LC_0} + \alpha\gamma\right)\dot{u} + \left(\omega_0^2\gamma + \gamma\frac{1-\frac{K}{2}}{2LC_0}\right)u = 0, \qquad (6)$$

where $\alpha = \frac{R}{L}$, $\gamma = \frac{\gamma}{\gamma C_0}$, $\omega_0^2 = \frac{1}{LC}$.

For $K = 4\frac{C_0}{C} - 2 \approx 4\frac{C_0}{C}$ the coefficient of the last term of eq.(6) vanishes.

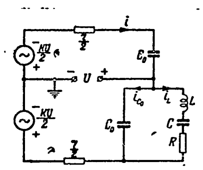

Fig.8

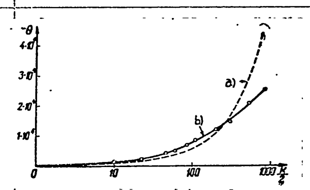

Fig.9

a) Theoretical; b) Experimental

For such a value of K, the roots of the characteristic equation are of the form:

$$p_{1,2} = -\frac{\alpha+\gamma}{2} \pm \sqrt{\left(\frac{\alpha+\gamma}{2}\right)^2 - \left(\omega_0^2 + \frac{1}{LC_0} + \alpha\gamma\right)}.$$

As may be seen from a calculation, the damping is characterized by two summands: α and γ. In practice, $\alpha \approx 10^2$, $\gamma \approx 10^5 - 10^7$, i.e., the damping has been increased by several orders of magnitude over the natural damping.

The relation θ (K) was studied for a specific quartz resonator by numerical solution of the characteristic equation. The same relation was also studied experimentally. The calculated damping, and that experimentally found (Fig.9) are over three orders of magnitude greater than the natural damping of the quartz crystal, which permits the successful practical use of such a system of forced damping. Thus, for instance, an amplification factor of 200 in the feedback circuit assures a decrease in amplitude by a factor of 2.8 in only eight periods.

d) Shock Quenching of a Quartz Resonator

It was found in studying shock quenching that by varying the duration of the

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exciting pulse the amplitude of the excited oscillations could be varied. If a rectangular pulse, however, is caused to act on a quartz crystal already excited, then, by varying the duration of the pulse, its amplitude, and its delay with respect to

Fig.10

the exciting pulse, the amplitude of the oscillations can be considerably reduced (almost complete quenching).

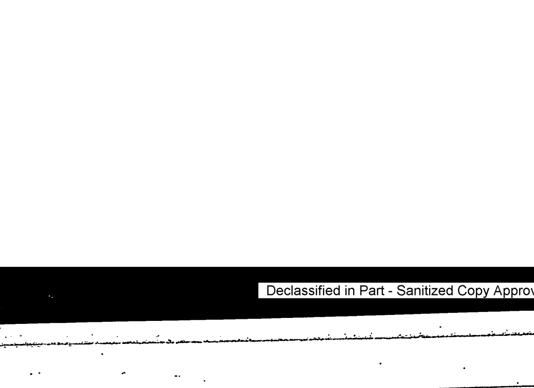

Fig.11

The oscillograms of Fig.10 show shock quenching for the case when the exciting and quenching pulses are of the same duration ($\tau = \frac{T}{2} = 5$ microsec), while the delay time of the second pulse with respect to the first is varied (Fig.10a : t_{del} = 5 microsec; Fig.10b : t_{del} = 10 microsec; Fig.10c : t_{del} = 15 microsec).

It is advisable to use this method of quenching in cases when the amplitude of

the oscillations must be rapidly reduced. Complete quenching is difficult, owing to the non-monofrequency of quartz crystals that has already been mentioned.

As a result of our investigations, a balanced circuit of excitation and quenching of the oscillations of quartz crystals was assembled. The complete circuit diagram is given in Fig.11. The operation of the auxiliary equipment, permitting quenching of the oscillations to be accomplished at any desired instant after their excitation, is not considered, in view of its simplicity.

Thus the equipment presented allows the oscillations of a quartz crystal to be quenched after a few periods, the instant of turning on the feedback and its duration being easily changed.

4. Conclusions

A theoretical and experimental investigation of the shock excitation of a quartz crystal has been performed. A complete solution has been obtained for the oscillations excited for the cases of action of a signal of stepped or rectangular shape on the quartz crystal has been found.

The optimum pulse duration at which the excited oscillations are at maximum amplitude and have the smallest number of harmonics has been established.

Four methods of forced quenching of quartz crystal oscillations have been studied: shunting the crystal with an active resistance, quenching in a circuit with negative feedback, quenching by means of a balanced circuit, and shock quenching.

It has been found that the damping of the quartz crystal can be increased by only one order of magnitude when the first two of these methods are used.

Calculation and experimental study of the quenching of quartz crystal oscillations by means of the balanced circuit proposed in the present paper have shown that the damping of the quartz crystal oscillations may be increased 10^4 times by this method. Further than that, the quality factor of the crystal may be artificially reduced to any required value by permanent feedback connection.

A balanced circuit has been developed for exciting and quenching quartz-crystal oscillations. The circuit is suitable for practical use.

This work has been done at the Department of Oscillation Theory, Faculty of Physics, Moscow State University. The authors express their thanks to Professor V.V. Migulin for suggesting the theme of this research and for his help during its execution.

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BIBLIOGRAPHY

1. Mynall,D.J. - Jour.Inst.Elec.Engrs. 93, III-A (7), (1946),pp.1207-1214
2. Yastrebtseva,T.N. and Galkin,O.P. - This Journal. (7), (1955),pp.69-73
3. - Proc.Inst.Radio Engrs. 29(4), (1941),pp.195-199

FEEDBACK IN TRANSISTOR CIRCUITS

by

Ya.K.Trokhimenko

Full Member of the Society

This paper gives a technique of analysis of the principal types of feedback in transistor circuits.

1. Introduction

The role of feedback in transistor circuits is even more important than in circuits using vacuum tubes. There are two reasons for this.

1. The transistor has a deeper internal feedback than the vacuum tube.

2. In modern transistors the instability of the equivalent parameters is considerably greater than in vacuum tubes.

The technique of designing radio is controlled primarily by the properties of the active elements. The transistor, as an element of the circuit, differs from the vacuum tube in the value and character of its feedback and in its low input resistance. It is the last-named fact that is primarily responsible for the specific features of the analysis of circuits using transistors. Thus, depending on the ratio of the internal resistance R_g of the signal source and the input resistance R_{in}, amplifiers may be arbitrarily divided into three classes:

1. Voltage amplifiers, for which the following inequality holds:

$$R_{in} \gg R_g.$$

2. Current amplifiers, for which the reverse inequality holds, i.e.:

$$R_{in} \ll R_g.$$

3. Power amplifiers, when R_{in} and R_g are comparable.

For electron tubes, the inequality $R_{in} \gg R_g$ usually holds, a fact which considerably simplifies the calculation of vacuum-tube circuits, allowing the replacement of a tube by an active dipole controlled by the input voltage (Fig.1a). Such an inequality can hold only for a very limited number of transistor circuits (for instance, in circuits with a common collector at very low values of the internal resistance of the signal source) and it is only in these circuits that the substitution circuit of Fig.1a and the technique of analysis corresponding to it are applicable.

Transistor circuits for which the opposite inequality $R_{in} \ll R_g$ holds are considerably more frequent. Such circuits may be analyzed by means of the equivalent circuit of Fig.1b, and, in particular, by the aid of a technique of calculation based on the dual analogy between the input circuits of vacuum tubes and transistors. In the equivalent circuits of Figs.1a and 1b, the amplifier proper is described by only two equivalent parameters, which considerably simplifies the analysis. In the overwhelming majority of transistor circuits, however, the input resistance is comparable to the resistance of the signal source, which requires the replacement of the amplifier, not by a dipole, as in the cases discussed above, but by a quadripole (Fig.1c), described by four equivalent parameters. It must be noted that in calculating the input circuit (especially in cases where the resistance of the signal source is assigned), a power amplifier may also be reduced to an active dipole, analogous, for example, to the circuit of Fig.1a for

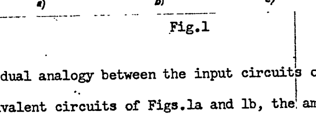


Fig.1

$$\mu = \frac{R_{21}}{R_{11}+R_g}, \quad R_i = R_{22} - \frac{R_{12}R_{21}}{R_{11}+R_g},$$

or to the substitution circuit of Fig.1b; but in this case the equivalent circuit will replace not only the amplifier, but also the passive part of the input circuit.

In the general case the transistor amplifier must be replaced by an active quadripole, while use of the methods of quadripole theory is advisable in the analysis, and particularly the use of methods based on the application of matrix algebra, which considerably simplifies the calculation.

2. Technique of Analysis of Transistor Circuits with Feedback

In a linear circuit, approximating the physical circuit for weak signals, the variation of voltage is related linearly to the variation of current, so that the division of the feedback into current feedback and voltage feedback is in large measure arbitrary. From the point of view of the connection of the feedback circuit to the amplifier itself, however, and also in connection with the fact that in an actual circuit the difference between current feedback and voltage feedback acquires substantial importance with nonlinear loads, it is advisable to classify circuits with feedback between the output and input into the following principal types: (1) parallel feedback (Fig.2a); (2) series feedback (Fig.2b); (3) series-parallel feedback (Fig.2c); and (4) parallel-series feedback (Fig.2d). It is easy to see that any feedback circuit may be reduced to one of these four types of feedback or to some combination of them. Depending on the value and sign of the parameters of the feedback circuit, that circuit may be either positive or negative, active or passive, etc.

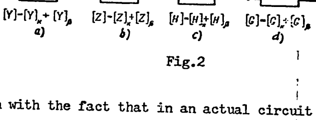


Fig.2

To analyze each of these fundamental types of feedback, it is convenient to make use of matrices of equivalent parameters corresponding to them, i.e., matrices of the constant coefficients of the equations:

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} Y_{11} Y_{12} \\ Y_{21} Y_{22} \end{bmatrix} \begin{bmatrix} u_1 \\ u_2 \end{bmatrix}, \qquad (1)$$

$$\begin{bmatrix} u_1 \\ u_2 \end{bmatrix} = \begin{bmatrix} Z_{11} Z_{12} \\ Z_{21} Z_{22} \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \end{bmatrix}, \quad (2)$$

$$\begin{bmatrix} u_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} H_{11} H_{12} \\ H_{21} H_{22} \end{bmatrix} \begin{bmatrix} i_1 \\ u_2 \end{bmatrix}, \quad (3)$$

$$\begin{bmatrix} i_1 \\ u_2 \end{bmatrix} = \begin{bmatrix} G_{11} G_{12} \\ G_{21} G_{22} \end{bmatrix} \begin{bmatrix} u_1 \\ i_2 \end{bmatrix}. \quad (4)$$

The matrix of the coefficients of eqs.(1) to (4) may be written as follows:

$$[k] = \begin{bmatrix} k_{11} k_{12} \\ k_{21} k_{22} \end{bmatrix}, \quad (5)$$

and the elements of the matrix, eq.(5), in accordance with their roles in eqs.(1) to (4), may be termed: k_{11}, the input parameter; k_{12}, the feedback parameter; k_{21}, the transmission parameter or the amplification parameter; k_{22}, the output parameter.

For the analysis of the principal feedback circuits any desired systems of equivalent parameters may be used; but the calculations are the simplest when the corresponding system of parameters is used for each type of feedback. In particular, in this case the matrix of the entire amplifier is equal to the sum of the matrices of the amplifier proper and of the feedback circuit, as shown in Fig.2.

It is convenient to perform the analysis of feedback for power amplifiers by the aid of the voltage amplification factor, which is connected with the power amplification factor by the simple relation:

$$K_M = \frac{4R_g}{R_n} K^2. \quad (6)$$

The voltage amplification factor in the general form may be written as follows:

$$K = w \frac{\Delta_{12}}{\Delta}, \quad (7)$$

where Δ = determinant of matrix of the corresponding equivalent parameters of the amplifier, including the external circuits as well;

Δ_{12} = algebraic complement of the element k_{12} of the matrix;

w = parameter whose dimensionality and value depend on the system of equivalent parameters. Thus, for equivalent impedance, $w = Z_H$; for equivalent conductivities, $w = Y_G$; for series-parallel H-parameters, $w = 1$; for parallel-series G-parameters, $w = Z_g Y_H$.

For a circuit with feedback, the amplification factor is equal to

$$K_{fdb} = w \frac{\Delta_{12}}{\Delta} = \frac{K}{A}, \quad (8)$$

where $A = K/K_{fdb}$ = coefficient of feedback. For the analysis it is more convenient to use, not the coefficient of feedback, but the feedback difference F (Bibl.1), which equals the ratio of the determinant of the entire circuit to the determinant from which the given element of feedback is absent:

$$F = \frac{\Delta}{\Delta^{\circ}}. \quad (9)$$

By simultaneous solution of eqs.(8) and (9), we find

$$K_{fdb} = w \frac{\Delta_{12}}{\Delta} = w \frac{\Delta^{\circ}_{12}}{\Delta^{\circ}} \cdot \frac{\Delta_{12}/\Delta^{\circ}_{12}}{\Delta/\Delta^{\circ}} = \frac{Km}{F}. \quad (10)$$

Since the parameter $m = \frac{\Delta_{12}}{\Delta^{\circ}_{12}}$, showing the influence of feedback on the transmission parameter $\Delta_{12} = k_{21}$, is rather close to unity, we may consider the feedback difference for the given feedback element to be approximately equal to the coefficient of feedback for the given element, i.e., that

$$A = F/m \approx F. \quad (11)$$

Equation (11) enables us to use instead of the quantity A the quantity F, which is simpler to calculate. The criterion of such substitution is that the parameter m shall differ from unity.

In the case where the feedback is being determined for a given channel, then to the value of K in eqs.(8) and (10) will correspond the amplification in the given channel in the absence of feedback, but in the possible presence of feedback in other channels, this feedback may be termed the original feedback. In the absence of external feedback in a transistor amplifier, the original feedback is equal to the internal feedback of the transistor itself. The influence of internal feedback on amplification in this case may be expressed by the relation:

$$K = \frac{K_0}{A_0}, \tag{12}$$

where K_o = amplification factor when the feedback parameter of the transistor vanishes, i.e., when $k_{12} = 0$; and A_o = feedback with respect to the parameter k_{12}.

For an amplifier with multi-channel feedback, the following expression holds:

$$K_{fdb} = \frac{K_0}{A_0 A_1 A_2 \ldots A_n}, \tag{13}$$

where A_j is determined in the absence of any feedback with a number greater than j with respect to the original feedback for the given channel.

The stabilizing action of feedback may be expressed by the aid of the "sensitivity" S of the amplification to the variation of the element w of the circuit, defined as the ratio of the percentage change in the value of the given element to the percentage change in the amplification factor (Bibl.1):

$$S = \frac{\partial w/w}{\partial E_{out}/E_{out}} = \frac{\partial w/w}{\partial K/K}. \tag{14}$$

The concept of sensitivity may also be utilized to determine the stabilizing action of feedback on the other characteristics of the circuit (for instance, on R_{in} or R_{out}, etc.). In this case, the value of the sensitivity

$$S(w) = \frac{\partial w/w}{\partial P/P} \tag{15}$$

Table 1

Calculation of Circuits with Direct Feedback

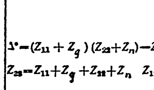
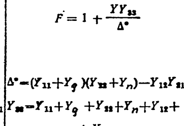
Type of Feedback	Series	Parallel
Equivalent circuit		
Matrix of circuit parameters	$\begin{bmatrix} Z_{11}+Z_g+Z & Z_{12}+Z \\ Z_{21}+Z & Z_{22}+Z_{11}+Z \end{bmatrix}$	$\begin{bmatrix} Y_{11}+Y_g+Y & Y_{12}-Y \\ Y_{21}-Y & Y_{22}+Y_{11}+Y \end{bmatrix}$
Parameter $m = \Delta_{12}/\Delta^o_{12}$	$\frac{Z_{21}+Z}{Z_{21}}$	$\frac{Y_{21}-Y}{Y_{21}}$
Feedback difference F	$F = 1 + \frac{ZZ_{33}}{\Delta^\circ}$ $\Delta^\circ = (Z_{11}+Z_g)(Z_{22}+Z_n) - Z_{12}Z_{21}$ $Z_{33} = Z_{11}+Z_g+Z_{22}+Z_n \quad Z_{12}-Z_{21}$	$F = 1 + \frac{YY_{33}}{\Delta^\circ}$ $\Delta^\circ = (Y_{11}+Y_g)(Y_{22}+Y_n) - Y_{12}Y_{21}$ $Y_{33} = Y_{11}+Y_g+Y_{22}+Y_n+Y_{12}+$ $+Y_{21}$
Input impedance Z_{in}	$Z_{in} = \frac{\Delta_{in}}{Z_{22}+Z_n+Z}$ $\Delta_{in} = \Delta$ for $Z_g = 0$	$Z_{out} = \frac{Y_{22}+Y_n+Y}{\Delta_{in}}$ $\Delta_{in} = \Delta$ for $Y_g = 0$
Output impedance Z_{out}	$Z_{out} = \frac{\Delta_{out}}{Z_{11}+Z_g+Z}$ $\Delta_{out} = \Delta$ for $Z = 0$	$Z_{out} = \frac{Y_{11}+Y_g+Y}{\Delta_{out}}$ $\Delta_{out} = \Delta$ for $Y_n = 0$
Value of feedback element for an assigned F or A	$Z = (F-1)\frac{\Delta^\circ}{Z_{33}} \approx (A-1)\frac{\Delta^\circ}{Z_{33}}$	$Y = (F-1)\frac{\Delta^\circ}{Y_{33}} \approx (A-1)\frac{\Delta^\circ}{Y_{33}}$

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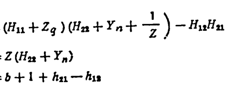
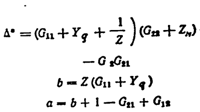
Table 1 (cont.)

Series-Parallel	Parallel-Series
$Z = Z_1 + Z_2 \quad q = Z_1/Z$  б)	$Z = Z_1 + Z_2 \quad q = Z_1/Z$  d)
$\begin{bmatrix} H_{11}+Z_g+Z(q-q^2) & H_{12}-q \\ H_{21}+q & H_{22}+Y_n+\frac{1}{Z} \end{bmatrix}$	$\begin{bmatrix} G_{11}+Y_g+\frac{1}{Z} & G_{12}+q \\ G_{21}-q & G_{22}+Z_n+Z(q-q^2) \end{bmatrix}$
$\frac{H_{21}+q}{H_{21}}$	$\frac{G_{21}-q}{G_{21}}$
$F = 1 + \frac{q}{\Delta^*}[a - qb]$	
$\Delta^* = (H_{11}+Z_g)(H_{22}+Y_n+\frac{1}{Z}) - H_{12}H_{21}$; $b = Z(H_{22}+Y_n)$; $a = b + 1 + h_{21} - h_{12}$	$\Delta^* = (G_{11}+Y_g+\frac{1}{Z})(G_{22}+Z_n) - G_{12}G_{21}$; $b = Z(G_{11}+Y_g)$; $a = b + 1 - G_{21} + G_{12}$
$Z_{in} = \frac{\Delta_{in}}{H_{22}+Y_n+\frac{1}{Z}}$; $\Delta_{in} = \Delta$ for $Z_g = 0$	$Z_{in} = \frac{G_{22}+Z_n+Z(q-q^2)}{\Delta_{in}}$; $\Delta_{in} = \Delta$ for $Y_g = 0$
$Z_{out} = \frac{H_{11}+Z_g+Z(q-q^2)}{\Delta_{out}}$; $\Delta_{out} = \Delta$ for $Y_n = 0$	$Z_{out} = \frac{\Delta_{out}}{G_{11}+Y_g+\frac{1}{Z}}$; $\Delta_{in} = \Delta$ for $Z_n = 0$
$q = \frac{a}{2b} \pm \sqrt{\frac{a^2}{4b^2} - \frac{\Delta^*(F-1)}{b}} \approx \frac{a}{2b} \pm \sqrt{\frac{a^2}{4b^2} - \frac{\Delta^*(A-1)}{b}}$	

will determine the ratio of the percentage change of the value of w to the percentage change of the value of P, with which we are concerned.

The stabilization under variation of the circuit parameters may be evaluated by the aid of the stabilization factor

$$N = \frac{S_1(\omega)}{S_2(\omega)}, \tag{16}$$

which is equal to the ratio of the sensitivity of the circuit under different conditions (variation of the depth of feedback, variation of the value or configuration of circuit elements, etc.).

The general relations presented above are fundamental for the analysis of feedback circuits.

3. <u>Properties of the Principal Types of Feedback</u>

With the same coefficient of feedback, different feedback circuits have different effects on the other characteristics of the amplifier (especially on its input and output resistance) and on the stability of the parameters under the action of destabilizing factors. The computational formulas obtained from the general relations we have considered are usually rather unwieldy, but they may be considerably simplified by taking account of the relations between the values of the parameters of a specific circuit. For illustrative purposes, Table 1 gives computational formulas for an amplifier with direct (transformerless) passive feedback. Each of these expressions may be simplified without impairing the accuracy of the calculation. Thus, for a single-stage amplifier using a transistor with grounded emitter, the usual assumptions may be adopted:

$$Y_{12} \ll Y_{21};\; Y_{12} \ll Y_{11};\; Y_{22} \ll Y_{21}.$$

Under these assumptions, the curves of the relation between the terminating resistances of the amplifier and the coefficient of feedback (Figs.3 and 4), as well as the other characteristics, are practically in agreement with the experimental

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curves.

As will be seen from the graphs of Figs.3 and 4, for a circuit with grounded emitter the series-parallel feedback, which raises the input resistance and lowers the output resistance, is the most advantageous, with respect to the terminating resistances, while the parallel-series feedback is the least advantageous. The absence of a common junction point with the load and signal source is a disadvantage shared by both these types of direct feedback.

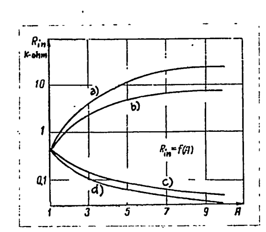

Fig.3

a) Series-parallel (R = 50 kiloohms) feedback; b) Series feedback; c) Parallel-series (R = 20 kiloohms) feedback; d) Parallel feedback

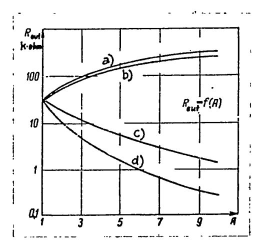

Fig.4

a) Parallel-series feedback; b) Series feedback; c) Series-parallel feedback; d) Parallel feedback

The formulas for calculating the stability of the amplifier parameters for different imposition of the feedback are more complicated; but they, too, may be considerably simplified by taking account of specific relations between the values of the circuit parameters. The experimental data and the calculation of the feedback effect on the stability of the amplifier parameters (by instability of parameters may be meant not only temperature or time instability, but also the deviation of the value of the equivalent parameters from the nominal values due to the influence of fluctuations, of the nonlinearity of characteristics, on the replacement of transistors, etc.) show that, for circuits with grounded base and grounded emitter, series and series-parallel feedbacks give the maximum degree of stabilization of the input equivalent parameter k_{11}, while parallel feedback or parallel-series feedback give the maximum stabilization of the output parameter k_{22}. The amplification parameter k_{21} is stabilized to about the same extent by all forms of feedback. With a sufficiently deep external feedback, the influence of the parameter of internal feedback k_{12} may be neglected. The application of DC feedback permits rather deep stabilization of the working point of a transistor.

4. Conclusion

The application of feedback allows considerable diminution of the influence of the equivalent parameters and of the influence of the internal feedback of transistors, which influences, in a number of cases, have a markedly adverse effect on the operation of semi-conductor equipment. It must be noted that a number of questions connected with the use of feedback in transistor circuits (especially the conditions of stability and the phase ratios) have up to now been given absolutely inadequate study. The development of a clear-cut engineering theory of the application of feedback to transistor circuits will permit substantial improvement of the working characteristics and operational properties of articles using semiconductor instruments, especially in their mass production.

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BIBLIOGRAPHY

1. Bode,G. - Circuit Theory and Design of Amplifiers Using Feedback. GIIL. 1948

DIMINISHING THE PULSE-FRONT DISTORTIONS IN VIDEO AMPLIFIERS USING JUNCTION TRANSISTORS

by

T.M.Agakhanyan

The author discusses a method of diminishing the distortions of the pulse fronts due to dispersion of the velocities of the minority carriers in the base region of a junction transistor.

1. Introduction

The distortions of the pulse front in video amplifiers using junction transistors are due mainly to two groups of phenomena. The first group is connected with the presence of parasitic capacitances. The methods of diminishing this type of distortion are well known from the theory of vacuum-tube video amplifiers. The second group is connected with the processes taking place in the base region of the junction transistor. These processes - diffusion, on the one hand, and impairment of the thermodynamic equilibrium between the processes of recombination and the process of thermal production of carriers, on the other (Bibl.1) - lead to a time shift of the input and output pulses and to an increase in the build-up time of the front (Bibl.2,3).

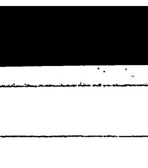
Fig.1

The build-up time of the front may be shortened by choosing parameters such that, at equal amplitudes, more carriers shall arrive in the region of the base, the steeper the pulse front. Such a redistribution of the minority carriers leads to an increased flow of carriers at the output during sharp drops of the input pulse, so that the build-up time of the front is shortened.

2. Circuit with Complex Feedback

The method of redistributing the carriers in the base region may be very clearly illustrated by the example of a circuit with complex current feedback (Fig.1). The input resistance of this circuit is the lower, the steeper the front of the input pulse, since with steep voltage drops the feedback circuit R_oC_o is almost inoperative. With decreasing input resistance, the flow of carriers into the base region increases, and therefore the distortion of the front of the output pulse diminishes.

Figure 2 gives a substitution circuit for such a stage (the equivalent circuit for the transistor is surrounded by a dashed line). In this circuit, r_b is the pure ohmic resistance of the base layer;

$$r_K = r_{22} - r_b \text{ and } r_e = r_{11} - r_b \qquad (1)$$

are the resistances, and C_k and C_e the capacitances, of the junctions (with subscript k, that of the collector junction, with subscript e, that of the emitter junction); and αi_e and αi_k, the current generators.

The operator expression of the mutual conductance of the circuit of Fig.2 is as follows:

$$S_c = \frac{di_K}{du_q} = \frac{\frac{Z_e + Z_o}{Z_K} - \alpha}{(R_q + r_b)\left[1 - \alpha + \frac{(1-\alpha_u) Z_e + Z_n + Z_o}{Z_K}\right] + (Z_e + Z_o)\left[1 + \frac{Z_n}{Z_K}\right] - Z_e \alpha \alpha_K}, \qquad (2)$$

where

$$Z_e = \frac{r_e}{p r_e C_e + 1}; \qquad Z_K = \frac{r_K}{p r_K C_K + 1}; \qquad Z_o = \frac{R_o}{p R_o C_o + 1}. \qquad (3)$$

Exact determination of the original S_c from eq.(2) is difficult. It may be considered, with an accuracy sufficient for practice, that

$$S_c = -\frac{\alpha_o}{(R_q + r_b + Z_e + Z_o)(p\tau_\alpha + 1) - \alpha_o[(R_q + r_b) + Z_e \alpha_{uo}]}. \qquad (4)$$

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In this expression we have neglected the shunting action of Z_k. Instead of the exact relationship for (Bibl.1,4), we used the approximate expression (Bibl.3):

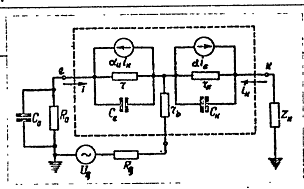

Fig.2

$$\alpha = \frac{\alpha_0}{p\tau_\alpha + 1}, \qquad (5)$$

where

$$\alpha_0 = \frac{r_{21} - r_b}{r_{22} - r_b}; \quad \tau_\alpha = \frac{1}{2\pi f_\alpha}, \qquad (6)$$

f_α = boundary frequency of α in the circuit with grounded base, and, finally*

$$\alpha_u \approx \alpha_{u0} = \frac{r_{12} - r_b}{r_{11} - r_b}. \qquad (7)$$

Let us consider circuits using transistors in which overcharging of the capacitance C_e takes place in a time considerably shorter than the build-up time of the pulse front. For such triodes, we may consider $Z_e \approx r_e$, and then

$$S_c = -S_{c0}\,\frac{p + \frac{1}{m\tau_\alpha}}{p^2 + p\left(\frac{a_\alpha}{\tau_\alpha} + \frac{a_e}{m\tau_\alpha}\right) + \frac{a_{\alpha e}}{\tau_\alpha^2 m}}\,\frac{a_\alpha}{\tau_\alpha}, \qquad (8)$$

where

$$S_{c0} = \frac{\alpha_0}{\Delta}; \quad \Delta = R_g + r_b + r_e + R_0 - \alpha_0\,[R_g + r_b + r_e\alpha_{u0}], \qquad (9)$$

$$a_{\alpha e} = \frac{\Delta}{R_g + r_b + r_e}; \qquad a_\alpha = 1 - \frac{\alpha_0\,(R_g + r_b + r\,\alpha_{u0})}{R_g + r_b + r_e}, \qquad (10)$$

$$a_e = 1 + \frac{R_0}{R_g + r_b + r_e}; \quad m = \frac{R_0 C_0}{\tau_\alpha}. \qquad (11)$$

* α_u depends on the rate of growth of the pulse fronts. But in the examples considered, this approximation is rather accurate.

If we select what is termed the coefficient of correction

$$m = \frac{a_e - 1}{a_{\alpha e} - a_e} \equiv 1, \qquad i.e., \quad R_0 C_0 = \tau_\alpha,$$

then

$$S_c(t) = -S_{c0}\left(1 - e^{-\frac{t}{\tau_\alpha}a_{\alpha e}}\right), \qquad (12)$$

while the build-up time

$$t_n = 2{,}2\,\frac{\tau_\alpha}{a_{\alpha e}}. \qquad (13)$$

It will be clear from eq.(13) that by appropriate choice of $a_{\alpha e}$ (practically of R_0) in the circuit of Fig.1, t_n may be reduced to a value equal to, or even less than, the build-up time in the circuit with grounded base. This reduction, however, is accomplished at the cost of decreasing the amplification of the individual stages. This latter circumstance makes it necessary to increase the number of stages, since the overall amplification is usually assigned. It turns out that even when we do increase the number of stages, the total build-up time of the transfer characteristic of an amplifier with feedback is decreased to almost an order of magnitude below that of the ordinary amplifier.

With an assigned amplification in the circuit of Fig.1, the build-up time is shorter, the greater the coefficient of correction m. Thus, for example, for $m = m_{crit}$, where

$$m_{crit.} = \frac{1}{a_\alpha^2}\left[2a_{\alpha e} - a_\alpha a_e - \sqrt{(2a_{\alpha e} - a_\alpha a_e)^2 - a_\alpha^2 a_e^2}\right], \qquad (14)$$

i.e., in the critical state (the roots of the characteristic equation are equal, and are real numbers):

$$S_c(t) = -S_{c0}\left\{1 + \left[\left(\sqrt{a_{\alpha e} m_{crit.}} - 1\right)\sqrt{\frac{a_\alpha}{m_{crit.}}}\cdot\frac{t}{\tau_\alpha} - 1\right] e^{-\frac{t}{\tau_\alpha}\sqrt{\frac{a_{\alpha e}}{m_{crit.}}}}\right\}. \qquad (15)$$

If $a_\alpha < 2$, then the transfer characteristic is a monotone increasing function.

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In this case the build-up time will be shorter than in the case of m = 1 by a factor of $\frac{a_e}{4}\sqrt{8 - a^2}$.

If $\alpha_e > 2$, then the reduction in t_n is still greater, but in this case aperiodic overshoots will appear. Their value may be determined by the formula:

$$A = (\sqrt{a_{\alpha e}\, m_{crit.}} - 1)\, e^{-\frac{\sqrt{a_{\alpha e} m_{crit.}}}{\sqrt{a_{\alpha e} m_{crit.} - 1}}}. \qquad (16)$$

For $m > m_{crit}$, the circuit will operate under oscillatory conditions:

$$S_c(t) = -S_{c0}\left[1 + \frac{\sqrt{a_{\alpha e}\, m\, [(am-1)^2 + b^2]}}{b}\, e^{-\frac{t}{\tau_\alpha} a} \sin\left(\frac{t}{\tau_\alpha} b - \lambda\right)\right], \qquad (17)$$

where

$$\lambda = \operatorname{arctg} \frac{b}{(a^2 + b^2)m - a}; \quad a = \frac{1}{2m}(a_e + ma_\alpha); \quad b = \sqrt{\frac{a_{\alpha e} - ma^2}{m}}. \qquad (18)$$

In the oscillatory regime, the build-up time

$$t_n = 2{,}2\frac{\tau_c}{a_{\alpha e}}\vartheta, \qquad (19)$$

where θ is a coefficient indicating how many times t_n is smaller than it is in the case m = 1. The build-up time is shorter, the greater the allowable value of the first overshoot:

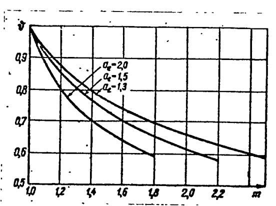

Fig.3

$$A = \sqrt{(am-1)^2 + b^2} \cdot e^{-\frac{t_m}{\tau_\alpha} a} \qquad (20)$$

Here

$$t_m = \frac{\pi - \operatorname{arctg}\frac{bm}{1 - ma}}{b} \quad \text{for } ma < 1 \qquad (21)$$

and

$$t_m = \frac{1}{b}\operatorname{arctg}\frac{bm}{ma - 1} \quad \text{for } ma > 1. \qquad (22)$$

Figures 3 and 4 give the values of ϑ and A related to m and a_e for a_α = 0.05 (when a_α varies from 0 to 0.7, ϑ and A vary only slightly).

When $r_e C_e > \tau_\alpha$, i.e. in circuits using high-frequency transistors (P 1-I, junction transistors), the effect of the capacitance of the emitter junction, C_e, cannot be neglected. This capacitance, too, leads to an increase in the flow of charge carriers into the base region, when the voltage drops are steep, and, consequently, also leads to shortening of the build-up time of the pulse fronts. It is precisely for this reason that the transfer characteristic of a stage with grounded emitter, using a junction transistor, is of oscillatory character at low values of R_g.

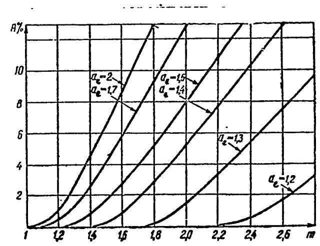

Fig.4

The analysis of the circuit of Fig.1 with high-frequency triodes is somewhat cumbersome in the general form. But in practice, in such circuits, the value of R_0C_0 is found to be close to that of r_eC_e. In this case, i.e., when

$$R_0C_0 = r_eC_e > \tau_\alpha \qquad (23)$$

the mutual conductance S_c may be expressed by eq.(8), and all the quantities necessary for the calculation (t_n, A, etc.) may be determined by the formulas given above, if we consider

$$a_{\alpha e} = \frac{\Delta}{R_g + r_b}; \quad a_\alpha = 1 - \alpha_0; \quad a_e = 1 + \frac{R_0 + r_e}{R_g + r_b} \qquad (24)$$

and

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$$m = \frac{R_0 C_0}{\tau_\alpha}. \qquad (25)$$

3. Circuit with Combination Current Distribution Circuit and Feedback

The source of the input pulses for the intermediate stages of the amplifier is the stages preceding them, which may in practice be represented as current generators. To diminish the distortion of the pulse fronts in such circuits, the use of what are termed current-distributing circuits is recommended (Bibl.3). The simplest of such circuits consists of a resistance and an inductance connected in series with it. A disadvantage of such circuits is that an appreciable shortening of the build-up time can be obtained only at high values of the inductance.

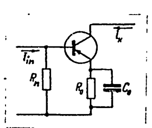

Fig.5

In circuits with a combination of a current-distributing circuit and feedback, induction coils are in practice no longer needed. In this case the current-distributing circuit consists of a resistance connected in parallel with the input of the stage (Fig.5).

The current amplification factor for such a circuit is as follows:

$$K_i = \frac{di_\kappa}{di_{in}} \approx -\frac{\alpha R_n}{(R_n + r_b + Z_e + Z_0)(p\tau_\alpha + 1) - \alpha_0[(R_n + r_b) + Z_e \alpha_{u0}]}. \qquad (26)$$

The structures of the expressions for K_i and S_c are entirely the same, and therefore the formulas for t_n, A, etc. are also the same, if R_n is substituted for R_G in them. For an assigned amplification, the build-up time of the circuit of Fig.5 is the shorter, the greater the values of R_H and R_o. In actual circuits, the resistor R_H is connected to the collector circuit of the preceding triode. An increase in this resistance, or in R_o, makes it necessary to increase the voltage of the supply source. In practice it is sufficient to confine ourselves to the values R_H = 2-3 kiloohms, since on further increase of this resistance the shortening of t_n is very slight.

4. Conclusion

The circuit with complex feedback is merely a simple special case, showing how effectively the distortions of the pulse fronts may be diminished by redistributing the carriers in the base region. The redistribution of the carriers may be effected by the most varied methods: use of feedback, employment of current distributing circuits, etc. The feedback circuits show the greatest promise, since they not only shorten the build-up time but also lead to qualitative and quantitative improvements of many other properties of the amplifier.

This method of diminishing the distortions of the fronts is applicable to pulse equipment designed either for amplification or for shaping pulses with steep drops.

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BIBLIOGRAPHY

1. Adirovich,E.I. and Kolotilova,V.G. - Zhur.eksp.i teor.fiz. 29,No.6(12), (1955), p.770
2. Chow,W.F. and Suran,J.J. - Proc.Inst.Radio Engrs. 41(9),(1953),p.1125
3. Shea,R.F. - Principles of Transistor Circuits. New York 1953
4. Schaffner,J.S. and Suran,J.J. - J.Appl.Phys. 24(11), (1953),p.1355

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THE PRINCIPAL PROPERTIES AND CHARACTERISTICS OF THE SYNCHRONOUS FILTER

by

N.K.Ignat'yev

The author discusses the action of an electric filter designed to separate or absorb oscillations of arbitrary form repeated at an assigned frequency. The basic parameters of the filter are established.

Relations for the frequency, phase and transfer characteristics of the filter are derived. The concept of the filtration factor with respect to the noise voltage is introduced, and the relations of this factor to the parameters of the filter are established.

The possibility of using a long open line as an accumulator for the filter is shown.

1. Introduction

In modern radio engineering it is rather often necessary to separate a complex oscillation into two components, one of which is an arbitrary function of time, periodically repeating at the assigned frequency f_o, while the other is either a nonperiodic or a periodic function of time, repeating at a frequency other than f_o. The first component may be termed synchronous, and the second, asynchronous, with respect to the assigned repetition frequency f_o. In some cases it is necessary to isolate the synchronous component and eliminate the asynchronous (for instance, in controlling interference in radar work); in other cases, on the other hand, the asynchronous must be isolated and the synchronous eliminated (for instance, in the selection of moving radar targets).

A number of devices to solve such a problem have already been constructed, tested, and described in the literature (Bibl.1, 2, 3, 4, 5). According to the purpose, the method of realizing it, and the methods of describing the principal properties, these devices have been given the most varied names. In some cases they are called canceling devices, in others, synchronous accumulators, in others, comb filters or synchronous filters, in still others, integrating devices, and so on.

We shall consider the action of all these devices (regardless of the form in which they are realized, or of their purpose) as special cases of the realization of a particular type of filter, hereafter termed synchronous filter. This paper is devoted to the elucidation of the principal properties and characteristics of such generalized filter.

The synchronous filter is a natural generalization of the resonant filter. In contrast to the usual resonant filter, which is able to separate (or absorb) only sinusoidal oscillations of assigned frequency, the synchronous filter can separate (or absorb) oscillations of arbitrary form, repeated at the assigned frequency f_o, to which it is "tuned".

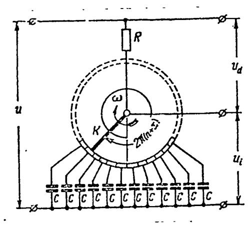

Fig.1

a) Differentiating output; b) Integrating output

The accumulator which must "remember" the form of the oscillations fed to it, is a principal and inseparable part of the synchronous filter. The action of the accumulators may be based on the utilization of the most varied physical phenomena. Thus, for example, the accumulators that have been practically realized and described in the literature, include capacitative storage devices with cathode-ray switching (Bibl.1,2), with vacuum-tube switching (Bibl.4), and with mechanical switching (Bibl.5), electrolytic storage devices (Bibl.3), wave storage devices, storing information in the form of waves propagated along a line, and others. According to the form of the accumulator, the circuit and design of the synchronous filter may be radically modified, but its basic properties as a filter must never-

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theless be maintained unchanged.

As the starting system, serving to elucidate the principle of operation of the synchronous filter, and to obtain the fundamental relations, an idealized system of a synchronous filter with a capacitative storage device (Fig.1) will be used in this paper. It is considered as an equivalent circuit for a synchronous filter of any form.

2. The Synchronous Filter with Capacitative Accumulator

The commutator K, rotating at the angular velocity $\omega_o = 2\pi f_o$, periodically passes a number of contacts, uniformly distributed along an arc of the circumference. Idealizing the operation of the commutator, we shall consider that, in passing from contact to contact, it does not connect them with each other, and at the same time does not remain separated from one of them. A capacitor of capacitance C is connected to each contact; the leads from the opposite plates of the capacitors are connected together. The system consisting of the commutator and the set of capacitors is called an accumulator. The resistance R is connected in series with the commutator.

The input voltage u is applied to the storage tube through the resistor R. It is assumed that the time Δt, expended on a single switching of each capacitor of the storage tube, is sufficiently short by comparison with the time during which u can vary appreciably in value; in other words, the input voltage may be represented, with a sufficient degree of accuracy, in the form of a sequence of rectangular pulses, each of duration Δt.

If the value of the time constant RC is taken sufficiently great by comparison with the time Δt, then a large number of revolutions of the commutator will be required for each of the capacitances of the accumulator to become charged to a voltage close to the input voltage. Under such conditions, the component of the input voltage that is asynchronous with respect to the frequency of rotation f_o of the

commutator can produce but slight fluctuations of the voltage on the capacitors of the accumulator. At the same time, the component of the input voltage that is synchronous with respect to the frequency f_o will fully charge those capacitors after the appropriate number of revolutions of the commutator (i.e., each of the capacitors will be charged up to that instantaneous value of the synchronous component of the input voltage which is taken on by that voltage at the instant of commutation of the given capacitor). As a result of this, the voltage u_i will be separated on the accumulator, and will approach the synchronous component of the input voltage, which, as it were, it "remembers", while on the resistor the voltage u_d will be separated, and will approach the asynchronous component of the input voltage. Accordingly (by analogy to an elementary filter, consisting of resistances and capacitances connected in series), this system of a synchronous filter has two outputs: an integrating output and a differentiating output.

The smaller the value of the coefficient

$$b = \frac{\Delta t}{RC}, \qquad (1)$$

which we shall term the damping coefficient, the closer u_i will be to the synchronous component, and u_d to the asynchronous component, of the input voltage. For practical purposes the cases when $b \ll 1$ are of the greatest interest.

The commutation frequency f_o and the damping coefficient b are the basic parameters of the synchronous filter, completely characterizing its action on the form of the input voltage.

In view of the fact that during the process of commutation each commutated capacitance is able to vary somewhat, the form of the voltage taken from the accumulator will prove to be "sawtooth". In analyzing the processes taking place in this synchronous filter system, we shall not take this "sawtooth" character into account, and shall take, as the true value of the voltage at each commutated contact, the voltage that becomes established at the end of its commutation.

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3. Derivation of the Starting Formula

We shall commence our analysis of the operation of the filter by deriving the starting formula, relating the voltage at its output to the voltage at its input. For this purpose we introduce the following additional notation: t = current time; n = number of complete revolutions of the commutator from time $t = 0$; x = coordinate of relative displacement of commutator (in fractions of a complete revolution), measured from the position occupied by the commutator at time $t = 0$; T_o = period of commutation:

$$T_0 = \frac{1}{f_0}. \tag{2}$$

According to the meaning of the notation just introduced, the current time may be expressed in the form

$$t = (n + x)\, T_0. \tag{3}$$

Let a single voltage pulse, of duration Δt, be fed to the filter input from some source of EMF, and let it fall on one of the capacitors of the accumulator. This capacitor will be charged to the voltage $1 - e^{-b}$, and then, at each successive commutation, it will be discharged by a factor of e^{-b}. As a result, a sequence of pulses will appear at the integrating output of the filter, and the value of these pulses will vary by the law

$$g_i(n) = (1 - e^{-b})\, e^{-nb}. \tag{4}$$

The function $g_i(n)$, represented in Fig.2, may be considered as the reaction of the integrating filter to the unit pulse function (i.e., as its transfer characteristic).

Using eq.(3), the input voltage, which is assigned in the form of the function of time $u(t)$, may be represented in the form of a function of the number of revolutions of the commutator:

$$u(t) = u[(n + x)\, T_0].$$

This permits analysis of the process of variation of the voltage at any capacitor (for any x) as a function of the number of revolutions of the commutator n. Knowing the function $g_i(n)$, the voltage on the capacitor for any n may be found as the sum of voltages, each of which is the result of the action of one of the successive pulses of the input voltage, i.e.,

$$u_i[(n+x)T_0] = u[x\,T_0](1 - e^{-b})\, e^{-nb} + u[(1+x)T_0] \times (1 - e^{-b})\, e^{-(n-1)b} + \ldots + u[(n-1+x)T_0]\,(1 - e^{-b})\, e^{-b} + u[(n+x)T_0](1 - e^{-b}) = (1 - e^{-b}) \sum_{k=0}^{n} u[(n + x - k)\, T_0]\, e^{-kb},$$

where k = a whole number.

The expression so obtained may be rewritten in the form

$$u_i(t) = (1 - e^{-b}) \sum_{k=0}^{n} u(t - kT_0)\, e^{-kb}. \tag{5}$$

This formula, relating the form of the voltage at the integrating output to the form of the voltage at the input, will serve us as our starting point for the solution of all problems, both of steady and transient character.

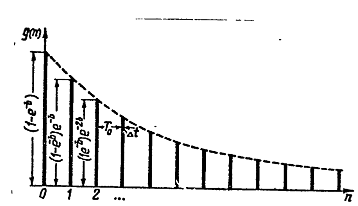

Fig.2

To find the form of the voltage at the differentiating output, the obvious equation may be used:

$$u_d(t) = u(t) - u_i(t), \tag{6}$$

first finding $u_i(t)$ by eq.(5).

Using eq.(5), let us consider the action on the synchronous filter of certain forms of input voltage which, on the one hand, are of the greatest practical interest, and, on the other, most completely disclose the characteristics and properties of the filter. Such forms of input voltage include: a periodic voltage of arbitrary form, a harmonic voltage, and "noise" voltage.

4. Action of Periodic Voltage

Beginning at time $t = 0$, let the periodic voltage $u(t)$ act on the filter input. Let that voltage have a repetition period T, differing little from the commutation period T_o, i.e. let the following inequality hold:

$$\Delta T \ll T, \tag{7}$$

where

$$\Delta T = T_0 - T. \tag{8}$$

If the form of such voltage is observed on an oscillograph with sweep beginning on the left part of the screen and having a repetition period that is a multiple of T_o, then for $\Delta T > 0$, the oscillogram will be seen as moving from right to left, while for $\Delta T < 0$, on the contrary, it will be seen as moving from left to right.

Using eq.(8), we may write

$$u(t - \kappa T_0) = u(t - \kappa \Delta T - \kappa T)$$

or, taking account of the fact that T = period of repetition of the function $u(t)$, while k is an integer, we may write:

$$u(t - \kappa T_0) = u(t - \kappa \Delta T). \tag{9}$$

Thus, according to eq.(5), the expression for the voltage at the integrating output of the synchronous filter in this case may be represented in the form

$$u_i(t) = (1 - e^{-b}) \sum_{\kappa=0}^{n} u(t - \kappa \Delta T)\, e^{-\kappa b}. \tag{10}$$

If the inequality eq.(7) holds, then variations of the quantity k, assuming

only integer values, will lead to relatively small variations of the function $u(t - k\Delta T)$, and this latter function may be considered, with a sufficient degree of accuracy, to be a continuous function of the variable k. Similarly, assuming the following inequality to hold:

$$b \ll 1, \tag{11}$$

the function e^{-kb} may likewise be considered to be a continuous function of the variable k. Here, as the limiting value for k, we may use, instead of n (which assumes only integer values), the continuously varying quantity $n + x$ [cf.eq.(3)].

These assumptions allow us to pass from the formula with the sum, eq.(10), to the formula with the integral:

$$u_i(t) = b \int_0^{n+x} u(t - \kappa \Delta T)\, e^{-\kappa b}\, d\kappa. \tag{12}$$

Further, taking eq.(3) into account, introducing the notation $\tau_o = \frac{\Delta T}{b}$, and passing to a different constant of integration, $\tau = k\Delta T$, we get, from eq.(12):

$$u_i(t) = \frac{1}{\tau_0} \int_0^{\frac{\Delta T}{T_0} t} u(t - \tau)\, e^{-\frac{\tau}{\tau_0}}\, d\tau. \tag{13}$$

In its form, the latter expression recalls the formula for the voltage at the integrating output of an elementary RC-filter, differing from it by the cofactor $\frac{\Delta T}{T_o}$ standing at the upper limit of integration. The sign and value of this cofactor, expressing the relative divergence between the period of the input voltage and the period of commutation, affects the form of the output voltage in the most substantial way.

In the case where the frequency of the input voltage coincides with the commutation frequency (i.e., if $\Delta T = 0$), eq.(13) is easily reduced to the form

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$$u_i(t) = u(t)\left(1 - e^{-\frac{t}{T_0}}\right). \qquad (14)$$

In this case, the form of the output voltage repeats that of the input, but increases exponentially in value, reaching only at the limit, the same value as the input voltage.

In the most general case, when $\Delta T \neq 0$, the output voltage is found to differ from the input voltage both in form and in magnitude, which vary at first, but then, with time, reach certain limiting values.

Let us obtain an expression for the limiting value of the output voltage. For this purpose, let us first represent eq.(13) as:

$$u_i(t) = \frac{1}{\tau_0}\int_0^t u(t-\tau)\, e^{-\frac{\tau}{\tau_0}}\, d\tau - \frac{1}{\tau_0}\int_{\frac{\Delta T}{T_0}t}^{t} u(t-\tau)\, e^{-\frac{\tau}{\tau_0}}\, d\tau.$$

It will be easily seen that for sufficiently great values of t (in the steady state) when the inequality

$$\frac{\Delta T}{T_0} t \gg \tau_0,$$

holds, the second number of the last inequality may be neglected in view of its smallness, and we may consider that

$$u_i(t) = \frac{1}{\tau_0}\int_0^t u(t-\tau)\, e^{-\frac{\tau}{\tau_0}}\, d\tau. \qquad (15)$$

This expression already agrees completely with the formula for the voltage at the integrating output of an elementary RC-filter with the time constant:

$$\tau_0 = \frac{\Delta T}{\Delta t} RC, \qquad (16)$$

Consequently, under appropriate conditions, the voltage at the output of the synchronous filter may coincide with the voltage at the integrating output of the equivalent RC-filter. For negative values of ΔT, however, this analogy loses its previous meaning, for τ_0 and τ become negative, and when the operation of reversion, expressed by eq.(15), is performed, we must summate the residual action on the system by the following values of the function u(t), and not by its preceding values. In this case it is as though we were dealing with the action of the input voltage u(-t) on the equivalent RC-filter instead of the action of the actual input voltage u(t).

Figure 3 gives examples of the steady values of the output voltages of the filter for a pulsed input voltage, obtained by eqs.(15) and (16) for the cases $\Delta T > 0$ and $\Delta T < 0$.

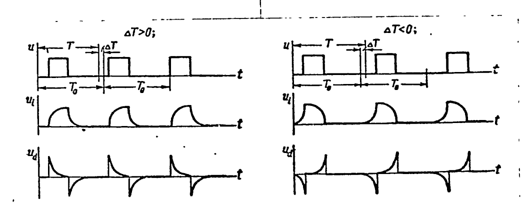

Fig.3

When these output voltages are observed on an oscillograph with a scanning having a repetition period that is a multiple of T_0, the "pulling" of their form in the direction opposite the apparent displacement along the screen may be observed.

5. The Action of Harmonic Voltage

Let the voltage

$$u(t) = U_m \sin \omega t, \qquad (17)$$

be imposed on the filter input. Consequently:

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$$u(t-\kappa T_0) = U_m \sin(\omega t - \kappa \omega T_0).$$

On introducing the notation, to shorten the formulas:

$$a = \omega T_0 \qquad (18)$$

and applying eq.(5), we obtain

$$u_i(t) = (1 - e^{-b}) U_m \sum_{\kappa=0}^{n} \sin(\omega t - a\kappa) e^{-\kappa b}.$$

We shall concern ourselves only with the steady process, and shall therefore perform the summation to k = ∞:

$$u_i(t) = (1 - e^{-b}) U_m \sum_{\kappa=0}^{\infty} \sin(\omega t - a\kappa) e^{-\kappa b}. \qquad (19)$$

It can be shown that

$$\sin(\omega t - a\kappa) e^{-\kappa b} = \frac{\sqrt{a^2+b^2}}{\sqrt{2(\operatorname{ch} b - \cos a)}} \times$$

$$\times \int_{\kappa-\frac{1}{2}}^{\kappa+\frac{1}{2}} \sin\left(\omega t - ay + \operatorname{arctg}\frac{a}{b} + \frac{a}{2} + \arccos\frac{(e^{-b} - \cos a) e^{\frac{b}{2}}}{\sqrt{2(\operatorname{ch} b - \cos a)}}\right) e^{-by}\, dy.$$

Consequently, eq.(19) may be written as follows:

$$u_i(t) = \frac{(1 - e^{-b})\sqrt{a^2+b^2}}{\sqrt{2(\operatorname{ch} b - \cos a)}} U_m \int_{-\frac{1}{2}}^{\infty} \sin\left(\omega t - ay + \operatorname{arctg}\frac{a}{b} + \frac{a}{2} + \right.$$

$$\left. + \arccos\frac{(e^{-b} - \cos a) e^{\frac{b}{2}}}{\sqrt{2(\operatorname{ch} b - \cos a)}}\right) e^{-by}\, dy. \qquad (20)$$

On integrating the expression so obtained and performing the appropriate transformations, we get

$$u_i(t) = \frac{2 \operatorname{sh}\frac{b}{2}}{\sqrt{2(\operatorname{ch} b - \cos a)}} U_m \sin(\omega t + \varphi), \qquad (21)$$

where

$$\varphi = \pi + a + \arccos\frac{(e^{-b} - \cos a) e^{\frac{b}{2}}}{\sqrt{2(\operatorname{ch} b - \cos a)}}$$

or, after obvious transformations,

$$\varphi = \arccos\frac{(e^{b} - \cos a) e^{-\frac{b}{2}}}{\sqrt{2(\operatorname{ch} b - \cos a)}}. \qquad (22)$$

It follows from a comparison of eqs.(17) and (21) that the voltage transmission coefficient for the integrating output is equal, in absolute value, to

$$K_i = \frac{u_i}{u} = \frac{2 \operatorname{sh}\frac{b}{2}}{\sqrt{2(\operatorname{ch} b - \cos a)}}, \qquad (23)$$

and its phase angle to

$$\varphi_i = \arccos\frac{(e^{b} - \cos a) e^{-\frac{b}{2}}}{\sqrt{2(\operatorname{ch} b - \cos a)}}. \qquad (24)$$

Knowing the absolute value and the phase angle of the transmission coefficient of the integrating output, its complex value is easily obtained:

$$\dot{K}_i = \frac{\dot{u}_i}{\dot{u}} \qquad (25)$$

After performing the appropriate trigonometric transformations, we get

$$\dot{K}_i = \frac{e^{-\frac{b}{2}} \operatorname{sh}\frac{b}{2}}{\operatorname{ch} b - \cos a}\left[e^{b} - \cos a - i \sin a\right]. \qquad (26)$$

Thence follows directly a more convenient expression for φ_i:

$$\varphi_i = -\operatorname{arctg}\frac{\sin a}{e^{b} - \cos a}. \qquad (27)$$

Using eq.(26), it is easy to obtain an expression for the complex transmission

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coefficient of the differentiating output:

$$\dot{K}_d = \frac{\dot{u}_d}{\dot{u}}. \qquad (28)$$

Indeed, since

$$\dot{u}_d = \dot{u} - \dot{u}_i,$$

then

$$\dot{K}_d = 1 - \dot{K}_i.$$

On performing the appropriate transformations, we obtain

$$\dot{K}_d = \frac{e^{-\frac{b}{2}} \operatorname{sh}\frac{b}{2}}{\operatorname{ch} b - \cos a}\left[(1-\cos a)\operatorname{cth}\frac{b}{2} - i\sin a\right]. \qquad (29)$$

The absolute value of the transmission coefficient of the differentiating output is:

$$K_d = \frac{2e^{-\frac{b}{2}}\sin\frac{a}{2}}{\sqrt{2(\operatorname{ch} b - \cos a)}}, \qquad (30)$$

and its phase angle is:

$$\varphi_d = \operatorname{arctg}\frac{\sin a}{(1-\cos a)\operatorname{cth}\frac{b}{2}}. \qquad (31)$$

Equations (23) and (27) express respectively the amplitude-frequency and phase-frequency characteristics of the integrating filter, while eqs.(30) and (31) express those respective characteristics of the differentiating filter. Examples of such characteristics, constructed for b = 0.1, are given in Fig.4.

For cases where $b \ll 1$, the formulas so obtained are considerably simplified. Thus, for example, eq.(23) is reduced to the form

$$K_i = \frac{1}{\sqrt{1+\left(\frac{2\sin\frac{a}{2}}{b}\right)^2}} \qquad (32)$$

and eq.(30) to the form

$$K_d = \frac{1}{\sqrt{1+\left(\frac{b}{2\sin\frac{a}{2}}\right)^2}}. \qquad (33)$$

For cases when the frequency of the input voltage is close to the commutation frequency or to one of its harmonics (the case of "slight detuning"), these formulas may be still further simplified.

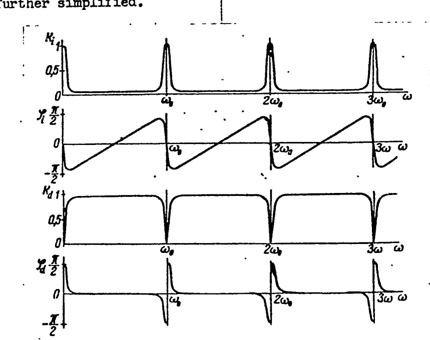

Fig.4

According to the notation of eq.(18), we may write

$$a = 2\pi\left(m + \frac{\Delta f}{f_0}\right), \qquad (34)$$

where m is zero or an integer, while $\frac{\Delta f}{f_0}$ is the relative "detuning" of the frequency of the input voltage.

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Putting $\frac{\Delta f}{f_o} \ll 1$ in this case, it may be considered that $2 \sin \frac{a}{2} \approx 2\pi \frac{\Delta f}{f_o}$.

Consequently,

$$K_i \approx \frac{1}{\sqrt{1+\left(\frac{\pi}{b} \frac{2\Delta f}{f_0}\right)^2}}, \tag{35}$$

$$K_d \approx \frac{1}{\sqrt{1+\left(\frac{b}{\pi} \frac{f_0}{2\Delta f}\right)^2}}. \tag{36}$$

Under the same conditions,

$$\varphi_i = -\operatorname{arctg}\left(\frac{\pi}{b} \frac{2\Delta f}{f_0}\right) \tag{37}$$

and

$$\varphi_d \approx \operatorname{arctg}\left(\frac{b}{\pi} \frac{f_0}{2\Delta f}\right). \tag{38}$$

It is interesting to note that for cases of "slight detuning", the amplitude-frequency and phase-frequency characteristics of the synchronous filter coincide in form with the corresponding characteristics of a resonant filter consisting of a single oscillatory circuit (Fig.5), provided that circuit is tuned to the frequency f_o and has a logarithmic damping decrement ϑ which is numerically equal to the damping factor b.

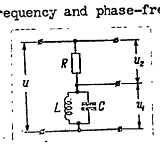

Fig.5

By analogy to the ordinary resonant systems, the concept of the pass band may be introduced for the synchronous filter. If a pass band is understood to mean a band of frequencies 2 F, included within the limits of the half-slope of the square of the amplitude-frequency characteristic (for one of the peaks of this characteristic), then from eq.(35) we get the following expression for such pass band:

$$2F = \frac{b}{\pi} f_0, \tag{39}$$

analogous to the well-known expression for the pass band of a single oscillatory circuit:

$$2F = \frac{\vartheta}{\pi} f_0. \tag{40}$$

6. The Action of Noise Voltage

On the filter input let there be impressed the noise voltage u(t) characterized by the dispersion D(u) and the mean square value σ, where

$$\sigma = \sqrt{D(u)}. \tag{41}$$

As a result, at the integrating output of the filter, there appears the voltage $u_i(t)$, characterized by the dispersion $D(u_i)$ and the mean square value σ_i, where

$$\sigma_i = \sqrt{D(u_i)}. \tag{42}$$

The ratio

$$G = \frac{\sigma}{\sigma_i}, \tag{43}$$

characterizing the power of the integrating filter to attenuate the noise voltage, will be termed the filtration factor.

Let us establish the relation between the filter parameters and the filtration factor. For this purpose let us make use of eq.(5), rewritten for the case of the steady state, assuming that $b \ll 1$:

$$u_i(t) = b \sum_{\kappa=0}^{\infty} u(t - \kappa T_0)\, e^{-\kappa b}.$$

Let us determine the dispersion of the left and right sides of this expression:

$$D(u_i) = b^2 D(u) \sum_{\kappa=0}^{\infty} e^{-2\kappa b}.$$

The holding of the inequality $b \ll 1$ enables us to pass from the sum to the integral:

$$D(u_i) = b^2 D(u) \int_0^{\infty} e^{-2\kappa b}\, d\kappa,$$

whose solution gives

$$D(u_i) = \frac{b}{2} D(u).$$

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From the relation so obtained, on the basis of the definition in eq.(43), we obtain

$$G=\sqrt{\frac{2}{b}}. \qquad (44)$$

Making use of the fact that the noise voltage has a uniform distribution of energy throughout the spectrum of frequencies, the filtration factor may likewise be determined from the amplitude-frequency characteristic of the filter.

Considering the spectrum of noise within the limits of repetition of the form of the amplitude-frequency characteristic (i.e., within the limits of variation of a from 0 to π), it is easy to reach the conclusion that the ratio of the noise energy at the filter input to the noise energy at its output will be expressed as

$$\frac{\pi}{\int\limits_0^{\pi} K_t^2\, da}.$$

The filtration factor, by definition, must equal the square root of this ratio. Making use of eq.(32) and taking account of the fact that b ≪ 1, we obtain from this the previous result of eq.(44).

7. Certain Properties of the Accumulator

Let us attempt to ascertain the equivalent impedance $\dot{Z}$ by which the capacitative accumulator in the synchronous filter system may be replaced.

When a sinusoidal voltage is applied to the filter input through the accumulator, in addition to a sinusoidal current of the same frequency, a sawtooth component of the charge-and-discharge current of its individual capacitors will also flow. When the resistance R in series with the accumulator increases, this component will decrease, and at the limit, as R →∞, it will disappear. For this reason the equivalent impedance Z of the accumulator (which has a meaning only for the sinusoidal current) must be sought at its connection to the filter circuit with an infinitely great resistance.

Setting up the equations relating the voltages $\dot{u}$ and $\dot{u}_d$ through the resistances $\dot{Z}$ and R, we get

$$\dot{Z}=R\left(\frac{\dot{K}_d-1}{\dot{K}_d}\right).$$

Hence, making use of eq.(29), interpreting the values of b and a in accordance with eqs.(1) and (18), and passing to the limit R → ∞, we finally obtain

$$\dot{Z}=-i\frac{\Delta t}{2C}\operatorname{ctg}\frac{\omega T_0}{2}. \qquad (45)$$

It is clear from this formula that the equivalent resistance of the accumulator coincides with the input resistance of a lossless open-circuit line of length

$$l=\frac{vT_0}{2} \qquad (46)$$

(where v = velocity of propagation of voltage wave along the line), and of wave impedance equal to

$$W=\frac{\Delta t}{2C}. \qquad (47)$$

Thus, by completely formal operations, we have discovered the possibility of using an open-circuit line as a wave accumulator for a synchronous filter. This result, however, is entirely natural, since the line constitutes a peculiar "memory" device.

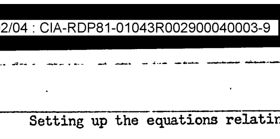

Fig.6

a) Line; b) Differentiating Output; c) Integrating Output

Figure 6 gives a diagram of the synchronous filter with line (i.e. with a wave accumulator). The application of a single voltage pulse from some source of EMF at the filter input causes the appearance of this voltage at the input of the line, followed by its wavelike motion along the line, reflection from the end of the line, motion in the opposite direction, reappearance at the input of the line, partial reflection

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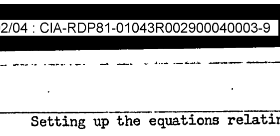

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from the input of the line, new motion along the line, and so on. As a result, there appears, at the input of the line, a sequence of voltage pulses, separated by time intervals $\frac{2l}{v}$ and damped by an exponential law (even in the case of a lossless line) owing to the losses of energy across the resistance R at each reflection from the input. This sequence of pulses, constituting the transfer characteristic for the integrating output of the synchronous filter system presented, coincides with the analogous transfer characteristic of the synchronous filter with capacitative accumulator (Fig.2). Consequently the two systems mentioned are equivalent in their operation.

It must be noted that if the voltage separated is of interest merely for a certain part of its repetition period, then the requirements to be met by the "volume of memory" of the accumulator are lowered by the corresponding factor. In this case the capacitative accumulator may be simplified on account of the corresponding decrease in the number of the capacitances to be switched. A line accumulator cannot be further simplified under these conditions, however, since the line length required depends only on the duration of the repetition period of the voltage separated.

In principle, any other device possessing the property of "remembering" the form of the oscillations fed to it may likewise be used as the accumulator of a synchronous filter.

Paper received by Editors 26 May 1955

BIBLIOGRAPHY

1. Jensen,A.S., Smith,J.P., Wesner,M.H., and Flory,L.E. - Barrier-Grid Storage Tube and Its Operation. RCA Review 9,3,1948

2. Harrington,J.V. and Rogers,T.E. - Signal-to-Noise Improvement through Integration in a Storage Tube. Proc.Inst.Radio Engrs. 38,1197,1950

3. Vavilov,V.S. - Experiments on Radiolocation of the Moon. Usp.fiz.nauk 39,359,1949

4. Vladimirskiy,K.V. - On the Synchronous Filter. Zhur.eksp.teor.fiz. 21,3,1951

5. Beard,C.I. and Skomal,E.N. - RC Memory Commutator for Signal-to-Noise Improvement. Rev.of Sci.Instr. 24,276,1953

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REVIEW BY AUTHOR

THE USE OF HARMONICS IN CALCULATING ENVELOPES

by

A.A.Kulikovskiy,

Full Member of the Society

In considering the nonlinear problems of pulse engineering one often encounters the influence on an oscillatory circuit (and consequently on the linear resonant stages following it) of a high-frequency current pulse in which the form of the "filling" oscillations differs sharply from harmonic, owing to the nonlinear state of the preceding tube. The operation of the oscillatory circuit in the plate circuit of a nonlinear resonant amplifier will serve as a typical example. It is of interest to determine, for this case, the variation with time of the voltage on the oscillatory circuit or at the output of the following linear stages. As far as I know, this problem has not yet been investigated. A paper discussing the distortions of telegraph signals in the power stage of a transmitter operating in class B (i.e., with a constant angle of cut-off and with constant first-harmonic factor) does approach this subject (Bibl.1).

Let us make use of the transfer characteristic $h(t) = \dfrac{u_{outp}(t)}{U_{inp} \cdot 1(t)}$ of a resonant amplifier containing n + 1 linear stages (Bibl.2,3). If we divide it by the transconductance S of the first tube, we get an expression for the contact resistance $r(t) = \dfrac{u_{outp}(t)}{I_{inp} \cdot 1(t)}$ of a quadripole consisting of an oscillatory circuit in the plate circuit of the first tube and n linear stages following it. This resistance may be represented in the form $r(t) = R(t) \sin \omega_o t$, where ω_o = resonant angular velocity of the quadripole in question; $R(t)$ = the slowly varying amplitude. Let us determine the voltage at the output of the quadripole, with an arbitrary current $i(t)$

in the plate circuit of the first tube, by means of the Duhamel integral:

$$u_{out}(t) = r(0) \cdot i(t) + \int_0^t r'(t-x)\, i(x)\, dx. \qquad (1)$$

Taking the slow variation of R(t) into account, and using methods similar to those employed in another paper of mine (Bibl.4), it may be shown that this expression leads to an integral relation of complex amplitudes:

$$\dot{U}_{m\ out}(t) = \frac{\omega_0}{2} \int_0^t \dot{R}(t-x)\, \dot{I}_{m1}(x)\, dx. \qquad (2)$$

Here $\dot{I}_{m1}(x)$ is the slowly varying complex envelope of the first harmonic of the expansion of i(t) into Fourier series in the "sliding" interval $(x;\ x + T_o)$, where $T_o = \frac{2\pi}{\omega_o}$ = period of resonant frequency of the quadripole.

If the fundamental frequency of the current i(t) coincides with the resonant frequency of the quadripole, then the phase of its first harmonic remains constant. In that case, eq.(2) relates real envelopes.

In the special case when the current i(t) consists of a pulse filled with harmonic oscillations of frequency coinciding with the resonant frequency of the quadripole $i(t) = I_m(t) \sin \omega_o t$, we get $I_{m1}(t) = I_m(t)$. Then eq.(2) coincides with the well-known integral relation obtained by S.I.Yevtyanov for this important special case (Bibl.3).

In order to use the relation so obtained, eq.(2), for calculating the output voltage of the nonlinear stage at known input voltage and at an assigned nonlinear tube characteristic, we must determine the plate current i(t) and then find the envelope $I_{m1}(t)$ of its first harmonic on expansion into a Fourier series in the interval $(x;\ x + T_o)$. If the plate reaction can be neglected, which is permissible for pulse stages with pentodes or tetrodes, then there will be no difficulty in determining the plate current from the known input voltage and from the assigned nonlinear tube characteristic. In the case of triodes, the investigation is impeded by the necessity of allowing for the plate reaction, but even in this case eq.(2) may serve as one of the equations of the operating conditions of the tube.

In calculating the envelope of the first harmonic of the plate current of a tube operating under nonlinear conditions, it is convenient to approximate the relation between the first harmonic and the ratio of the DC bias voltage U_o to the amplitude of the alternating voltage of the input signal U_m. Thus, with the nonlin-

Fig.1

Fig.2

earity shown in Fig.1, the amplitude of the first harmonic of the current is $I_{m1} = SU_mA$, where $A = \frac{1}{\pi}(\psi - \sin\psi\cdot\cos\psi)$; S=transconductance of the ascending characteristic; $\psi = \cos^{-1}\frac{U_o}{U_m}$ = angle of cut-off. The relation $A\left(\frac{U_o}{U_m}\right)$ is represented on Fig.2 and is rather well approximated by the broken line, whose equation is

$$A = 0 \quad \text{at} \quad \frac{U_o}{U_m} > 1,$$

$$A = \frac{1}{2}\left(1 - \frac{U_o}{U_m}\right) \quad \text{at} \quad 1 > \frac{U_o}{U_m} > -1,$$

$$A = 1 \quad \text{at} \quad \frac{U_o}{U_m} < -1.$$

In accordance with this approximation, for an assigned envelope of input voltage $U_m(t)$, the interval of integration of eq.(2) must be divided into parts such that, within the limits of each part, $\frac{U_o}{U_m}$ shall remain either greater than 1, or less than 1, or have intermediate values. In the former case, A = 1, which corresponds to the linear state, i.e., to operation without cut-off. In the second case, the tube is blocked (A = 0), and the particular integral vanishes. On those parts

of the interval of integration where $\frac{U_o}{U_m}$ varies within the limits (-1; +1) the process is complicated by the variation of the angle of cut-off and of the value of the first harmonic of current with the variation of $U_m(t)$. The integral of eq.(2) in this region takes the form:

$$\frac{S\omega_0}{2}\int R(t-x)\,U_m(x)\left[\frac{1}{2}\left(1-\frac{U_o}{U_m(x)}\right)\right]dx =$$

$$= \frac{S\omega_0}{4}\int R(t-x)\,U_m(x)\,dx - \frac{S\omega_0}{4}U_o\int R(t-x)\,dx.$$

The first summand characterizes the linear process of amplitude build-up, but contains the correction factor 1/2, while the second summand characterizes the process independent of the envelope of the input signal.

It also follows from the above that, when pulse signals pass through a frequency changer, the distortion of the envelope obeys the same regularities as when they pass through a resonant amplifier with the same oscillatory circuit in the plate circuit as the frequency changer has, provided that the usual requirement, that the envelope of the intermediate frequency current shall coincide with the envelope of the input signal, is satisfied.

Paper received by Editors 4 May 1953.

Review by Author of paper 3 March 1956

BIBLIOGRAPHY

1. Gonorovskiy,I.S. - Contribution to the Theory of Distortions of Telegraph Signals in the High-Frequency Parts of Radio Equipment. IEST No.6,1940

2. Ageyev,A., Kobzarev,Yu. - Transient Phenomena in a Resonant Amplifier. Zhur.Tekh. Fiz. 5,8,1935

3. Yevtyanov,S.I. - Transient Processes in Receiver-Amplifier Circuits. Svyaz'zdat 1948

4. Kulikovskiy,A.A. - Build-Up Processes in Detection of Pulse Signals. This

Journal, No.6,1955

5. Kulikovskiy,A.A. – Passage of Pulse Signals through Nonlinear Selective Stages of Radio Receiving Equipment. Trudy VVIA im.Zhukovskiy. No.328,1950

6. Kulikovskiy,A.A. – Steady and Transient States of Nonlinear Stages of Radio Receiving Equipment. Ibid. No.544,1955

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(continued)

ELECTRONICS SECTION

L.A.Vaynshteyn, in his paper Electron Waves in Delay Systems, discussed the linear theory of such waves (theory of amplification of weak signals in the travel traveling-wave tube); the results of generalization of studies in this field allow derivation of an expression for the forces acting on an electron beam in a nonlinear state of amplification of a monochromatic signal.

Z.S.Chernov reported the results of a study of spiratrons, which are new electronic devices of the type of the traveling-wave tube (LBV) with centrifugal-electrostatic focusing of the electron streams.

The Section noted that the theoretical and experimental material contained in the paper evidence the promise inherent in the utilization of the new type of focusing for the construction of spiratrons; it also recommended accelerating the introduction of spiratrons, in the first place, in the equipment of radio relay lines.

Ye.D.Naumenko communicated the results of the development of laboratory models of reflecting klystrons for measuring purposes. The 7–19 mm range is covered by klystrons of two sub-types with a power of the order of 5 – 10 mw.

A paper by A.S.Tager presented the results of a theoretical study of the effect of noise waves in the electron flux on the noise characteristics of traveling wave tubes.

The Section recommended the continuation of the work on the theoretical and experimental study of noise in electron fluxes at super-high frequencies.

In his paper, Nonlinear Theory of the Traveling Wave Tube, L.A.Vaynshteyn formulated the fundamental equations of the nonlinear theory of the traveling-wave tube, operating in the state of amplification of a monochromatic signal, and simplified equations for "slowly varying" functions, which hold at small values of the fundamental parameter of smallness of the tube.

The Section noted the considerable interest of the nonlinear theory of the traveling-wave tube reported by the speaker.

A.M.Chernushenko spoke on an oscillograph for studying the electrical processes in pulse generators of the centimeter-wave band. The oscillograph has a marker generator and calibrated delay lines. The Section noted that super-high frequency cathode-ray tubes should find widespread application in super-high frequency engineering, and pointed out the necessity for their extremely rapid introduction into measurement practice.

The paper by S.G.Afanasova investigated the reactive properties of diode intermediates by introducing them into a coaxial passive oscillatory circuit with external exciter. She considered the possibility of using diode intermediates for the electronic returning of the frequency of HF generators.

In the resolutions on the paper the Section noted that the work on the methods of frequency control (modulation) of superhigh-frequency generators was of great practical interest, and recommended its continuation.

V.S.Troitskiy, A.M.Starodubtsev, and V.S.Serebrennikov spoke on the results of a study of the phase fluctuations of the oscillations of certain super-high frequency generators.

The Section noted the timeliness of the investigation of the fluctuations in the phase and spectral composition of the oscillations of super-high frequency generators, and recommended the expansion of the work in this field.

M.I.Kuznetsov reported on the design, and presented the characteristics, of super-high frequency generators with electronic tuning, in the centimeter-wave band.

SECTION OF RECEIVING AND TRANSMITTING EQUIPMENT

A paper by V.A.Klyaznik discussed a method of compensating the injurious action

of pulse noise in the radio reception area, both after and before the detector stage, and also pointed out methods of practically complete suppression of pulse noise by compensation methods.

M.G.Golubtsov and V.A.Morozov devoted their report to the suppression of pulse noise by the compensation method, with the formation of artificial noise. They showed the possibility, in principle, of total compensation of pulse noise, without distortion of the signal received.

L.T.Remizov and L.S.Tyufyakin devoted their paper to the suppression of pulse noise by the compensation method, using frequency conversion, and showed that the suppression of noise in the IF amplifier tract requires two additional channels besides the reception channel (i.e. that the total band of the receiver must be widened).

The Section noted the considerable scientific interest and practical value of the studies in the field of compensation methods of pulse noise suppression.

A paper by A.A.Gorbachev gave the essential features of a method of controlling pulse noise by limiting and converting the spectrum of the useful signal and of the noise. He gave the results of an experimental check of the method, and discussed the circuit diagram of the equipment. The Section recommended the realization of this method in broadcast receivers.

A.A.Pirogov, in his paper Balanced Plate Modulation, considered a number of questions connected with the problem of increasing the efficiency of radio transmitters, by improving their noise-proof qualities, by reducing their power consumption, and by narrowing the frequency band. The Section noted that the proposed system was one of the most progressive and promising methods of radio transmission.

Yu.V.Bogoslovskiy reported on possible methods of accomplishing auto-plate modulation with triodes and pentodes, and analyzed the technique of calculating the modulation characteristics of a stage. The Section recommended the widespread use of the materials presented, and advised further development of the methods of calculating auto-plate modulation.

In his paper, B.I.Rassadin discussed the experimentally confirmed advantages of signal transmission on a single frequency band in multi-channel radiotelegraph and radiotelephone systems. He recommended a method for considerably reducing nonlinear distortions.

The Section noted the original formulation of the question and the value of the technical improvement of the single-band transmitter, allowing a considerable increase in its efficiency.

V.V.Malanov described the essential features of a high-efficiency pulse method of power amplification of audiofrequency oscillations, its advantages, and its qualitative indexes.

The Section noted the great scientific and practical interest of the method so developed, and recommended continued research in this field.

D.V.Ageyev reported on his theory of the FM receiver with follow-up tuning and controlled oscillatory circuit. He defined the conditions under which an FM receiver with follow-up tuning will have greater noise-immunity than the ordinary FM receiver, without introducing appreciable frequency or nonlinear distortions of the received signal.

Ya.G.Rodionov gave the results of an experimental study of FM receivers with controlled resonant frequency, which had the object of elucidating the advantages of this receiver over the ordinary wide-band FM receivers. The results of the experiments were compared with the conclusions of the theory developed by D.V.Ageyev.

The Section noted the great scientific and practical value of Agayev's studies and the interest of Rodionov's experimental work.

M.K.Belkin gave an analysis of the superregenerative receiver operating in the super-high frequency range with a reflecting klystron. The Section recommended continued study of the methods of reducing klystron noise and the noise factor of the receiver.

TELEVISION SYSTEM

A.P.Angaforov considered two basic principles of construction, together with the design of television receiving tubes for direct reception of colored images: the three-ray tube with dim-out mask and mosaic screen (of the colortron type), and the single-ray tube with control grid and line screen (of the chromatron type), which may be applied to color television and radar.

M.E.Gos discussed various methods of narrowing the frequency band in modern systems of color television, based mainly on reducing the quantity of color information to be transmitted.

In its resolutions on these papers, the Section pointed out the necessity of considerably expanding the work on creating a system of color television.

P.Ye.Kodes told of a new television system using eight studio channels, designed to transmit one program in black-and-white television. The television stations of large USSR cities will be equipped with these installations.

I.V.Ostrovskiy considered the fundamental principles of construction of the new Moscow and Leningrad television centers, as well as schematic diagrams of the equipment.

The Section recommended paying particular attention, in developing the equipment for the new television centers at Moscow and Leningrad, to the qualitative and operational indices of that equipment. An interdepartmental conference should be called in the near future to deal with the questions of selecting methods of television transmission of motion pictures and distributing the research and experimental-designing work among the organizations concerned.

Yu.V.Lobov proposed a more objective approach to the evaluation and measurement of the qualitative indices of television transmitters. The measurements are made by means of a special electronic measuring instrument giving readings of the amount of distortion directly on the scale of the instrument itself.

V.S.Gladin reported the results of a study of two qualitative indices of television receiving tubes: contrast and definition of the television image. The approach to the evaluation of these indices had not previously been uniform. The speaker proposed passing from these qualitative indices to three others: contrast of major details, contrast of medium details, and contrast of minor details. The introduction of a characterization of the transmission of detail in the evaluation of the quality of the television image will open up great opportunities in the development of television receiving tubes.

The report by A.P.Yefimov gave a critical analysis of the existing technique of evaluating the influence of periodic noise, and pointed out that the disturbing effect of various noises must be evaluated according to their visible manifestation on the television image.

The Section noted the value of the reported methods of measuring television parameters and allowances, and recommended continuation of this work. Considerable expansion and intensification is required in the work of improving television receivers, and in that of determining the optimum parameters for the entire television tract.

V.B.Ivanov analyzed the specific distortions that arise on the reception of television with partially suppressed lateral distortions alone, and especially the distortions due to the square component in the system adopted as the USSR television standard.

G.V.Braude and M.A.Ushakov discussed the planning of pre-amplifiers for television motion-picture channels, and the basic requirements they must meet.

Considering one of the most important problems of television broadcasting to be the further improvement of reception quality, the Section recommended further work in determining and standardizing the optimum characteristics of the amplifier system of television receivers.

In his paper, I.I.Tsukkerman gave the principal results of a study of the

electron-optical method of varying the angle of "vision" of the television camera, as applied to the super-orthicon.

The Section approved the advisability of using matched magnetic and electric fields to control electron-optical aberrations, both in the construction of television cameras with a variable field of vision and to enhance the dynamic range of super-orthicons.

A.Ya.Breitbart reported on the principal data and characteristics of the new types of televisors scheduled for initial production in 1956. The trends of future development of television reception engineering were discussed. The Section noted the necessity for increasing the amount of research in that field.

Yu.I.Kaznacheyev, in his paper Prospects of the Application of Waveguides in Communications and Television, discussed the advantages of long-distance signal transmission on H_{01} waves along circular-section waveguides. The Section noted the great importance of the investigations conducted in the USSR on the questions of employing waveguide communications lines, and recommended the formulation of a single plan for such work in the research organizations of the various government departments.

Ye.V.Gershzon considered the possibility and advisability of using semiconductor devices in televisors. He remarked that a number of televisor units could already be built with the semiconductor devices produced in the USSR.

In its resolutions, the Section recommended acceleration of the mass production of the corresponding types of semiconductor devices.

A.D.Azat'yan presented the technical characteristics of a number of tube types in wide use in modern Western European and American radio receivers and televisors, and gave a survey of the new USSR bantam tubes for radio receivers and televisors.

The Section recommended considerable expansion of the research work in the design of new tube types, with special attention to the noise characteristics of the tubes designed for the input equipment.

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RADIO BROADCASTING AND ACOUSTICS SECTION

I.Ye.Goron delivered a report on the draft standard for the principal qualitative characteristics of radio broadcasting channels. He discussed the results and technique of investigating the audibility of various distortions, and presented a technique for working up the results of such investigations.

He presented a justification of the classification adopted, recommended the draft standards for pass band, nonlinear distortions, principal forms of noise, and gave a justification of these standards.

The paper by G.S.Genzel' was devoted to a study of the audibility of crosstalk. He discussed questions of standardization, and gave an analysis of the factors affecting the results of subjective-statistical expert tests of the auditory perception of such noise.

The Section recommended that measures be taken for the most rapid introduction of the draft standards so worked out, and that persons preparing specifications for new equipment, or for the rebuilding of existing equipment, guide themselves by such standards.

L.M.Zayezdniy proposed criteria of classification of methods of measurement and identification of nonlinear distortions, and presented a survey of the existing methods.

In the paper by N.L.Bezladnov, the shortcomings of the methods now in use for measuring nonlinear distortions were discussed. The speaker also presented requirements which, if met, would permit attainment of the necessary correspondence between the results of measurements and the perception of distortions.

The Section's resolutions on these papers noted the great timeliness and the necessity for working out modern methods of measuring nonlinear distortions in radio broadcasting channels.

M.V.Laufer considered the circuit of the recording magnetometer with modulation

magnetic head and presented the design data of an instrument for examining magnetic phonograms and the magnetic fields of recording heads. The Section noted the originality and value of the work done by the staff under the direction of the speaker.

The report by V.A.Nyurenberg discussed the application of a system of remote measurement to radio broadcasting. The system developed encourages the improvement of the quality of the operation of radio broadcasting installations.

In its resolutions the Section noted the timeliness of the work and recommended the extension of this system of remote measurements to other means of communication and radio broadcasting.

A.K.Bektabegov told of a new piezoelectric ceramic pickup with a number of advantages.

The Section recommended the international standardization of the piezoelectric ceramic heads of phono pickups so as to make the heads and their spare parts interchangeable. The phono pickups now in production, using piezoelectric elements of rochelle salt or ammonium phosphate, do not meet modern demands, and should be replaced by piezoelectric ceramic phono pickups.

V.S.Kissel'gof presented a survey of the electrical and design features of modern USSR and foreign radio broadcast receivers and radio consoles of various classes. He formulated the problems of development of radio broadcast equipment during the next few years.

In his report, P.Ye.Shifman discussed the possibilities of a sharp improvement in the quality of the sound of modern radio broadcast receivers, which is of particular importance in connection with the introduction of ultrashort-wave FM broadcasting.

In its resolutions on these reports, the Section noted that the new models of unified radio broadcast receivers and radio consoles recently developed are a substantial step forward in the field of radio broadcast receiving apparatus. The Section recommended acceleration of the creation of various types of radio receivers, extensive use of semiconductor devices and printed circuits in them, and development of high-grade speakers and sound systems for the new models of radio and television receivers. In drafting a new State Standard it will be advisable to provide for the possible production of radio receivers in the 13-19 meter band.

SECTION FOR SEMICONDUCTOR DEVICES

A.I.Stefanovskiy surveyed the semiconductor devices produced in 1955 and analyzed their shortcomings. He reported the studies on improving the electrical parameters and other characteristics of semiconductor devices and presented the results.

The resolutions of the Section noted that a comparison of the data given in the report with the data on the state of semiconductor technology abroad indicates the unfavorable position in the design and series production of USSR semiconductor devices.

V.B.Pestryakov presented data on the actual effectiveness of the use of semiconductors in radio equipment of the receiving type: diminished weight and power consumption, improvement of the thermal conditions, prolonged service life.

G.Z.Gol'dshteyn discussed the specific features of semiconductor amplifiers in automatic regulation systems.

Yu.D.Lindenbraten, on the basis of a criterion of evaluation for various semiconductor amplifier circuits operating over a wide temperature range, discussed the conditions of stabilization of the amplifier parameters and the methods of stabilizing the DC conditions.

The Section noted the practical significance of the materials of these papers.

On the basis of a representation of a transistor in the form of an active, irreversible quadripole, S.G.Kalikhman stated methods of experimental determination of the characteristic parameters of the junction transistors used in the radio-frequency range.

Yu.A.Kamenetskiy discussed the qualitative indices of transistors and an in-

strument for measuring them, based on the determination of the parameters of an equivalent circuit.

The report by Ya.K.Trokhimenko was devoted to the use of feedback in transistor amplifiers.

The Section noted the importance of the use of feedback in transistor circuits.

A.A.Rizkin showed in his paper that by introducing a regeneration factor it is easy to obtain simple formulas for calculating certain important quantities characterizing the operation of an amplifier stage using a transistor.

I.N.Migulin considered a system of low-frequency and high-frequency parameters of junction transistors which permits the generalization of the theory of crystal and vacuum-tube amplifiers.

V.N.Yakovlev compared the methods of obtaining the input or output volt-ampere characteristics with a falling part, applicable to the analysis of pulse circuits using point-contact triodes; and presented a technique of calculating pulse circuits from assigned pulse parameters.

P.P.Toroshchin discussed simplified methods of calculating amplifier circuits using transistors, and showed the applicability of the proposed technique to the calculation of transformer-coupled amplifier stages.

A.G.Muradyan analyzed the operation of semiconductor amplifiers with series and parallel feedback.

S.M.Gerasimov considered the variation of generator states of a junction transistor at supercritical frequencies, and also the variation of the energetic indices, and gave an analysis of these variations.

In the resolutions on these papers, the Section noted the practical importance of the methods of calculating various circuits using semiconductor devices.

RADIO ENGINEERING SECTION

P.P.Mesyatsev discussed the technological planning of radio equipment. He gave a method of determining flow in the production of radio apparatus.

The Section noted the timeliness of this work and its practical value, and pointed out the fact that the further development of radio engineering questions demanded the study of the theory of probability, mathematical statistics, and the theory of tolerances on radio electric elements.

Ye.M.Khodov discussed questions of combined mechanization and automation of the production of radio engineering apparatus.

The Section noted the great value of the materials presented and recommended improving the work on normalization and standardization of semimanufactured goods and their parts.

M.P.Yemel'yanov, in his paper, considered the method of "comparison by sample", which is used in tuning, regulating and checking modern radio equipment during its production. The Section noted the promising nature of the method presented and recommended continuing the work on introducing it into production.

N.N.Ivliyev devoted his report to methods of making radio equipment air-tight, and protecting the parts from moisture, and to the materials employed for this purpose.

The Section called attention to the necessity of very rapidly reorganizing the production of filling and impregnating materials.

I.P.Chesnokov discussed questions of the mechanization of welding processes in mass production of radio equipment.

The Section called attention to the importance of this work and recommended its continuation.

G.A.Shevtsov and Yu.A.Fedoseyev reported on theoretical questions of parallel reserves as one of the methods of increasing the reliability of electronic apparatus. They gave recommendations for the construction of circuits using parallel reserves.

The Section recommended continuation of the development of the theory and tech-

nique of calculation systems of enhanced reliability.

The report by B.A.Krasyuk considered the data of experimental studies of the variation of the magnetic properties of alloys of the "permalloy" type under the action of gamma-rays.

The Section pointed out the timeliness of these studies and recommended comprehensive tests of the effects of radioactive radiations on new materials for use in radio technology, during the course of their development and introduction.

P.A.Arutyunov presented a report entitled "Study and Calculation of the Accuracy of Certain Elements of Radio Technology"; on which the Section recommended the development of methods of calculating radio technological equipment.

P.Ya.Tsygankov discussed the basic causes for the ageing of quartz resonators, and reported the results of work in building slow-ageing quartz resonators. The Section recommended very speedy practical utilization of the new technique of building slow-ageing quartz resonators.

NEW BOOKS

M.L.Volin. Intermediate frequency amplifiers. Publishing House "Soviet Radio". Moscow, 1956. 232 pp. Price 7.20 rubles.

This book gives the theory and simplified methods of calculating intermediate frequency amplifiers of a radio receiver. It discusses the questions of shielding, decoupling the circuits, and suppressing parasitic feedback in amplifiers. It gives an analysis of various amplifier circuits and designs. The distortions in amplification of various types of signals are considered, and the selection of the basic parameters of amplifiers is justified.

It is written for radio specialists concerned with the design, production and adjustment of radio receivers of all wavelengths. It may be used as a text by students and instructors of colleges and technicians when studying the relevant parts of the course in "Radio Receiving Equipment" and in doing the incourse and graduation research projects.

A.B.Ivanov, L.N.Sosnovkin. Super-High Frequency Pulse Transmitters. Publishing House "Soviet Radio". Moscow,1956. 611 pp. + 1 insert. Price 14.35 rubles.

The theory and calculation of the elements of super-high frequency generators and modulators used in super-high frequency pulse transmitters is discussed. The book gives an exposition of the principles of constructing pulse transmitter circuits, and of questions of their testing and operation.

It considers generators with external excitation, auto-generators using triodes, klystrons, and magnetrons, resonant autogenerators, and pulse modulators.

The book is written for engineering workers in the radio engineering field, and also for students of radio engineering and radar faculties.

G.A.Remez. Course in Basic Radio Engineering Measurements. 2nd Edition. Svyaz'izdat, Moscow,1956. 448 pp. Price 11.25 rubles.

The theory and technique of the basic radio engineering measurements is given (measurements of currents, voltages, power, frequency, phase, nonlinear distortion, modulation, basic parameters of circuits with lumped and distributed constants, field strength). The book is approved as a text for radio engineering colleges and faculties.

M.A.Okrainskiy and V.D.Snetkov. Basic Telephone Cable Measurements. Voyenizdat, Moscow, 1956. 134 pp. Price 3.45 rubles.

This book gives the general concepts and definitions from the theory of long-distance communications that are necessary for a better mastery of the methods of basic measurements of telephone cables. The arrangement and principles of operation of certain special measuring instruments are briefly described (signal generators, nepermeters, frequency-meters, noise-voltage meters, etc.). It considers in detail the practical methods assuring the conduct of that minimum of measurement that is essential for the correct tuning and operations of telephone channels.

Radio Astronomy. Transactions of Fifth Conference on Questions of Cosmogony, March 9-12, 1955. Academy of Sciences USSR, 1956. 567 pp. Edition 1500 copies. Price bound 23.90 rubles.

This book contains the papers and addresses delivered by the members of the Fifth Conference of Questions of Cosmogony, devoted to radio astronomy.

Collection of Scientific Papers. Svyaz'izdat, Moscow, 1956. 226 pp. Ministry of Telecommunications USSR. Central Telecommunications Research Institute. Gratis.

This collection contains the following papers: A.D.Apanasenko "Equations of electromagnetic energy transmission on multiconductor lines";G.F.Pramnek "Expectation of priority call on a single line"; M.I.Mikhaylov "Protection of underground cables from lightning"; A.F.Marchenko "Theoretical principles of determining the aggressiveness of soils and ground waters on the lead sheathing of communications cables"; N.N.Gerasimov and A.A.Kon'kov "Method of measuring electromagnetic coupling on aerial lines"; N.A.Stepanova "Methods of measuring electrical fluctuations"; L.T.Kim "Rational methods of obtaining the driving frequency for multi-channel systems designed for short-distance communication"; V.P.Rurevich "Measurement of the parameters of transistors at low frequencies for faint signals"; M.D.Korf "Effect of deviation of filter elements on the frequency characteristic of damping of the filter in measuring a small number of elements".

A.F.Ioffe. Semiconductors and their use. Academy of Sciences USSR, Moscow-Leningrad, 1956. 71 pp. (AN SSSR. Popular Scientific Series). Price 1 ruble.

Describes the electrical properties of semiconductors in industry and in daily life, and tells of the history and prospects of the study of semiconductors.

ANNOUNCEMENT

On and after October 1, 1956, subscriptions will be accepted for 1957 for the journal Radiotekhnika, organ of the Scientific and Technical Society of Radio Engineering and Telecommunications imeni A.S.Popov.

(12th year of publication)

The journal presents information on the accomplishments of Soviet radio engineering and on the working experience of research institutes, laboratories, scientists and specialists in the field of radio engineering.

It contains reviews and bibliographies on radio engineering literature, as well as individual papers printed in the periodical literature.

It reports the activities of the Scientific Society for Radio Engineering and Telecommunications imeni A.S.Popov.

The journal is intended for scientists, engineers, instructors and students of advanced courses in higher technical educational institutions.

The subscription rate for 12 months is 72 rubles, or 36 rubles for 6 months.

Subscriptions are accepted at the city branches of Soyuzpechat', at telecommunications offices, sections and agencies, at subscription agencies, and by authorized agents at industrial plants, in educational institutions and organizations.

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