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SOME BASIC RELATIONS PERTAINING TO HIGH-POWER KLYSTRON AMPLIFIERS

by

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This article survey some basic relations determining the design of high-power klystron amplifiers. Special attention is turned to analysis of band limitations of the frequencies passed by the output oscillatory system and to the minimum allowable feed voltage. Also a description of the conditions in which these limitations become less valid.

1. Introduction

Klystron amplifier engineering has been lately progressing along an arduous, peculiar and tortuous path of development. Klystron amplifiers have been developed less rapidly than other types of super-high-frequency amplifiers. However, the pessimistic opinions often voiced about the possibilities of their further development have been refuted by facts again and again.

The paramount idea underlying the nature of the operation of klystron amplifiers, namely, the idea of velocity modulation of the electron beam has been formulated by D.A.Rozhanskiy as early as in 1932, as published literature (Bibl.1) indicates. However, a year after the invention in 1938 of toroidal vibrators the first workable klystron amplifiers began appearing in 1939 (Bibl.2).

At that time, klystron amplifiers had a low efficiency of the order of 15 to 20% owing to the large losses incurred at collisions of electrons with the grids and walls of tubes in the drift space, and also owing to the relatively disadvantageous grouping of electrons into bunches at velocity modulation in only one control gap. Therefore, klystron amplifiers used to be designed for low power only; power of the order of tens of watts. Such a situation continued throughout nearly the entire nineteen forties.

It was only at the turn of the forties and the beginning of the fifties that major modifications had been introduced in klystron amplifiers so as to increase sharply their efficiency and amplification factor.

A longitudinal focusing magnetic field was applied, and this reduced the losses from the collisions of electrons with the side walls of drift space. The grids were eliminated and the interaction of electrons with the fields between grids was supplanted by their interaction with the fields of the gaps formed by transverse slits in drift-space tubes. This eliminated the losses formerly incurred by collision of electrons with grid conductors. Lastly, there were introduced two or more successive drift spaces with passive resonators near the intermediate gaps. This had improved the longitudinal focusing of electrons into bunches, and it also had increased sharply the amplification factor.

The considerable increase of the actually attainable electron efficiency to values of the order of 40 to 50% and more has made the development of high-power klystron amplifiers more profitable. In the last few years numerous descriptions of these devices have been published (Bibl.3 to 6). The power these devices develop already is reaching tens of kilowatts at continuous-wave operation and tens of megawatts at pulsed operation.

Still, all modern high-power klystron amplifiers require very high feed voltages and have narrow frequency passbands. Moreover, for waves longer than 50 - 60 cm, the dimensions of klystron amplifiers are too large. These shortcomings hamper greatly any more widespread use of klystron amplifiers.

The present paper analyzes the causes of the abovementioned shortcomings and the possible ways and means for their partial overcoming.

2. Design Limitations of Klystron Amplifiers

Hereafter in this paper length will be expressed in centimeters and electrical parameters in amperes, volts, watts and ohms.

Let us survey the conditions in which an electron beam traverses the drift space of a two-resonator transit klystron amplifier.

According to the elementary theory of longitudinal defocusing, which assumes an infinite cross-sectional area of the beam in the absence of side walls (Bibl.7), the degree of focusing of electron bunches for the fundamental-frequency current is determined by Bessel's function of the first kind $J_1(x)$, whose argument equals

$$x = \frac{2\pi}{\lambda}\,\frac{c}{v_0}\,\frac{\mu}{2}\,\frac{\sin hs}{h}. \qquad (1)$$

Here λ and c are the wave length and velocity of free electromagnetic waves; v_0 is the velocity of the electrons started by feed voltage U_0; s is the length of drift space; μ is the ratio of amplitude of control voltage to voltage U_0; and h is the parameter of longitudinal defocusing, related to the action of space charge. This parameter depends on density j_0 of the electron current and is expressed by the following formula (Bibl.8)

$$h = \sqrt{3}\,\pi \cdot 10^{2}\,\frac{j_0^{\frac{1}{2}}}{U_0^{\frac{3}{4}}}. \qquad (2)$$

On introducing some correction factor $\varkappa$ that is higher than unity and takes into account the effective attenuation of the defocusing parameter owing to the ultimate value of the cross-sectional area of the electron beam, and which also takes into account the influence of the side walls, eq.(1) can be represented in this form

$$x = \frac{2\pi}{\lambda}\,\frac{c}{v_0}\,\frac{\mu}{2}\,\frac{\sin\frac{hs}{\varkappa}}{\frac{h}{\varkappa}}. \qquad (3)$$

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At

$$\frac{hs}{x} = \frac{\pi}{2} \qquad (4)$$

the magnitude of x is at its maximum, equal to

$$x_{max} = \frac{2\pi}{\lambda}\frac{c}{v_0}\frac{\mu x}{2h} = \sqrt{\frac{\pi}{3}}\cdot 10^{-2}\frac{c}{v_0}\frac{\mu x}{\lambda}\frac{U_0^{\frac{3}{4}}}{j_0^{\frac{1}{2}}}.$$

Assuming that

$$\frac{v_0}{c} \approx \frac{U_0^{\frac{1}{2}}}{500},$$

which is correct at $\left(\frac{v_0}{c}\right)^2 \ll 1$, when the correction deductible from the theory of relativity is still small, we find

$$x_{max} = 5\sqrt{\frac{\pi}{3}}\frac{\mu x}{\lambda}\frac{U_0^{\frac{1}{4}}}{j_0^{\frac{1}{2}}}. \qquad (5)$$

The maximum of the function $J_1(x)$ is present at $x = 1.84$. Having to satisfy the condition

$$x_{max} \geqslant 1{,}84\ \beta, \qquad (6)$$

we obtain the following condition for current density

$$j_0\lambda^2 \leq \frac{25\pi}{3\cdot 1{,}84^2}\left(\frac{\mu x}{\beta}\right)^2 U_0^{\frac{1}{2}}. \qquad (7)$$

Considering that the maximum of Bessel's function is obtained in a rather blunted form, the multiplier β can be assumed to be somewhat below unity.

The corresponding length of drift space s equals, in accordance with eqs.(4) and (2)

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$$s = \frac{1}{200}\sqrt{\frac{\pi}{3}\frac{U_0^{\frac{3}{4}}}{j_0^{\frac{1}{2}}}}\varkappa, \tag{8}$$

or, considering eq.(7),

$$\frac{s}{\lambda} \geqslant \frac{1,84}{1000}\frac{\beta}{\mu}U_0^{\frac{1}{4}}. \tag{9}$$

The application of eqs.(4), (7), (8), and (9), corresponds to the conditions present at short waves when the lengths of drift spaces are not large. At longer waves, for instance waves of the order of 100 cm, when it is desirable to reduce the lengths of drift spaces, it may prove expedient to select the argument of the sine in eq.(3) below $\frac{\pi}{2}$, i.e., to assume that

$$\frac{hs}{\varkappa} = \beta' \frac{\pi}{2}, \tag{4}$$

where $\beta' < 1$.

Considering that, when the argument of that sine is selected below $\frac{\pi}{2}$, the sine varies somewhat, it is possible to take the coefficient β' much below unity. Then, eqs.(7) and (9) will assume the following form:

$$j_0\lambda^2 \leqslant \frac{25\pi}{3\cdot 1,84^2}\left(\frac{\mu\varkappa}{\beta}\sin\frac{\pi\beta'}{2}\right)^2 U_0^{\frac{1}{2}}; \tag{7'}$$

$$\frac{s}{\lambda} \geqslant \frac{1,84}{1000}\frac{\beta\beta'}{\mu}U_0^{\frac{1}{2}}. \tag{9'}$$

The multipliers β and β' should, of course, be so selected as to preserve a quite satisfactory bunching of electrons.

Proceeding from here on we will employ eq.(7), on keeping in mind the necessity of replacing β by $\frac{\beta}{\sin\frac{\pi\beta'}{2}}$ at a reduced length of drift space.

Considering that

$$U_0 = \frac{P_0}{I_0} = \frac{P_0}{i_0 S},$$

where P_o is the power of electron beam, I_o is the feed current, and S is the cross-sectional area of beam, we find according to eq.(7) the limitation for the lowest value of feed voltage as imposed by the conditions of longitudinal repulsion of electrons

$$U_0^{\frac{3}{2}} \geqslant \frac{3 \cdot 1,84^2}{25\pi} P_0 \frac{\lambda^2}{S} \left(\frac{\beta}{\mu\varkappa}\right)^2. \qquad (10)$$

As can be seen from the above inequality, the cross-sectional area of the electron beam should be large in order to ensure the possibility of applying a low feed voltage.

3. Case of Solid Cylindrical Electron Beam

Present-day high-power klystron amplifiers operate with an electron beam having a circular cross-section (Fig.1). On designating the diameter of the beam by D, we have

$$S = \frac{\pi D^2}{4}.$$

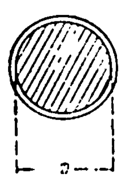

Fig.1

For a good interaction of electrons with gap fields, there should be

$$D < Ad,$$

where d stands for gap width. This width is, in turn, limited by the requirement

$$d < \frac{\lambda_q}{B} = \frac{\lambda}{B} \frac{v_0}{c} \approx \frac{\lambda U_0^{\frac{1}{2}}}{500 B}, \qquad (11)$$

where λ_q stands for the distance traversed during one oscillatory period by an electron accelerated by voltage U_o. The adopted values of the coefficients A and B are determined by the conditions of the interaction between the electron beam and gap fields (Bibl.8, 9).

The following limiting requirements are obtained for the diameter and the cross-sectional area of the electron beam:

$$D < \frac{A}{500\,B} \lambda U_0^{\frac{1}{2}}, \qquad (12)$$

$$\frac{S}{\lambda^2} \leqslant \frac{\pi}{10^6} \left(\frac{A}{B}\right)^2 U_0. \qquad (13)$$

On considering the latter limitation we obtain in accordance with eq.(10) the requirement for the minimum allowable feed voltage

$$U_0^{\frac{5}{2}} \geqslant 0{,}415 \cdot 10^6 \cdot P_0 \left(\frac{\beta}{\mu x} \cdot \frac{B}{A}\right)^2. \qquad (14)$$

For instance, assuming that $\frac{\beta}{\mu x} = \frac{4}{3}$ and B = 5.5 and A = 1.6, we obtain at P_o = 20 kwt

$$U_0^{\frac{5}{2}} \geqslant 1{,}72 \cdot 10^{10}$$

or

$$U_0 > 12{,}5 \text{ кв}.$$

Published literature (Bibl.(4) contains a description of a 30-cm-wave klystron amplifier with very similar data. For that amplifier, P_o = 19.5 kwt and U_o = 13 kilovolts. Gap width measures half an inch, which is corresponded by B ≈ 5.7. Thus, the assumed value of $\frac{\beta}{\mu x}$ is evidently approximate to the real one.

4. A Band of Frequencies Passed by Klystron Amplifiers With a Solid Cylindrical Electron Beam.

In view of the very high voltages of the klystron amplifiers of the here-described type it is necessary to apply output oscillatory systems with a very high resonant resistance R, and hence with a high quality Q when in loaded state. This is equivalent to a narrow frequency passband.

The following are the formulas for the necessary resonant resistance R and quality Q:

$$R = \frac{U}{I_1} = \frac{\xi}{\gamma_1}\frac{U_0}{I_0} = \frac{\xi}{\gamma_1}\frac{U_0^2}{P_0}, \quad (15)$$

$$Q = \frac{R}{\rho_u} = \frac{\xi}{\gamma_1}\cdot\frac{U_0^2}{P_0\,\rho_u}. \quad (16)$$

Here U is the amplitude of alternationg voltage on output gap; I_1 is the first harmonic of the current exciting the output oscillatory system; $\xi = \frac{U}{U_0}$ is the intensity of mode; $\gamma_1 = \frac{I_1}{I_0}$ is the coefficient of the first harmonic of current; and ρ_u is the parallel active characteristic (Bibl.10) of the output oscillatory system, related to voltage U.

On considering requirement (14) we obtain

$$R \geqslant 0{,}5\cdot 10^4\,\frac{\xi}{\gamma_1}\left(\frac{\beta}{\mu\lambda}\,\frac{B}{A}\right)^{\frac{8}{5}}\frac{1}{P_0^{\frac{1}{5}}}\ldots \quad (17)$$

Assuming that in the above-reviewed example P_0 = 20 kwt and U_0 = 12.5 kilovolts, and also assuming that $\frac{\xi}{\gamma 1}$ = 0.8, we find

$$R = \frac{\xi}{\gamma_1}\frac{U_0^2}{P_0} = 0{,}8\,\frac{12{,}5^2\cdot 10^6}{2\cdot 10^4} = 6300 \text{ ohms}.$$

Thus, there is obtained a relatively very high resistance and, consequently, a narrow passband.

In effect, it is difficult to increase the characteristic ρ_u of the toroidal **vibrator** even at a relatively large gap between the ends of drift tubes. In practice it can barely be made higher than 60 – 70 ohms. Therefore, the necessary quality obtained amounts to a value of hundred or more units, which corresponds to an output-resonator frequency passband of the order of 1% or less from the carrier

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frequency (at computations in terms of half power level on the edges of the frequency passband). Passbands of such an order are actually observed in practice (Bibl.4).

As can be seen from formula (17), resonant resistance varies very slowly with a change in power, that is to say, as the root of the fifth degree of power. Therefore any much more favorable conditions for the width of the frequency passband are present only at very high powers of klystron amplifiers. For example, according to formula (17), a fourfold expansion of the passband is corresponded by an approximately thousandfold rise in power.

In this respect, a characteristic amplifier described in literature (Bibl.5) is the super-high-power pulsed klystron amplifier with a pulse output power of the order of 20 mwt and with U_0 = 400 kilovolts, I_0 = 250 amperes, and Q ≈ 30. The passband in such an amplifier is thus comparatively wide.

For that amplifier, we have (assuming $\frac{\varepsilon}{\gamma_1} = 0.8$)

$$R = \frac{\varepsilon}{\gamma_1}\frac{U_0}{I_0} = 0{,}8 \cdot \frac{4 \cdot 10^5}{250} = 1280 \text{ ohms}.$$

Therefore, quality Q = 30 at loaded state is corresponded by the conveniently obtained characteristic

$$\rho_\mu = \frac{R}{Q} = \frac{1280}{30} = 43 \text{ ohms}.$$

Let us also note that, as can be concluded from eq.(17), a reduction in resistance R and expansion of frequency passband are also possible to achieve by reducing the amplification factor (reducing the magnitude of $\frac{\beta}{\mu}$) or by reducing intensity ε i.e., weakening the braking of electrons by the field of the gap.

5. Klystron Amplifiers With Enlarged Cross-Sectional Area of the Electron Beam

As can be seen from the aforecited formulas, it is possible to reduce the necessary feed voltages and expand the frequency passband by eliminating the limitation imposed by requirement (13) on the cross-sectional area of the electron beam.

This limitation becomes dispensable in, for example, the event of a multi-beam klystron amplifier described in the book by Warnecke and Guenard (Bibl.9), Chapter XXXVII, and to an even greater degree, in the event that none of the electron beams is solid and all adhere to the walls of the drift-space tube (Fig.2).

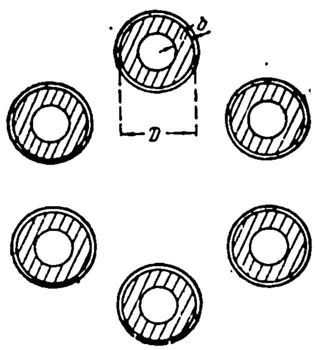

Fig.2

For such a multi-beam klystron amplifier the inequality (12) is replaced by the

$$b \leqslant \frac{A'}{2} d, \qquad (17)$$

where b stands for thickness of the electron layer (Fig.2), and A' stands for a somewhat smaller coefficient than A.

The overall area of all n rays (at D ≫ b) approximately equals

$$S \approx \pi D b n \leqslant \pi D \frac{A'}{2} n d, \qquad (18)$$

or, on taking into account inequality (11)

$$\frac{S}{\lambda^2} \leqslant \frac{\pi}{1000} \frac{A'}{B} \frac{nD}{\lambda} U_0^{\frac{1}{2}}, \qquad (19)$$

where D is the diameter of each beam.

Correspondingly, inequality (10) assumes this form

$$U_0 \geqslant \frac{1,84}{\pi} \frac{\beta}{\mu \varkappa} \sqrt{120 P_0 \frac{B}{A'} \frac{\lambda}{nD}}. \qquad (20)$$

For resonant resistance we obtain

$$R = \frac{\xi}{\gamma_1} \frac{U_0^2}{P_0} \geqslant \left(\frac{1,84}{\pi}\right)^2 120 \frac{\xi}{\gamma_1} \frac{B}{A'} \frac{\lambda}{nD} \left(\frac{\beta}{\mu \varkappa}\right)^2. \qquad (21)$$

As can be seen from the above, resonant resistance is not affected by power. On selecting a sufficiently high value of $\frac{nD}{\lambda}$ it is possible to obtain comparatively low values of U_o and R. For example, if it is necessary that R be not over 1000 ohms, this should be the formula

$$\frac{nD}{\lambda} \geqslant 0{,}12\left(\frac{1{,}84}{\pi}\right)^2 \frac{\xi}{\gamma_1}\frac{B}{A'}\left(\frac{\beta}{\mu x}\right)^2.$$

Assuming that A' = 12 and, as before, B = 5.5, $\frac{\xi}{\gamma_1}$ = 0.8, and $\frac{\beta}{\mu x} = \frac{4}{3}$, we find

$$\frac{nD}{\lambda} > 0{,}27.$$

Naturally, the amplifier's design should be adapted for ensuring a fairly adequate decoupling between the output oscillatory system and the preceding oscillatory system.

For the required voltage we have

$$U_0 = \sqrt{\frac{\gamma_1}{\xi} R P_0}\,. \qquad (22)$$

In the example described here, assuming that R = 1000 ohms, P_o = 20 kwt, and $\frac{\xi}{\gamma_1}$ = 0.8, we obtain

$$U_0 = 5000\ \text{v},$$
$$I_0 = \frac{P_0}{U_0} = 4\ a.$$

Thus we obtain a relatively low feed voltage and a relatively high feed current.

Let us now turn our attention to the frequency passband of the output oscillatory system. Let us take note of the capacitance of the gaps, which, at the presence of several large-diameter drift-space tubes, constitutes nearly the entire capacitance of the output oscillatory system and may, therefore, be used for an approximate appraisal of its characteristic ρ_u.

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The linear capacitance of gap width d between two flat tapes having a width h (Fig.3) is determined by the following approximate formula

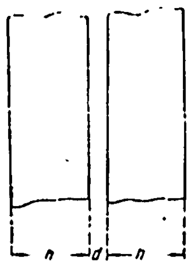

Fig.3

$$C_1 = \frac{1}{2\pi^2}\ln\frac{h}{d}\ \frac{cm}{cm}. \qquad (23)$$

Therefore; for the above-discussed example, it can be approximately assumed (on increasing capacitance somewhat) that

$$C = C_1\pi Dn = \frac{nD}{2\pi}\ln\frac{h}{d}\,cm. \qquad (24)$$

Correspondingly

$$\rho_u \approx \frac{15\lambda}{\pi C} = \frac{30\lambda}{nD}\,\frac{1}{\ln\frac{h}{d}}. \qquad (25)$$

Assuming, for instance, that h = 20d, we obtain, in the event of $\frac{nD}{\lambda} = 0.27$

$$\rho_u \approx 37\ ohms$$

For the necessary loaded quality of the output oscillatory system, the following formula is obtained on taking requirements (21) and (25) into account

$$Q = \frac{R}{\rho_u} \geqslant \left(\frac{1,84}{\pi}\right)^2 4\,\frac{\xi}{\gamma_1}\,\frac{B}{A'}\left(\frac{\beta}{\mu\varkappa}\right)^2 \ln\frac{h}{d}. \qquad (26)$$

Here, the right-hand part does not depend on P_o, U_o and λ.

Assuming; as before, that B = 5.5, A' = 1.2, $\frac{\beta}{\mu\varkappa} = \frac{4}{3}$ and $\frac{h}{d} = 20$, we find

$$Q_{min} \approx 27.$$

Thus, in these conditions, we obtain a low necessary loaded quality corresponding to a comparatively wide frequency passband. Of course, the above computations are approximate.

As before, it is possible here too to expand further the frequency passband by reducing the amplification factor, i.e., the magnitude of $\frac{\beta}{\mu}$, or by reducing intensity F. The enlargement of the cross-sectional area of the electron beam permits not only to reduce the feed voltage and to expand the frequency passband but also to obtain certain other benefits.

1. In accordance with inequality (7), the density of the electron-beam current decreases as $U_o^{\frac{1}{2}}$. This may be of vital importance in short-wave operations when the current density necessary for the most effective performance of an amplifier and increasing in **proportion to the reduction to lambda$_2$,** has a very high value that is difficult to attain.

2. At long-wave operation, a reduction in voltage may prove expedient from the viewpoint of reducing the necessary lengths of drift spaces determined by requirement (9') and proportional to $U_o^{\frac{1}{2}}$, and hence also from the viewpoint of reducing the lengthwise dimensions of the tube and the length of paths traversed by electrons. At the same time, the dimensions of the oscillatory systems are **reduced** by increasing their usefully utilized capacitance, which is also important at very-long-wave operation.

These circumstances may expand somewhat the boundaries of the wave band within which the klystron amplifiers can be applied.

6. Klystron Amplifiers With Two and More Drift Spaces **Arranged in Series**

Although the formulas described above pertain to klystron amplifiers with a single drift space, they can also be related to amplifiers with two or more drift spaces, upon being somewhat modified, (in particular, with regard to coefficients β and x).

We have been so far concerned only with the problem of expanding the passband of the output oscillatory system. A corresponding expansion of passband for the input and intermediate oscillatory systems can be achieved by their ballast loading. In view of the relative smallness of the amplitudes of r – f voltages occurring in these oscillatory systems, the powers emitted in the ballast loads will be low compared to output power, so that the overall efficiency will be decreased only a little. Naturally, the amplification factor will be somewhat below its maximum possible value. However, this circumstance is not very important in view of the very high amplification factors inherent in klystron amplifiers with two or more drift spaces.

Appendix

Beside the above-described, as per inequality (7), limitation for the electron-beam current, imposed by the longitudinal debunching action of the space charge, it is also necessary to take into account the current limitation due to the decrease (the so-called "sagging") of the potential inside the drift tube at the presence of the electron beam.

This second limiting factor assumes the following form (see Bibl.8, p.78) at the presence of a longitudinal focusing magnetic field and allowing, as is usually done), for a decrease in potential by no more than 10%

$$I_0 < 6{,}25 \cdot 10^{-6} U_0^{\frac{3}{2}}. \quad (27)$$

At the same time, with respect to a single-beam klystron amplifier, according to (7) and (13), the limitation pertaining to the longitudinal debunching effect is obtained in the following form

$$I_0 \quad j_0 S < 24{,}3 \cdot 10^{-6}\left(\frac{A}{B} \cdot \frac{\mu x}{\beta}\right)^2 U_0^{\frac{3}{2}}$$

or, at the afore-adopted values of A = 1.6, B = 5.5, and $\frac{\beta}{\mu x} = \frac{4}{3}$

$$I_0 < 1{,}15 \cdot 10^{-6} \cdot U_0^{\frac{3}{2}}. \quad (28)$$

Considering that this requirement appears to be more exacting than (27), it has also been taken into account in the text.

With respect to a multi-beam klystron amplifier, the potential-decrease requirement assumes this form

$$I_0 < 6,25 \cdot 10^{-6}\, n\, U_0^{\frac{3}{2}}. \qquad (29)$$

As for the limitations pertaining to the longitudinal debunching effect, on the basis of (7) and (19) it follows that

$$I_0 < 24,3 \cdot 10^{-3} \frac{A'}{B} \frac{nD}{\lambda} U_0 \, ,$$

or, at the afore-adopted values of A' = 1.2, B = 5.5 and $\frac{nD}{\lambda} = 0.27$

$$I_0 < 1430 \cdot 10^{-6} \cdot U_0 \qquad (30)$$

If

$$n > \frac{230}{U_0^{\frac{1}{2}}} \, ,$$

then requirement (30) proves to be more exacting than (29), and therefore it, too, has been considered in the text.

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APPROXIMATE METHODS OF COMPUTING THE FIELD STRENGTH OF ULTRA SHORT WAVES WITH CONSIDERATION OF TERRAIN RELIEF

by

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This article provides approximate methods of computing the field strength of ultra short radio waves on considering the influence of terrain relief. Interference formulas are employed for computing field strength on open routes. Field strengths on half-open and closed routes are determined by an approximate method based on the approximation of route obstacles by spheres whose radius is determined by the specific features of the obstacle, and on the exponential relationship between the attenuation multiplier and the diffraction angle.

At present, owing to the studies of the diffraction of radio waves conducted by Academician V.A.Fok and other savants, the problem of determining field strength at propagation of radio waves over smooth spherical surface of the Earth and at linear variation of the permittivity of the air with altitude, can be regarded as being completely solved (Bibl.1, 2).

However, the simplifying postulates made at a rigorous solution of this problem (smooth spherical Earth, linear variation of permittivity of the air with altitude, and others) cease to apply when the ultra short wave range is concerned, and therefore computations of field strength of ultra short waves can be regarded as correct only if they take into account the concrete features of terrain relief and the fluctuations in weather conditions.

Approximate methods of computing field strength on real routes upon taking into account the influence of terrain relief and refraction, will be described below.

These methods are derived from the approximate methods mentioned in Bibl.5 and 6.

1. Classification of Routes

Let us examine the profile of a real route illustrated in Fig.1. In computations of field strength on real routes it is convenient to employ the extent of clearance between line of sight AB and route profile, rather than the heights of antennas above sea level or other nominal zero level. Route profiles are always known, and the extent of clearance can be determined directly from route profile.

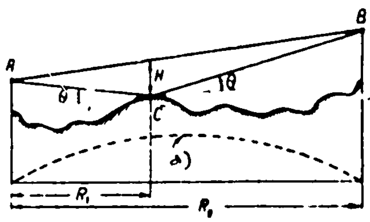

Fig.1

a) Sea level or other nominal zero level

Let us regard the extent of clearance as positive when line of sight AB passes over route profile, and as negative when this line intersects route profile.

As depending on the extent of clearance between line of sight AB and the highest point on route profile, we will divide routes into open, half-open and closed ones.

Let us term open routes those routes on which the extent of clearance at the highest point of route profile, H, is greater than the value of H_0, which is yielded by the following formula

$$H_0 = \sqrt{\frac{1}{3} R_0 \lambda \kappa (1 - \kappa)}, \qquad (1)$$

where R_0 is route length, λ is wave length, and k is the relative coordinate of the highest point on route profile,

$$\kappa = \frac{R_1}{R_0}, \qquad (2)$$

R_1 is the distance from the left-hand end of route to the highest point on route

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profile.

The value H_o refers to an extent of clearance at the highest point on route profile at which the difference in mode of travel between a direct wave and a wave reflected at the said point will equal $\lambda/6$. On real routes the phase of the reflection factor approximates 180^o, and therefore when the extent of clearance reaches the value H_o then the attenuation multiplier will equal unity practically outside of its relationship to the value of the modulus of reflection factor.

We will term as half-open those routes for which

$$H_0 > H > 0,$$

and we will term a route as closed when

$$H < 0.$$

Let us note that the classification of routes into open, half-open and closed, as depending on extent of clearance, has sense only at definite conditions of refraction. Extent of clearance varies with conditions of refraction and it can be found, for example, that a route that is half-open or closed at some conditions of refractions will become an open route at other conditions of refraction.

2. Computing the Attenuation Factor on Open Routes

1. The field on open routes has an interference character. Generally speaking, field strength in locus of reception represents the geometric sum of the fields of direct wave and of the waves reflected from the ground, forest, etc., surfaces.

If there are p reflection points on a route, the attenuation multiplier V can be computed by means of this formula

$$V = \left[\left(1 + \sum_{i=1}^{p} |\Phi|_i \cos\gamma_i\right)^2 + \left(\sum_{i=1}^{p} |\Phi|_i \sin\gamma_i\right)^2\right]^{\frac{1}{2}}. \qquad (3)$$

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where $|\Phi|_i$ stands for the modulus of the reflection factor at the i^{th} reflection point, and γ_i stands for the phase shift between the direct and the i^{th} reflected waves, equal to

$$\gamma_i = \frac{2\pi}{\lambda} \Delta r_i + \beta_i.$$

Here Δr_i stands for the difference in travel between the direct and the i^{th} reflected waves, which should be determined upon taking into account the specific features of terrain relief on the route

β_i stands for the phase of the reflection factor for the i^{th} reflection point.

When only one reflection point is present on a route, eq.(3) is reduced to this customary form

$$V = \sqrt{1 + |\Phi|^2 + 2|\Phi|\cos\gamma}. \quad (5)$$

2. The difference in wave travel Δr_i can be determined as follows: On the profile of a real open route we locate the reflection points, i.e., the points at which the angles between a line tangential to the profile and the straight lines plotted from each such reflection point to the transmission and reception points are mutually equal angles. Fig.1 depicts a route with a single reflection point denoted by the letter C.

After locating the reflection points we determine from the profile the values of clearance H_i for every such point. The wave travel difference Δr_i for the i^{th} reflection point can be calculated according to this formula

$$\Delta r_i = \frac{H_i^2}{2 R_0 \kappa_i (1 - \kappa_i)}, \quad (6)$$

where k_i is the relative coordinate of reflection point, determinable from the formula

$$\kappa_i = \frac{R_i}{R_0}, \quad (7)$$

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R_i stands for the distance from the left-hand end of route to the $\underline{i}^{th}$ reflection point.

3. The diversity of the forms of the Earth's surface complicates extremely a rigorous theoretical determination of the reflection factors for such surface, and yet, in the practically important cases of computing the attenuation multiplier in interference minimums it is necessary to determine fairly accurately the values of the reflection factor. The theory can merely supply some criterions for evaluating the expected values of reflection factors.

One of such criterions is the value of the minimum zone encompassing the geometrical reflection point, which - when there is no influence from any remaining reflection surface - yields the same value of reflection factor as an infinitely larger surface with the same electrical parameters. For a planar reflecting surface such a minimum zone represents an ellipse encompassing the geometrical reflection point and having an area equal to one third of the area of Fresnel's first zone.

If the planar reflection surface is level* within the confines of the minimum zone, at least, then the reflection factor can be determined from the ordinary graphs used for a flat reflection surface according to known values of electrical parameters of surface, wave length, and angle of slide, which last can be determined from the following formula**

$$\theta = \frac{H}{2 R_0 \kappa (1 - \kappa)}. \qquad (8)$$

It should be noted that the presence of ground sections with differing relief and plant-covered surface can cause a change in the value of the reflection factor determined on the assumption of the action of a single minimum zone, by increasing or reducing that value.

If the reflection surface has a smooth convex form, the value of the modulus

* In the sense that it satisfies Rayleigh's criterion.

** For simplicity, the subscript i will be hereafter discarded.

of reflection factor decreases owing to the divergence of reflected waves. The question of the reflection from convex surfaces has been discussed as a whole in another paper (Bibl.3). In the particular instance when a reflection surface can be approximated by a sphere of a radius b, the divergence factor D by which the value of the modulus of the factor of reflection from a planar surface $|\Phi|_{pl}$ has to be multiplied in order to obtain $|\Phi|$ can be computed according to the following formula

$$D = \frac{1}{\sqrt{1 + \frac{4\kappa^2(1-\kappa)^2 R_0^2}{bH}}}. \qquad (9)$$

Considering that usually $b \gg H$ and $b \gg R_0$, then $D < 1$ only at low values of H. $D \to 1$ at high values of H, and $D \to 0$ at $H \to 0$.

The radius of the approximating sphere, b, is determined by the form of the reflection surface and is expressed by r and Δy through the formula

$$b \approx \frac{r^2}{8\Delta y}. \qquad (10)$$

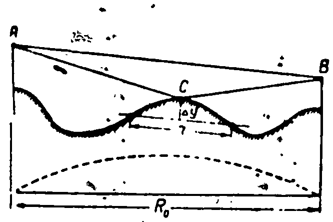

Fig.2

On substituting (10) into (9) we obtain the formula for D

$$D = \frac{1}{\sqrt{1 + \frac{32\alpha\kappa^2(1-\kappa)^2 H_0}{l^2 H}}}, \qquad (11)$$

where

$$\alpha = \frac{\Delta y}{H_0}, \qquad (12)$$

$$l = \frac{r}{R_0}. \qquad (13)$$

4. On real routes the angles of slide θ are always small and the phase of the reflection factor can be considered equal to 180°. Therefore, on routes with a single reflection point, the interference maximum can be observed at values of clearance H_m equal to

$$H_m = \sqrt{(2m-1) R_0 \lambda \kappa (1-\kappa)}, \qquad (14)$$

where m = 1, 2, 3 . . . is the number of the interference maximum.

Interference minimums occur at these values of clearance

$$H_n = \sqrt{2n R_0 \lambda \kappa (1-\kappa)}, \qquad (15)$$

where n = 1, 2, 3 . . . is the number of the interference minimum.

The attenuation multiplier V_m for the $\underline{m}^{th}$ interference maximum equals

$$V_m = 1 + |\Phi|_m. \qquad (16)$$

and for the $\underline{n}^{th}$ interference minimum

$$V_n = 1 - |\Phi|_n. \qquad (17)$$

If the reflection surface is a smooth spherical one, the values of $|\Phi|_m$ and $|\Phi|_n$ are determined by these formulas

$$\left.\begin{array}{l} |\Phi|_m = |\Phi|_{plm} D_m \\ |\Phi|_n = |\Phi|_{pln} D_n \end{array}\right\}, \qquad (18)$$

where $|\Phi|_{plm}$ and $|\Phi|_{pln}$ stand for values of the moduli of reflection factors for the smooth planar surface, corresponding to angles of slide:

$$\Theta_m = \frac{1}{2}\sqrt{\frac{(2m-1)\lambda}{R_0 \kappa (1-\kappa)}}, \qquad (19)$$

$$\Theta_n = \sqrt{\frac{n\lambda}{2 R_0 \kappa (1-\kappa)}}, \qquad (20)$$

and D_m and D_n stand for values of divergence factor, determined by the following formulas

$$D_m = \frac{1}{\sqrt{1 + 18{,}5 \frac{a \kappa^2 (1-\kappa)^2}{l^2 \sqrt{2m-1}}}}, \qquad (21)$$

$$D_n = \frac{1}{\sqrt{1 + 13{,}1\,\frac{a\kappa^2(1-\kappa)^2}{l^2\sqrt{n}}}}. \qquad (22)$$

Equations (19) to (22) are commensurately derived from eqs.(8) and (11) after substituting into them the values of clearance H_m and H_n as determined from eqs.(14) and (15).

Considering that the value of the attenuation multiplier in interference minimums is very critical in relation to the value of the modulus of reflection factor, it is necessary to know the dimensions of the minimum zone for precisely these minimums. It can be demonstrated that the dimension of the large axis $2a_n$ of the ellipse representing the minimum zone is determined by the following formula with regard to the $\underline{n}^{th}$ interference minimum

$$2a_n = R_0 \frac{\sqrt{\frac{1}{3}\left(\frac{1}{3} + 2n\right)}}{\frac{1}{3} + \frac{n}{2\kappa(1-\kappa)}}, \qquad (23)$$

and the small axis $2b_n$ of this ellipse equals

$$2b_n = \frac{1}{2}\sqrt{R_0\lambda}\,\sqrt{\frac{\frac{1}{3}\left(\frac{1}{3} + 2n\right)}{\frac{1}{3} + \frac{n}{2\kappa(1-\kappa)}}}. \qquad (24)$$

3. Computing the Attenuation Multiplier on Half-Open and Closed Routes

1. At low values of clearance between line of sight and route profile the interference formulas yield incorrect results and the attenuation multiplier has to be calculated on the basis of the theory of electrodynamic diffraction.

The below-described approximate method of computing the attenuation multiplier on half-open and covered routes is based on the approximation of shielding obstacles by spheres whose radius is determined by the specific features of obstacle. Such an approximation has been chosen because it allows the application, in a somewhat modified form, of all the results of the theory of the electrodynamic diffraction of radio waves around smooth spherical surface of the Earth.

On first glance it may be found that in a number of cases real obstacles can be more effectively approximated by other geometrical bodies, for instance, by ellipsoids of rotation, cylinders, etc. However, it can be demonstrated that field strength is not particularly affected by the form of approximation. In effect, the extent of space that is of basic importance to the process of propagation of ultra short radio waves has the form of an ellipsoid of rotation stretched along the route, with its focii represented by the transmission and reception points (see Bibl.4). The large axis of this ellipsoid is hundreds and thousands of times larger than its small axis. Consequently, the value of field strength is affected chiefly by the radius of the curvature of obstacle on the plane passing through corresponding points and the Earth's center. The value of the radius of curvature of obstacle on the perpendicular plane is of no essential importance so long as it is not too small in comparision with the former radius.

2. Let us now turn to Fig.3, which depicts a route profile with θ denoting the angle of diffraction. Let us assume that terrain relief before and behind the obstacle is such as to cause the absence of any interference phenomena before and behind the obstacle. It is known from the diffraction theory that the value of the attenuation multiplier V in the umbral and penumbral zones decreases exponentially at an increase in diffraction angle θ. Considering that angle θ is related to clearance by this formula

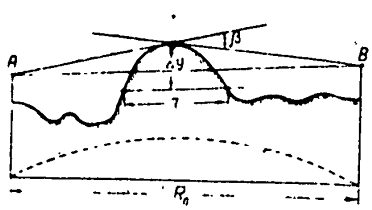

Fig.3

$$\beta = -\frac{H}{R_0 \kappa (1-\kappa)}, \qquad (25)$$

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the value of V also decreases exponentially at a decrease in clearance H. Consequently, the relationship between attenuation multiplier V, expressed in decibels, and the value of clearance at $H \leq H_o$, can be considered linear. At a given value of H the value of V can be determined by this simple formula

$$V_{db} = V_{odb}\left(1 - \frac{H}{H_0}\right), \qquad (26)$$

where V_{db} is value of attenuation multiplier in decibels

H is value of clearance ($H \leq H_o$), which is determined directly from route profile

H_o is value of clearance at which the value of attenuation multiplier equal to unity (zero decibels) is determined by eq.(1)

V_{odb} is the value of attenuation multiplier in decibels at H = 0, i.e., in the case when the line of sight is tangential to the highest point on route profile

A paper by V.A.Fok (Bibl.2) demonstrates that at sufficiently short waves and great antenna heights the attenuation multiplier is not dependent on the electrical parameters of the wave-bending sphere, and is determined by the following two parameters:

$$\xi = \sqrt[3]{\frac{\pi b}{\lambda}} \sin \beta, \qquad (27)$$

$$\mu = \sqrt[12]{\frac{8\pi^3}{b\lambda^3}} \sqrt{\frac{\sqrt{h_1 h_2}}{\sqrt{h_1} + \sqrt{h_2}}}, \qquad (28)$$

where b is radius of the sphere around which diffraction occurs

h_1 and h_2 are the heights of the transmitting and receiving antennas, respectively, above the surface of this sphere.

Considering that the value of V_o is determined at H = 0, ξ = 0, and V_o depends only on the parameter μ. The graph of the relationship $V_{o\,db} = f(\mu)$, cited from

Bibl.6, is depicted in Fig.4.

From Fig.4 it can be seen that at large values of μ (small values of b, short waves, and large values of heights h_1 and h_2) the curve $V_{o\ db} = f(\mu)$ asymptotically approximates - 6 db, which corresponds to transition to Fresnel's optical diffraction.

Antenna heights h_1 and h_2 are related to the values H, k, R_o and b by the following formulas:

$$\left.\begin{aligned} h_1 &= \frac{\kappa}{1-\kappa}H + \frac{\kappa^2 R_0^2}{2b} \\ h_2 &= \frac{1-\kappa}{\kappa}H + \frac{(1-\kappa)^2 R_0^2}{2b} \end{aligned}\right\} \qquad (29)$$

On substituting (29) into eq.(28) and considering (10), we obtain the formula for μ

$$\mu = \sqrt[6]{\frac{64\pi a^2}{3}}\sqrt[3]{\frac{\kappa^2(1-\kappa)^2}{l^2}}\sqrt[4]{1+\frac{l^2}{4\alpha\kappa(1-\kappa)}\frac{H}{H_0}}, \qquad (30)$$

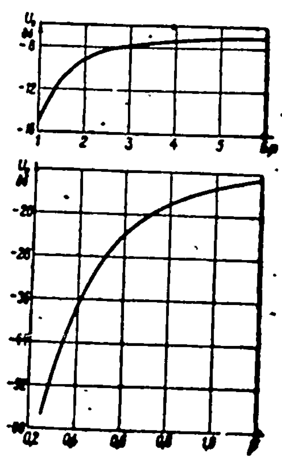

Fig.4

where

$$l = \frac{r}{R_0}, \qquad (31)$$

r is distance between the points of intersection of the obstacle with the line parallel to line of sight AB and spaced by value of Δy apart from the summit of obstacle (Fig.3),

$$\alpha = \frac{\Delta y}{H_0}; \qquad (32)$$

The value of μ is obtained sufficiently high in the range of decimetric and centimetric waves, and from Fig.4 it can be seen that at high μ the value of V_o is not very critical in relation to the value of μ. Moreover, the ratio $\frac{H}{H_o}$ is usually not very high and therefore the last multiplier in eq.(30) can be considered approximately equal to unity, and

$$\mu \approx \sqrt[6]{\frac{64\pi a^2}{3}} \cdot \sqrt[3]{\frac{\kappa^2(1-\kappa)^2}{l^2}}. \qquad (33)$$

If, for instance, we proceed from the assumption that the value of Δ_y equals H_o ($\alpha = 1$), eq.(33) will acquire this form

$$\mu \approx 2\sqrt[3]{\frac{\kappa^2(1-\kappa)^2}{l^2}}. \qquad (34)$$

3. Let us make a few remarks regarding the determination of the radius of approximating sphere for obstacles of a complex form.

We use the abovedescribed method to determine from the route profile the radius of the circumference b by which the cross-sectional area of obstacle is approximated by a plane passing through the Earth's center and the points of transmission and reception and not, of course, to determine the radius of the approximating sphere. If the form of an obstacle strongly differs from the spherical form, the radius of curvature of the obstacle on the perpendicular plane passing through the center of the Earth and the summit of obstacle may differ considerably from the value of b. However, the value of field strength is practically unaffected by the radius of the curvature of obstacle on the perpendicular plane when the extent of that radius is small in comparison with the extent of b. This is attributable to the small dimension of the essential zone in the ultrashort wave range, in the direction perpendicular to the route.

If two or more obstacles exist on a route and are in some proximity to each other, in the computations of the attenuation multiplier such obstacles can be regarded as a single equivalent obstacle upon determining the value of r as shown in Fig.5.

If two obstacles are spaced considerably far apart, they cannot be regarded as a single equivalent obstacle, and it is necessary to investigate separately the diffraction around each obstacle. It can be apparently assumed that in this case the

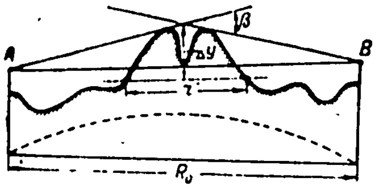

Fig.5

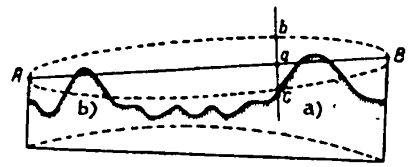

Fig.6

a) First obstacle; b) Second obstacle

value of the attenuation multiplier equals the sum of the attenuations (in decibels) introduced by every individual obstacle, when more than one obstacles are assumed to exist. The qualitative justification of this method can be provided on the basis of the following reasoning: Let transmitting antenna be at point A (Fig.6) and receiving antenna at point B. The ellipse in Fig.6 represents the cross-sectional area of space by a plane passing through the transmission and reception points and the center of the Earth. Let us assume in the beginning that there is only one obstacle on the route, and that it is in the proximity of transmission point B. In this case the obstacle is exposed to a spherical wave with a definite distribution of amplitudes and phases on the plane passing through points a, b, and c on the perpendicular plane of the diagram. The presence of a second obstacle will cause a change in the distribution of amplitudes and phases on the said plane, this change being the greater the nearer is the second obstacle to the first. It is obvious that the two obstacles can be considered separately only in the event when the distance between them is such as to cause the changes in amplitudes and phases, stemming from the presence of the second obstacle, at points b and c to be small compared with the change at point a.

It is possible to determine the minimum distance between two obstacles at which they can be considered separately, either by a rigorous solution of the problem of the diffraction of radio waves around two bodies of a definite geometrical form (for example, two spheres, two cylinders, etc.) or by comparing the computing results obtained by approximate methods with the results obtained experimentally.

4. It is known that when a terrain with obstacles in its front and rear is a level terrain, then interference phenomenons can be observed in the shadow zone owing to the overradiating effect of the summit of obstacle. The values of interference multipliers can be determined on taking into account the specific features of terrain relief, analogously to the procedure applied to open routes.

4. Consideration of Refraction

1. In all the aforecited formulas refraction was not considered in the computation of the attenuation multiplier, and it was assumed that all geometrical features of a route are determinable from its profile, plotted without considering refraction.

At linear variation of the permittivity of the air with altitude the consideration of refraction is effected by introducing the vertical gradient of permittivity of the air g, into rating formulas. The value of g is generally assumed to be negative when the permittivity of the air ε decreases with altitude and positive when ε increases.

2. Refraction can be taken into account by replacing the values of clearance H in all aforecited formulas with the value of H(g), which equals

$$H(g) = H - \frac{R_0^2}{4} g\kappa(1-\kappa), \qquad (35)$$

where g is vertical gradient of permittivity of the air

H is value of clearance at absence of refraction.

If $g < 0$, value of clearance H increases. At $g > 0$ value of clearance H(g) decreases.

It can be concluded from eq.(35) that the maximum change in clearance at a change in g takes place in the middle of route (k = 0.5).

3. A change in g leads to a change not only in the value of clearance but also in the coordinates of the reflection points and in the value of the radius of approximating sphere. However, it can be concluded from eqs.(6), (26), and (28) that the value of the attenuation multiplier is principally affected by changes in the value of clearance, and therefore changes in the coordinates of reflection points and in the radius of approximating sphere can be ignored.

4. A change in g owing to change in clearance on open routes involves a change in the travel difference between direct and reflected wave, and thus the reception point may achieve interference minimums and maximums of the field. On routes with a single reflection point the interference maximums occur at values of $g = g_m$ determinable by the following formula

$$g_m = 4\left[\frac{H}{R_0^2 \kappa(1-\kappa)} - \sqrt{\frac{(2m-1)\lambda}{R_0^3 \kappa(1-\kappa)}}\right], \qquad (36)$$

where m = 1, 2, 3 . . . is the number of the interference maximum and H is the value of clearance at reflection point with relative coordinate k at g = 0.

Interference minimums will occur at values of $g = g_n$ equal to

$$g_n = 4\left[\frac{H}{R_0^2 \kappa(1-\kappa)} - \sqrt{\frac{2n\lambda}{R_0^3 \kappa(1-\kappa)}}\right], \qquad (37)$$

where n = 1, 2, 3 . . . is the number of the interference minimum.

5. Limits of Applicability of Approximate Methods

At low values of clearance the interference formulas (3) and (5) yield incorrect values of the attenuation multiplier. On the other hand, at high positive values of clearance when the field has an interference character, the diffraction formula (26) also yields incorrect (exorbitant) values of the attenuation multiplier, and therefore it is important to know the limits of applicability of the abovedescribed approximate methods of determining the attenuation multiplier.

The diffraction formula (26) can be presented in this form

$$V_{db} = V_{0db}(\mu)(1 - m), \tag{38}$$

where

$$m = \frac{H}{H_0}, \tag{39}$$

and μ is determined by eqs.(30) or (33).

If we are dealing with a route with a single reflection point and if the reflection takes place on a smooth convex surface, then interference formula (5) can be thus written

$$V = \sqrt{1 + |\Phi|_{pl}^2 D^2 + 2|\Phi|_{pl} D\cos\gamma}. \tag{40}$$

Considering that angles of slide θ are usually small, $|\Phi|_{pl}$ can be usually regarded as equal to unity, and the phase of reflection factor can be regarded as being at 180°. Furthermore, on the basis of eqs.(33) and (39), the formula for the reflection factor can be reduced to this form

$$D \approx \frac{1}{\sqrt{1+3,9\frac{\mu^2}{m}}} = D(m, \mu). \tag{41}$$

On, lastly, considering eqs.(1), (16), and (39), eq.(40) can be thus rewritten

$$V = \sqrt{1+D^2(m,\mu) - 2D(m,\mu)\cos\left(\frac{\pi m^2}{3}\right)}. \tag{42}$$

By designating as m_o the value of m at which eqs.(38) and (42) yield identical values of attenuation multiplier, we obtain the following transcendental equation for determining m_o

$$V_{0db}(\mu)(1 - m) = 10\lg\left[1 + D^2(m,\mu) - 2D(m,\mu)\cos\left(\frac{\pi m^2}{3}\right)\right]. \tag{43}$$

which can be derived graphically. Graphical derivation demonstrates that at the actually encountered conditions of μ the value of m_o approximates unity.

Thus, the diffraction formula is to be employed when clearance $H \leq H_o$, and interference formulas have to be employed when $H > H_o$.

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THE APPLICATION OF FICTITIOUS MAGNETIC CURRENT FOR SOLVING THE PROBLEM OF ANTENNA RADIATION OVER A PLANE WITH NONHOMOGENEOUS LEONTOVICH BOUNDARY CONDITIONS

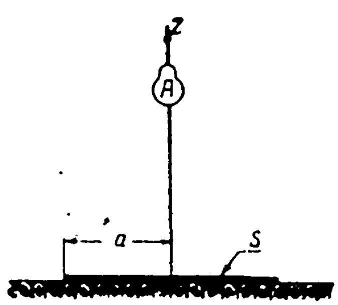
by

O.N.Tereshin

Description of a method of solving the problem of the radiation of a system of side currents over a plane with nonhomogeneous Leontovich boundary conditions.

1. Solution of the Problem

A system of currents in area A exists over surface z = 0 (see diagram). The part of surface directly under the antenna is metallized within the confines of area S. Outside the confines of this area, Leontovich boundary conditions apply on surface z = 0



$$\frac{E_{tg}}{H_{tg}} = \alpha \qquad (1)$$

Here the subscript tg denotes tangential components.

The problem is solved on considering the the following limitations:

1. Current distribution in antenna A has a circular symmetry in relation to axis z.

2. Area S represents a circle with a diameter of 2 a. The beginning of the cylindrical system of coordinates lies at the same point as the center of the circle.

3. The field created by antenna A in free space has a wave structure of the

TE type. This limitation is not important, because the problem can be solved completely analogously for the case of TE waves from primary antenna A.

Despite the above-introduced limitations the analysis of many metallized-system antennas with such mounting can be reduced to a problem of electrodynamics.

In solving the here posed problem we shall use the formulas obtained in G.T.Markov's paper (Bibl.1). Let us separate the process of solution into two stages:

1. First we will solve the problem of the radiation of the here-examined system of currents over a plane with uniform threshold conditions (1), which, at the limitations adopted, have this form

$$\frac{E_r}{H} = \alpha \tag{2}$$

In accordance with Bibl.1 the rectangular component of the electrical field strength for the given system of currents in free space will be written as follows

$$E_{z\,\text{primary}} = \int\limits_V dV \sum_{n=-\infty}^{\infty} \int\limits_{x=-\infty}^{\infty} E_{znx} dx, \tag{3}$$

where

$$E_{znx} = \frac{1}{S\pi} e^{\pm\sqrt{x^2-\kappa^2}(z'-z)} e^{in(\varphi-\varphi')} \frac{x}{\sqrt{x^2-\kappa^2}} \begin{cases} J_n(xr) F_1^{\vartheta}(x) & \text{for } r < r' \\ H_n(xr) F_2^{\vartheta}(x) & \text{for } r > r' \end{cases},$$

$$F^{\vartheta}_{\{{1 \atop 2}\}} = \frac{x^2}{i\omega\varepsilon} j_z^{\vartheta} \begin{Bmatrix} H_n^{(2)}(xr') \\ J_n(xr') \end{Bmatrix} - \left(\frac{\pm\sqrt{x^2-\kappa^2}}{i\omega\varepsilon} j_\varphi^{\vartheta} - j_r^{M} \right) \frac{in}{r'} \begin{Bmatrix} H_n^{(2)}(xr') \\ J_n(xr') \end{Bmatrix} +$$

$$+ \left(\frac{\pm\sqrt{x^2-\kappa^2}}{i\omega\varepsilon} j_r^{\vartheta} + j_\varphi^{M} \right) x \begin{Bmatrix} H_n^{(2)\prime}(xr') \\ J_n'(xr') \end{Bmatrix}.$$

Here j_z, j_r, and j_φ stand for the components of vectors of the density of the current set for area A; r, φ and z stand for the coordinates of observation points; and r', φ' and z' stand for coordinates of the points of current distribution.

If it be taken into account that

$$H_n^{(2)}(-x) = e^{i(n+1)\pi} H_n^{(1)}(x),$$
$$H_n^{(2)}(x) + H_n^{(1)}(x) = 2J_n(x),$$

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eq.(3) can be rewritten in another form

$$E_{z\,primary} = \int_V dV \sum_{n=-\infty}^{\infty} \int_{x=0}^{\infty} E'_{znx} dx, \qquad (4)$$

where

$$E'_{znx} = \frac{1}{4\pi} e^{\pm\sqrt{x^2-\kappa^2}(z'-z) - in(\varphi'-\varphi)} \frac{x}{\sqrt{x^2-\kappa^2}} J_n(xr) F'(x),$$

$$F'(x) = -x^2 j_z^э J_n(xr') - \left(\frac{\pm\sqrt{x^2-\kappa^2}}{i\omega\varepsilon} j_\varphi^э - j_r^м\right) \frac{in}{r'} J_n(xr') +$$

$$+ \left(\frac{\pm\sqrt{x^2-\kappa^2}}{i\omega\varepsilon} j_r^э + j_\varphi^м\right) x J'_n(xr).$$

Considering that we confined ourself to φ symmetrical current distributions, the sum by n vanishes in eqs.(4)

$$\underset{at\ (z'-z)>0}{E_{z\,primary}} = \frac{1}{2} \int_0^{\infty} F(x) \frac{x}{\sqrt{x^2-\kappa^2}} e^{-\sqrt{x^2-\kappa^2}(-z)} J_0(xr)\, dx. \qquad (5)$$

Here

$$F(x) = \int_{z'=-\infty}^{\infty} \int_{r'=0}^{\infty} e^{-\sqrt{x^2-\kappa^2}z'} \left\{\frac{x^2}{i\omega\varepsilon} j_r^э J_0(xr') - \right.$$

$$\left. - \left[\frac{-\sqrt{x^2-\kappa^2}}{i\omega\varepsilon} j_z^э + j_\varphi^м\right] x J_1(xr')\right\} r'dr'dz'. \qquad (6)$$

We will seek for the field reflected from boundary z = 0 thus

$$\underset{at\ z>0}{E_{z\,secondary}} = \frac{1}{2} \int_0^{\infty} f(x)\, e^{+\sqrt{x^2-\kappa^2}(-z)} J_0(xr)\, dx. \qquad (7)$$

Here f(x) is the sought-for function, which is a part of the spectral density of the Fourier-Bessel dissociation for the secondary field. Let us write the components of the electromagnetic field which are included in conditions (2). In this connection, we will employ the formulas derived in Bibl.1.

$$\left.\begin{aligned}
E_{r\,\text{primary}\atop \text{at } z=0} &= -\frac{1}{2}\int_0^\infty F(x)\,J_1(xr)\,dx,\\
H_{\varphi\,\text{primary}\atop \text{at } z=0} &= \frac{i\,\omega\varepsilon}{2}\int_0^\infty \frac{F(x)}{\sqrt{x^2-\kappa^2}}\,J_1(xr)\,dx,\\
E_{r\,\text{secondary}\atop \text{at } z=0} &= \frac{1}{2}\int_0^\infty \frac{f(x)\sqrt{x^2-\kappa^2}}{x^2}\,J_1(xr)\,xdx,\\
H_{\varphi\,\text{secondary}\atop \text{at } z=0} &= \frac{i\,\omega\varepsilon}{2}\int_0^\infty \frac{f(x)}{x^2}\,J_1(xr)\,xdx.
\end{aligned}\right\} \tag{8}$$

In accordance with condition (2) the full field on the boundary z = 0 should satisfy the following equation

$$\int_0^\infty\left[f(x)\frac{\sqrt{x^2-\kappa^2}}{x^2}-\frac{F(x)}{x}\right]J_1(xr)\,xdx = \\ = i\,\omega\varepsilon\alpha\int_0^\infty\left[\frac{f(x)}{x^2}+\frac{F(x)}{x\sqrt{x^2-\kappa^2}}\right]J_1(xr)\,xdx. \tag{9}$$

Inasmuch as the Fourier-Bessel transforms are mutually synonymous, condition (9) is reduced to the following equation

$$\frac{f(x)\sqrt{x^2-\kappa^2}}{x^2}-\frac{F(x)}{x} = i\,\omega\varepsilon\alpha\left[\frac{f(x)}{x^2}+\frac{F(x)}{x\sqrt{x^2-\kappa^2}}\right],$$

whence, on introducing the designation $i\,\omega\varepsilon\alpha = B$, we will find

$$f(x) = F(x)\,\frac{x}{\sqrt{x^2-\kappa^2}}\,\frac{(\sqrt{x^2-\kappa^2}+B)}{(\sqrt{x^2-\kappa^2}-B)}. \tag{10}$$

On substituting (10) into (8) we obtain the formulas for the full field on the plane of z = 0:

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$$\left. \begin{aligned} E_{r\ full} &= \int_0^\infty \frac{BF(x)}{\sqrt{x^2-\kappa^2}-B} J_1(xr)\,dx \\ H_{\varphi\ full} &= i\omega\varepsilon \int_0^\infty \frac{F(x)}{\sqrt{x^2-\kappa^2}-B} J_1(xr)\,dx \end{aligned} \right\} \qquad (11)$$

2. Let us now investigate an instance when only magnetic currents j_φ^M, distributed symmetrically by the azimuth, are set over a plane with homogeneous boundary conditions (2). In this case the solution will be again yielded by eq.(10), because this case represents a specific instance of the afore-described problem. However, then $F_M(x)$, which is determined by eq.(6), has the following form

$$F_M(x) = -\int_{z'=-\infty}^{\infty} \int_{r'=0}^{\infty} e^{\pm\sqrt{x^2-\kappa^2}z'} j_\varphi^M(z';\ r')\,x\,J_1(xr')\,r'dr'dz'. \qquad (12)$$

If the factor given is the surface magnetic current lying on the dividing boundary, the density of this surface current is related to the volume density of magnetic current by the following formula

$$j_\varphi^M(z';\ r') = I_\varphi^M(0;\ r')\,\delta(z'-0),$$

where $I_\varphi^M(0;\ r')$ is the surface density of current; $\delta(z'-0)$ is the delta function. Therefore eq.(12) can be thus rewritten

$$F_M(x) = -\int_{r'=0}^{\infty} I_\varphi^M(r')\,x\,J_1(xr')\,r'dr'. \qquad (13)$$

It can be concluded from eqs.(10), (5), and (7) that at $z > z'$ the field for the given magnetic surface current will be expressed by the following formulas:

$$\left. \underset{at\ z>0}{E_r^M} = \int_0^\infty \frac{F_M(x)\sqrt{x^2-\kappa^2}}{(\sqrt{x^2-\kappa^2}-B)} e^{-\sqrt{x^2-\kappa^2}\,z} J_1(xr)\,dx \quad \right| \qquad (14)$$

$$\underset{\text{at } z>0}{H^{M}_{\varphi}} = i\omega\varepsilon \int_{0}^{\infty} \frac{F_{M}(x)}{(\sqrt{x^2-\kappa^2}-B)} e^{-\sqrt{x^2-\kappa^2}\,z} J_1(xr)\,dx$$

In turn, we shall obtain the following formula from eq.(14) at the limit involving z = + 0 (at proximity of positive z's to surface z = 0)

$$\underset{\text{at } z=+0}{E^{M}_{r}} = \int_{0}^{\infty} \frac{F_{M}(x)\sqrt{x^2-\kappa^2}}{(\sqrt{x^2-\kappa^2}-B)} J_1(xr)\,dx. \tag{15}$$

Let us analyze eq.(15). Here

$$\frac{F_{M}(x)\sqrt{x^2-\kappa^2}}{(\sqrt{x^2-\kappa^2}-B)} = F_{M}(x) + \frac{BF_{M}(x)}{(\sqrt{x^2-\kappa^2}-B)}. \tag{16}$$

Considering this, let us find, upon taking into account (13)

$$\underset{\text{at } z=+0}{E^{M}_{r}} = -\int_{0}^{\infty}\left[\int_{0}^{\infty} I^{M}_{\varphi}(r')\,x\,J_1(xr')\,r'dr'\right] J_1(xr)\,dx +$$

$$+\int_{0}^{\infty} \frac{B\,F_{M}(x)}{\sqrt{x^2-\kappa^2}-B} J_1(xr)\,dx = -I^{M}_{\varphi}(r) + \int_{0}^{\infty} \frac{BF_{M}(x)}{\sqrt{x^2-\kappa^2}-B} J_1(xr)\,dx.$$

The first item in the above-obtained formula corresponds to an instance when a magnetic-current sheet spreads over an infinitely conducting plane of boundless dimensions (instance B = 0). The second item owes its appearance to the fact that B is necessary in the instance cited. Considering that the introduction of boundary conditions (1) is based on the assumption that B is low, we can justifiably expect that the magnitude of E_r at the surface of magnetic current will be chiefly determined by the first item in eq.(16). Hence a wholly comprehensible physical order of the construction of successive approximations for solving the problem posed. On regarding the parameter B as low, we can assume in the first approximation that

$$E^{M}_{r_{full}} = -I^{M}_{\varphi}(r). \tag{17}$$

Let us set such a magnetic current distribution as to cause the following equation to be valid within the limits of r = 0 to r = a on surface z = 0

$$E_{r\,full} + E^{M}_{r\,full} = 0, \tag{18}$$

where the first item is determined by eq.(11) and the second by eq.(15). By the same token, in the first approximation, an infinitely conducting shield is formed on surface z = 0 within the limits of r = 0 to r = a, and the boundary condition $E_{tg} = 0$ is satisfied on its surface. Outside this area the field components expressed by eqs.(11) and (14) satisfy separately the boundary conditions (2), whence it follows that the full field which is their sum also satisfies these boundary conditions. Thus, we obtain from eq.(18), upon considering eqs.(17) and (11), the following formula

$$\underset{0<r'<a}{I^{M}_{\varphi}}(r') = \int_0^\infty \frac{B\,F(x)}{\sqrt{x^2-\kappa^2}-B} J_1(xr')\,dx. \tag{19}$$

On substituting eq.(19) into eq.(13) and integrating by r', we will find

$$F_M(x) = -xB\int_0^\infty \frac{F(\xi)}{\sqrt{\xi^2-\kappa^2}-B}\cdot\frac{xa\,J_1(\xi a)\,J_0(xa) - \xi a\,J_0(\xi a)\,J_1(xa)}{\xi^2 - x^2}\,d\xi. \tag{20}$$

On further substituting eq.(20) into eq.(14) and on adding to this formula the field of the set antenna A over a surface with homogeneous boundary conditions (2) we will obtain in the first approximation the field of this antenna over a surface with nonhomogeneous boundary conditions.

The second approximation for solving the above problem can be thus determined. If surface magnetic current be given according to eq.(17) then, in accordance with eq.(16), the following field will prove to be not compensated on surface z = 0 within the range from r = 0 to r = a

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$$E^M_{r \text{ at } z \to +0} = \int_0^\infty \frac{F_M(x)\,B}{\sqrt{x^2-\kappa^2}-B} J_1(xr)\,dx, \tag{21}$$

where $F_M(x)$ is determined from eq.(20). The remaining field (21) can be compensated by again superimposing on surface z = 0, within the area of the shield, a magnetic current with surface density as follows

$$J^M_r(r) = \int_0^\infty \frac{B\,F_M(x)}{\sqrt{x^2-\kappa^2}-B} J_1(xr)\,dx.$$

From eq.(20) it can be concluded that for such magnetic field

$$F'_M(x) = -\int_0^\infty \frac{B\,F_M(\xi)}{\sqrt{\xi^2-\kappa^2}-B} \; \frac{xa\,J_1(\xi a)\,J_0(xa) - \xi a\,J_0(\xi a)\,J_1(xa)}{\xi^2-x^2} d\xi.$$

Considering that here $F_M(\xi)$ is determined by eq.(20), we obtain

$$F'_M(x) = xB \int_0^\infty \frac{\left[\xi B \int_0^\infty \frac{F(\eta)}{\sqrt{\eta^2-\kappa^2}-B} \frac{\xi a\,J_1(\eta a)\,J_0(\xi a) - \eta a\,J_0(\eta a)\,J_1(\xi a)}{\eta^2-\xi^2}\,d\eta\right]}{\sqrt{\xi^2-\kappa^2}-B} \times$$
$$\times \frac{xa\,J_1(\xi a)\,J_0(xa) - \xi a\,J_0(\xi a)\,J_1(xa)}{\xi^2-x^2}\,d\xi.$$

The subsequent approximations can be formulated according to the same schematic. As a result, the solution for the field of the given system of currents over a surface with nonhomogeneous Leontovich boundary conditions will be obtained in the form of the following series of successive approximations:

$$E = \int_0^\infty \frac{\sqrt{x^2-\kappa^2}}{x} \frac{e^{-\sqrt{x^2-\kappa^2}\,z}}{(\sqrt{x^2-\kappa^2}-B)} \left\{ \frac{F_1(x)(\sqrt{x^2-\kappa^2}-B) + F(x)(\sqrt{x^2-\kappa^2}+B)}{2\sqrt{x^2-\kappa^2}} - \right.$$
$$- xB\int_0^\infty F(\xi)\,\Phi(x,\xi)\,d\xi + xB\int_0^\infty \xi B\left[\int_0^\infty F(\eta)\,\Phi(\xi;\eta)\,d\eta\right]\Phi(x,\xi)\,d\xi -$$
$$- xB\int_0^\infty \xi B\left[\int_0^\infty \eta B\left(\int_0^\infty F(\psi)\,\Phi(\eta,\psi)\,d\psi\right)\Phi(\xi,\eta)\,d\eta\right]\Phi(x,\xi)\,d\xi +$$
$$\left. + \ldots \right\} J_1(xr)\,x\,dx. \tag{22}$$

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Here $\phi(x,\xi) = \dfrac{xaJ_1(\xi a)J_0(xa)-\xi aJ_0(\xi a)J_1(xa)}{(\sqrt{\xi^2-k^2}-B)(\xi^2-x^2)}$, $F(x)$ and $F_1(x)$ are determined by eq.(4), but, according to Bibl.1, the sign for $F_1(x)$ should be taken " + " before the radical.

2. Analysis of Computing Results

The obtained formulas were employed to conduct calculations of radiation patterns for the case of the radiation of an antenna having the form of a quarterwave dipole located over a surface with nonhomogeneous boundary conditions. At field calculations for the distant zone, the method of drawing out was applied to eq.(22) so that the external integral could be taken in eq.(22).

The internal integrals were computed by the method of numerical integration. In this connection, it was found that the function $\phi(x;\xi)$ has a sharply expressed filtering character at large parameters of disk ($ka \geq 5$), so that integration was conducted on a small segment around the point $x = \xi$. This facilitates the calculations considerably and makes it possible to obtain a solution at the third or fourth approximation without any special difficulties.

The analysis of the obtained computing results demonstrates that:(1) the series of successive approximation converges very satisfactorily (as the geometrical progression of 0.5^n, and even better) even at considerable diameters of disks ($ka \approx$ ≈ 20 to 30) at mean values of the parameter of smallness B ($B = -ik\,0.2$); and (2) on the introduction of metallization on surface $z = 0$ under the antenna, the field strength in the distant zone increases several times, at mean parameters of the Earth, as compared with the case of installing the same antenna over a surface with homogeneous Leontovich boundary conditions for all values of the elevation angle (at $ka \approx 20$ approximately twice). This coincides with the results obtained in E.L.Feynberg's work (Bibl.2).

The computing results were experimentally tested and corroborated.

3. Conclusions

In concluding it is to be noted that, as employed in the present paper, the method of imposing a fictitious surface magnetic current is in many ways analogous to the method of the double-layer potential which is widely applied in electrostatics. However, the imposition of magnetic current on the boundary between two media makes it possible to obtain, on isolating the external-integral member in eq.(16), the parameter of smallness B in the capacity of the multiplier before the integral member. This parameter B is related to boundary conditions on the surface of the other medium. This feature makes possible the rapid convergence of the obtained series of successive approximations, whose members have as their multiplier the parameters of smallness in degrees commensurate with the order of approximation.

It is easy to ascertain that if the problem posed here be solved by the conventional method of imposing a surface electric current, the series of the thus obtained successive approximations will lack the abovementioned feature.

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SOME BASIC CONCEPTS OF THE SIGNAL THEORY

by

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An attempt at defining some basic concepts of the signal theory as according to time-averaging operations; and an attempt to explain that correspondingly certain definite characteristics - the correlation function and the spectrum - are subordinated to laws applicable to a class of uniform signals that is broader than the class of stationary - in the probabilitative sense - processes. An example is provided by reviewing the autocorrelation function and the AM-signal spectrum. Indication of the necessity of distinguishing between the physical and statistical signal spectra.

The signal theory, which is a fundamental branch of the general theory of communications, is extensively based on the concepts and methods of the theory of random functions. It is to be noted that in signal transmission practice we have more often than not to deal with the statistical characteristics of the signal as a partial realization of the random function of time, whereas the random process theory deals with an assembly of function representing all the possible realizations of the process of this or that statistical type. In theory, communication engineering is concerned with computing and noting only the time-average values of the magnitudes pertaining to a received signal, whereas the random function theory virtually always interprets average values as those that are average for the assembly of functions. The opportunities for utilizing the concepts and methods of the random function theory for solving problems of the general theory of communications are therefore confined to the class of the so-called (broadly interpreted) stationary processes whose

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time and function-assembly averages coincide with an accuracy of up to any small value according to the ergodic theorem, upon satisfying some additional requirements.

Such a situation involves difficulties of the twofold kind. First, no information pertaining to one sample (or several samples) of a signal of this or that type suffices for judging whether the given signal is stationary or not; therefore, the reference to the ergodic property is in most cases a convenient but hardly well-founded hypothesis. Second, some types of the signals employed in communication engineering do not belong in the stationary groups (for instance, AM signals), as is known.

However, it can be demonstrated that the major actual concepts of the signal theory can be defined by means of time-averaging operations applied to processes in a form quite independent of the groups or assemblies to which these processes belong. Here, it is found that the laws governing the correlation functions of stationary random functions acquire a rational meaning in relation to a broader class of processes. At the same time, this will explain the symptoms by which we can judge whether a given signal does or does not belong in that class.

The signal theory is usually concerned with the following time-average values related to moment t and determined on the interval $t - \frac{T}{2} \leq \xi \leq t + \frac{T}{2}$:

$$M(t,T) = \frac{1}{T} \int_{t-\frac{T}{2}}^{t+\frac{T}{2}} f(\xi)\, d\xi, \qquad (1)$$

$$P(t,T) = \frac{1}{T} \int_{t-\frac{T}{2}}^{t+\frac{T}{2}} f^2(\xi)\, d\xi, \qquad (2)$$

$$R(\tau, t, T) = \frac{1}{T} \int_{t-\frac{T}{2}}^{t+\frac{T}{2}} f(\xi) f(\xi - \tau)\, d\xi. \qquad (3)$$

The first of the above values characterizes the average value of the signal; the second – the mean power of the signal; and finally the third – the degree of

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coherence of the signal and its delayed (by a time τ) repetition.

Let us term structurally uniform the signals whose mean values (1) to (3) tend to reach some definite limits (especially, zero) at an increase in T. In engineering practice we conclude, for example, the existence of a limit of function (3)

$$R(\tau) = \lim_{T \to \infty} \frac{1}{T} \int_{t-\frac{T}{2}}^{t+\frac{T}{2}} f(\xi) f(\xi - \tau)\, d\xi \qquad (4)$$

in the cases when, at an increase in T_o, the difference

$$\varepsilon = | R(\tau, T) - R(\tau, T_0) | \qquad (T_s > T > T_0)$$

decreases and may be made sufficiently small; in this connection, it is of course assumed that the duration of signal, T_s, always exceeds the averaging time T, because otherwise an increase in T beyond the limits of the presence of the signal could not yield any results. On selecting on the basis of some consideration a given value of $\varepsilon = \varepsilon_o$ characterizing the allowable error in the power measurement we can assume that the signal is quite structurally uniform if there exists a minimum averaging time T_o at which the requirement

$$\varepsilon < \varepsilon_0 \quad \text{at} \quad T > T_0 \qquad (5)$$

is satisfied for any value of τ. If this requirement is observed, the limit value of mean power is present also, because P = R (0); as for the mean value (1), it is determined by the magnitude of the constant component of the signal, which is virtually always known and in most cases equals zero.

Attention should be also turned to another essential circumstance. Experience indicates that the signals which are uniform in the sense of requirement (5) are actually present, but that time T_o (let us term it "limit of uniformity") can have a value that varies within very wide boundaries as depending on the type of signal. The limit of uniformity T_o constitutes a quantitative characteristic of the signal

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which must be absolutely known prior to the solving of a number of engineering problems. However, this characteristic is not (and cannot be) included in the concept of the stationary process as defined by the random function theory.

A principal characteristic of structurally uniform signals is the circumstance that the limit values P and R (τ) with an accuracy of up to a small value of ε_0 are not affected by time, that is, by the section of a signal with a duration of $T \geq T_0$ which is selected for investigation. This is a consequence of the fact that any sufficiently long section of a uniform signal does not differ from its other sections in those properties of its temporal structure which determine the energy effect when the signal is combined with its delayed repetition; in a case to the contrary, any increase in averaging time T should affect the value of R (τ) and the mean power P of the signal.

The function R (τ), which is determined for uniform signals from eq.(4), should be termed the autocoherence function, in accordance with its physical meaning. However, it is to be kept in mind that the communication theory has adopted quite durably the term "autocorrelation function" which it borrowed from the random process theory, although, generally speaking, this term is interpreted differently by the latter theory, (see below). The autocoherence and autocorrection functions coincide mutually only with regard to the stationary processes having the ergodic property. In view of the terminology adopted in the theory of communications, the term "autocorrelation function" can be supported by definition (4) in those cases, at least, when any confusion as to the method of definition is absent.

In an analogous manner, time averaging can be used for determining the mutual-correlation (or to be more precise, mutual coherence) function of two different uniform signals $f_1(t)$ and $f_2(t)$, of which one has generally some temporal displacement τ in relation to the other

$$R_{12}(\tau) = \lim_{T \to \infty} \frac{1}{T} \int_{t-\frac{T}{2}}^{t+\frac{T}{2}} f_1(\xi) f_2(\xi - \tau)\, d\xi \qquad (6)$$

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If signals $f_1(t)$ and $f_2(t)$ are uniform, the sum signal

$$f(t)=f_1(t)+f_2(t-\tau)$$

also is uniform in the sense of requirement (5). In effect, the mean power of signal f (t) is

$$P=\lim_{T\to\infty}\frac{1}{T}\int_{t-\frac{T}{2}}^{t+\frac{T}{2}}[f_1(\xi)+f_2(\xi-\tau)]^2d\xi=P_1+P_2+2R_{12}(\tau),$$

where P_1 and P_2 stand for the mean power of each of the signals being combined. Physical considerations make it clear that

$$|R_{12}|\leqslant\sqrt{P_1P_2};$$

therefore, limit (6) exists if the limit values of mean powers P_1 and P_2 exist, i.e., if the signals being combined are structurally uniform.

For instance, all periodic signals are uniform in the sense of requirement (5)

$$(t)=\sum_{\kappa=1}^{\infty}C_\kappa\cos(\kappa\omega t+\varphi_\kappa), \tag{7}$$

because a limit value exists for them

$$R(\tau)=\lim_{T\to\infty}R(\tau,T)=\sum_{\kappa=1}^{\infty}\frac{C_\kappa^2}{2}\cos\kappa\omega\tau. \tag{8}$$

Let us note, however, that if signals (7) be regarded as a group determinable by the probability characteristics of random amplitudes C_k and initial phases φ_k, such a group will not always be stationary and ergodic. AM signals of this form

$$f(t)=[A+\varphi(t)]\sin\Omega t \tag{9}$$

are also uniform, if the modulating signals $\varphi(t)$ are uniform. In effect, for AM signals

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$$R(\tau, T) = \frac{1}{2} r(\tau, T) \cos \Omega \tau -$$

$$- \frac{1}{2T} \left\{ \left[\int_{-\frac{T}{2}}^{\frac{T}{2}} F(t) \cos 2\Omega t \, dt \right] \cos \Omega \tau + \left[\int_{-\frac{T}{2}}^{\frac{T}{2}} F(t) \sin 2\Omega t \, dt \right] \sin \Omega \tau \right\},$$

where

$$r(\tau, T) = \frac{1}{T} \int_{-\frac{T}{2}}^{\frac{T}{2}} [A + \varphi(t)] [A + \varphi(t - \tau)] \, dt$$

and

$$F(t) = [A + \varphi(t)] [A + \varphi(t - \tau)].$$

On exploring the limit of the above formula we note that at an infinitely increasing T, the following values:

$$a(2\Omega) = \lim_{T \to \infty} \int_{-\frac{T}{2}}^{\frac{T}{2}} F(t) \cos 2\Omega t \, dt, \quad b(2\Omega) = \lim_{T \to \infty} \int_{-\frac{T}{2}}^{\frac{T}{2}} F(t) \sin 2\Omega t \, dt$$

determine the spectral density of the component amplitude with a frequency of 2Ω in the spectrum of the function F (t). Inasmuch as the carrier frequency Ω considerably exceeds the upper boundary of the spectrum of the modulating signal $\varphi(t)$, the upper boundary of the spectrum of function F (t) lies considerably below the frequency 2Ω; therefore, $a(2\Omega) = b(2\Omega) = 0$. Thus, there exists a limit value of the autocorrelation function

$$R(\tau) = \lim_{T \to \infty} R(\tau, T) = \frac{1}{2} r(\tau) \cos \Omega \tau = \frac{A^2}{2} \cos \Omega \tau + \frac{1}{2} r_{\varphi}(\tau) \cos \Omega \tau, \qquad (10)$$

where $r_{\varphi}(\tau)$ stands for the autocorrelation function of the modulating signal $\varphi(t)$ which is assumed to be structurally uniform. We can see that the AM signal proves

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itself to be uniform; it is also clear that the group of AM signals (9) corresponding to the group of modulating signals $\varphi(t)$ or, say, even the stationary group, is in no way stationary and displays no ergodic properties.

Random (nonperiodic) signals can be also structurally uniform. These signals are of fundamental interest to the communication theory, because only such random processes can convey a continuous flow of information. While the uniformity of periodic signals is determined by the multiple repetition of an identical time structure, the uniformity of nonperiodic signals stems from the time stability of certain statistical characteristics of that structure, and from their sameness on any sufficiently long signal section.

From the viewpoint of practical applications of the theory, an important fact is that uniform signals are subject to well known and often used theorems relating the autocoherence function $R(\tau)$ to the spectrum $\Phi(\nu)$ of mean signal power:

$$\Phi(\nu) = 4\int_0^\infty R(\tau)\cos\omega\tau\, d\tau, \tag{11}$$

$$R(\tau) = \int_0^\infty \Phi(\nu)\cos\omega\tau\, d\nu \tag{12}$$

$\omega = 2\pi\nu$). To derive these formula, it suffices to present the spectrum of function $R(\tau)$ in the form of

$$g(\omega) = \int_{-\infty}^{\infty} R(\tau)\, e^{i\omega\tau}\, d\tau$$

and to substitute here the definition of $R(\tau)$ from eq.(4). Thereupon

$$g(\omega) = \lim_{T\to\infty}\frac{1}{T}\int_{-\frac{T}{2}}^{\frac{T}{2}} d\tau \int_{-\frac{T}{2}}^{\frac{T}{2}} f(t)\, f(t-\tau)\, e^{i\omega\tau}\, dt;$$

on multiplying the formula under the integral by $e^{i\omega t}\cdot e^{-i\omega t} = 1$, and on introducing a new variable, $t' = t - \tau$, so that $dt' = -d\tau$, it is possible, on transposing

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the interval of integration T by the time axis (which is allowable if the signal is uniform), to obtain the following result

$$g(\omega)=\lim_{T\to\infty}\frac{S_T(\omega)\,S_T^*(\omega)}{T}=\lim_{T\to\infty}\frac{|S_T(\omega)|^2}{T},$$

where

$$S_T(\omega)=\int_{-\frac{T}{2}}^{\frac{T}{2}} f(t')\,e^{-i\omega t'}\,dt' \text{ and } S_T^*(\omega)=\int_{-\frac{T}{2}}^{\frac{T}{2}} f(t)\,e^{i\omega t}\,dt$$

stand for the coupled-complex values of the signal spectrum on the time segment from $-\frac{T}{2}$ to $\frac{T}{2}$. The subsequent transition to formulas (11) and (12) presents no difficulties if it be only considered that the independence of function R (τ) from time (i.e., the uniformity of the corresponding signal) ensures the parity of the autocoherence function.

Formulas (11) and (12) are subject to direct verification when applied to signals that are uniform although known to be nonstationary. Thus for instance, on substituting the autocoherence function (10) of an AM signal into (11), we obtain

$$\Phi(\nu)=2A^2\int_0^\infty \cos\Omega\tau\cos\omega\tau\,d\tau+2\int_0^\infty r_\varphi(\tau)\cos\Omega\tau\cos\omega\tau\,d\tau=\frac{A^2}{2}\delta(\omega-\Omega)+$$

$$+\int_0^\infty r_\varphi(\tau)\cos(\Omega+\omega)\tau\,d\tau+\int_0^\infty r_\varphi(\tau)\cos(\Omega-\omega)\tau\,d\tau,$$

where $\delta(\omega-\Omega)$ stands for Dirak's delta function. This result, as was to be expected, coincides with the generally known interpretation of the spectrum of the AM signal: The spectrum consists of a carrier frequency Ω with an effective value of $\frac{A^2}{2}$, on whose both sides there are sidebands with displaced spectra of the modulating signal

$$\Phi\left(\frac{\Omega}{2\pi}\pm\nu\right)=\int_0^\infty r_\varphi(\tau)\cos(\Omega\pm\omega)\tau\,d\tau.$$

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The considerations touched upon here necessitate a more precise definition of the concept of the spectrum of signal power, and a more distinct separation of the two different aspects of that concept.

Let us consider eq.(11) as the definition of the concept of spectral density of power; it is not difficult to observe that this concept can be endowed with totally different contents and meaning as depending on the definition of the function R (τ). On selecting the definition (4), i.e., the autocoherence function, we compute the spectrum of the time-average signal power P. This spectrum can be also found experimentally by investigating signal f (t) by means of a suitable spectral apparatus and by measuring a number of frequencies for sufficiently long time intervals that should at any rate be no shorter than the uniformity limit of the signal. Let us conditionally term this a physical spectrum, because it may be determined up to any desired degree of accuracy by means of physical (spectral) measurements. The concept of spectral density will have a totally different meaning if the autocorrelation function be interpreted from the viewpoint of the random process theory

$$R(\tau) = \iint x_1 x_2 w_2 (x_1, x_2, \tau) \, dx_1 \, dx_2$$

Here $w_2(x_1, x_2, \tau)$ stands for the twodimensional distribution, which depends, in the case of stationary signals, only on the value of the time interval τ between the correlated values of signals and not on its position relative to the time axis; the integration applies to any value x of the signals in the group. In this case $\psi(\nu)$ stands for some statistical characteristic of the group whose sense becomes clear if we represent the group by the Fourier integral

$$f(t) = \int_0^{\infty} [a(\nu) \cos 2\pi\nu t + b(\nu) \sin 2\pi\nu t] \, d\nu,$$

on considering the coefficients a (ν) and b (ν) as random functions; the power spectrum $\psi(\nu)$ stands for the group-average value of the randon magnitude $a^2(\nu) + b^2(\nu)$, in which connection

$$\int_0^\infty \Phi(\nu)\, d\nu = \int x^2 w_1(x)\, dx ,$$

where w_1 (x) stands for unidimensional distribution independent of time. Such a spectrum could be termed statistical, as distinguished from the physical spectrum. Naturally, it cannot be observed by means of any physically realizable spectroscope.

If a group of signals is stationary and ergodic, both definitions of the concept will lead to one and the same quantitative result, because the ergodic property ensures the equivalency of time-and group-averaging operations. However, as we have seen above, uniform signals with a wholly definite physical spectrum are not necessarily stationary, and in such cases there is no ergodic group and hence also no statistical spectrum. A typical example of such a case is the group (9): the existence of the physical spectrum of AM signals cannot be doubted, although the same cannot be said of their statistical spectrum.

The differentiation between the concepts of the physical and the statistical spectra apparently serves to eliminate the not infrequently arising misunderstandings and to gain a clearer comprehension of the term "statistical spectrum".

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SIMPLIFIED ANALYSIS OF THE CIRCUITS OF RADIO-FREQUENCY JUNCTION-TRANSISTOR OSCILLATORS WITH SELF-EXCITATION

by

P.D.Berestnev

This article derives simplified formulas for oscillated frequency and self-excitation conditions for two self-oscillator circuits (common-emitter and common-base) as linked to collector circuit. The coupling between the input and output circuits can be of the transformer, autotransformer or capacitance type.

Approximate formulas for oscillated frequency and self-excitation conditions of an oscillator can be obtained on proceeding from the assumption that a transistor represents an active linear four-terminal network with Y parameters. These parameters are easily measured and lead to an equivalent pi network of the transistor, similar to the equivalent circuit for a super-high-frequency electron tube. Moreover, the formula for the voltage amplification factor, as used below, is obtained in its simplest form when expressed in Y parameters.

Figure 1 depicts the circuit of a self-oscillator where Y_{ol} stands for admittance of oscillator load. Let us disrupt the feedback circuit at points aa, and let us load the feedback coil L_{feedb} with the input conductance of transistor, Y_{inp}, (Fig.2). We will solve the problem of the self-excitation of the circuit and of the frequency oscillated, by applying Nyquist's stability criterion in Fig.2.

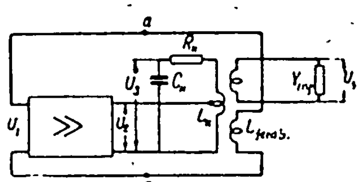

Fig.1

For the circuit in Fig.2 the voltage amplification factor is

$$\dot{K} = \frac{\dot{U}_4}{U_1} = |K| e^{i\varphi} = a + ib. \qquad (1)$$

The system is unstable if

$$|K| \geq 1 \text{ and } \varphi = 0,$$

or

$$a \geq 1 \text{ and } b = 0. \qquad (2)$$

Consequently, it is necessary to find the formula for the voltage amplification factor and to separate the real and the imaginary parts. On making the real part equal to unity it is possible to determine the conditions for self-excitation. On making the imaginary part equal to sero it is possible to find the oscillated frequency.

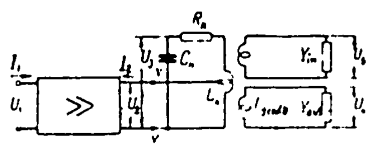

Fig.2

Before deriving a formula for the voltage amplification factor for the circuit in Fig.2, let us introduce the following designations:

circuit admittance

$$Y_{\kappa} = \frac{1}{i\omega L_{\kappa}} + \frac{C_{\kappa} R_{\kappa}}{L_{\kappa}} + i\omega C_{\kappa}, \qquad (3)$$

active component of circuit admittance

$$G_{\kappa} = \frac{C_{\kappa} R_{\kappa}}{L_{\kappa}} = \frac{R_{\kappa}}{\rho_{\kappa}^2}; \qquad (4)$$

circuit characteristic impedance

$$\rho_{\kappa} = \sqrt{\frac{L_{\kappa}}{C_{\kappa}}}; \qquad (5)$$

quality factor of nonloaded circuit

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$$Q = \frac{\rho_K}{R_K}; \qquad (6)$$

active component of the admittance of oscillator load - G_{ng} (the reactive component of load conductance will be ignored, because it can be compensated by tuning the circuit)

coefficient of circuit connection from the output side of the four-terminal network

$$n_1 = \frac{U_2}{U_3}; \qquad (7)$$

transformation ratio from the input side of the four-terminal network

$$n_2 = \frac{U_4}{U_3}. \qquad (8)$$

At small signals the relationship between the voltage and current variable components of the four-terminal network is determined by the following equations in Y parameters:

$$\left. \begin{aligned} \dot I_1 &= Y_{11}\dot U_1 + Y_{12}\dot U_2, \\ \dot I_2 &= Y_{21}\dot U_1 + Y_{22}\dot U_2. \end{aligned} \right\} \qquad (9)$$

Here $Y_{11} \ldots Y_{22}$ - character admittances of the four-terminal network representing the transistor. These admittances appear to be of the complex kind and, for a common-emitter circuit, can be represented by comparatively simple combinations or resistances, capacitances and inductances (Fig.3) (Bibl.3).

As can be seen from Fig.3 the analytic formulas for these admittances can be presented in the following form:

$$Y_{11} = \frac{r_{11} - i\frac{1}{\omega C_{11}}}{r_0 r_{11} - i\frac{1}{\omega C_{11}}(r_{11} + r_0)} \approx \frac{r_0 r_{11} + \frac{1}{\omega^2 C_{11}^2}}{r_{11}\left(r_0^2 + \frac{1}{\omega^2 C_{11}^2}\right)} + i\frac{\frac{1}{\omega C_{11}}}{r_0^2 + \frac{1}{\omega^2 C_{11}^2}} \qquad (10)$$

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at $r_o \ll r_{11}$

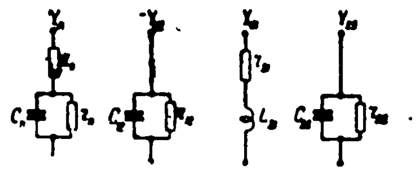

Fig.3

$$Y_{12} = \frac{1}{r_{12}} + i\omega C_{12}, \qquad (11)$$

$$Y_{21} = \frac{1}{r_{21} + i\omega L_{21}} = \frac{r_{21}}{r_{21}^2 + \omega^2 L_{21}^2} - i\frac{\omega L_{21}}{r_{21}^2 + \omega^2 L_{21}^2}, \qquad (12)$$

$$Y_{22} = \frac{1}{r_{22}} + i\omega C_{22}. \qquad (13)$$

As for the common-base circuit, the characteristic admittances of the network can be expressed by characteristic admittances of the common-emitter circuit in the following form:

$$Y'_{11} = Y_{11} + Y_{12} + Y_{21} + Y_{22} \approx Y_{11} + Y_{21}, \qquad (14)$$

$$Y'_{12} = -(Y_{12} + Y_{22}), \qquad (15)$$

$$Y'_{21} = -(Y_{21} + Y_{22}) \approx -Y_{21}, \qquad (16)$$

$$Y'_{22} = Y_{22}. \qquad (17)$$

The formula for network load (admittance of circuit in Fig.2 to the right of the section bb), the related formula can be thus presented

$$Y_n = \frac{1}{n_1^2}\left[Y_k + n_2^2 Y_{in} + Y'_{ol}\right], \qquad (18)$$

where

$$Y'_{ol} = Y_{ol}\, n_3^2, \quad n_3 = \frac{U_3}{U_2}.$$

The formula for input admittance of the network has, as is known (Bibl.2) the following form

$$Y_{in} = \frac{I_1}{U_1} = Y_{11} - \frac{Y_{12} Y_{21}}{Y_{22} + Y_k}. \qquad (19)$$

Substituting formula (19) in eq.(18), we will determine the value of Y_n by the parameters of the circuit, load, and network

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$$Y_n = \frac{-D + \sqrt{D^2 - 4 n_1^2 \varepsilon}}{2 n_1^2}, \tag{20}$$

where

$$D = n_1^2 Y_{22} - Y - n_2^2 Y_{11}, \tag{21}$$

$$Y = Y_\kappa + Y_{ol}, \tag{22}$$

$$\varepsilon = n_2^2 Y_{12} Y_{21} - Y_{22} Y - n_2^2 Y_{11} Y_{22}. \tag{23}$$

Taking into account eqs.(7) and (8), and also considering that $I_2 = - U_2 Y_n$, the second equation in the system (9) can be used to obtain the formula for the full voltage transmission ratio of the entire system in the following form

$$\dot{K} = \frac{\dot{U}_2}{\dot{U}_1} = -\frac{n_2}{n_1} \cdot \frac{Y_{21}}{Y_{22} + Y_n}. \tag{24}$$

Substituting into eq.(24) the values of Y_{21}, Y_{22}, and Y_n for specific circuits, and taking into account eqs.(1) and (2) it is possible to determine the conditions for self-excitation and for the frequency of generated oscillations.

For the common-emitter circuit the formula for the frequency of generated oscillations assumes the following form

$$\omega = \sqrt{\frac{B + \sqrt{B^2 + 4 A r_{21}}}{2A}}, \tag{25}$$

where

$$A = L_\kappa C_{11}^2 r_0 \left[r_{21} r_0 (C_\kappa + n_1^2 C_{22}) + L_{21} r_0 \left(G + \frac{n_1^2}{r_{22}} \right) + n_1^2 L_{21} \right],$$

$$B = L_\kappa \left[r_{21} C_{11} \left(\frac{r_0^2}{L_\kappa} C_{11} - n_1^2 \right) - r_{21} (C_\kappa + n_1^2 C_{22}) - L_{21} \left(G + \frac{n_1^2}{r_{22}} \right) \right].$$

If it be considered that the following inequalities apply for the radio frequencies of (f > megacycle):

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$$r_0 r_{11} \omega^2 C_{11}^2 \gg 1, \tag{26}$$

$$r_1^2 \omega^2 C_{11} \gg 1, \tag{27}$$

$$\frac{L_{21} n_1^2}{r_{22}} \gg n_2^2 C_{11} \tag{28}$$

$$1 \gg \frac{n_1^2 L_{21}}{r_{21} r_0 C_K}, \tag{29}$$

$$1 \gg \frac{n_1^2 L_K}{r_0 C_{11}}, \tag{30}$$

then eq.(25) can be easily reduced to this form

$$\omega = \frac{\omega_0}{\sqrt{1 + \frac{1}{C_K}\left[n_1 C_{22} + \frac{L_{21}}{r_{21}} \left(G + \frac{n_1^2}{r_{22}} \right) \right]}}. \tag{31}$$

where

$$\omega_0 = \frac{1}{\sqrt{L_K C_K}}.$$

From eq.(31) it can be seen that the greater is the collector capacitance C_k, the higher is the stability of the self-oscillation frequency. It is to be noted that eq.(31) is also correct for lower radio frequencies (f ≈ 100 kilocycles).

The condition for self-excitation in a common-emitter circuit will be written in the following form

$$n_2 = \frac{-n_1 \pm \sqrt{n_1^2 - 4MN}}{2M}, \tag{32}$$

where

$$M = \frac{r_{21}(r_0 r_{11} \omega^2 C_{11}^2 - 1)}{r_{11}(r_{0}^2 \omega^2 C_{11}^2 + 1)} - \frac{\omega^2 L_{21} C_{11}}{r_{0}^2 \omega^2 C_{11}^2 + 1}, \tag{33}$$

$$N = r_{21}\left(G + \frac{n_1^2}{r_{22}} \right) - \omega^2 L_{21} (C_K + n_1^2 C_{22}) + \frac{L_{21}}{L_K}. \tag{34}$$

The inequalities (26) and (27) are correct for frequencies of f > 1 megacycle,

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and the formula for M becomes simplified

$$M \approx \frac{r_{21}}{r_0} - \frac{L_{21}}{r_0^2 C_{11}}. \qquad (35)$$

On frequencies of f < 100 kilocycles

$$M \approx \frac{r_{21}}{r_{11}} (r_0 r_{11} \omega^2 C_{11}^2 - 1) - \omega^2 L_{21} C_{11}. \qquad (36)$$

For the common-base circuit the formula for the frequency of generated oscillation is the same as that for the common-emitter circuit [eq.(25)], but the value B is to be construed as referring to the following expression

$$B = \frac{n_1^2 L_{21}}{r_{11}} - L_\kappa \left[r_{21} C_{11} \left(\frac{r_0^2}{L_\kappa} C_{11} - n_2^2 \right) - r_{21} (C_\kappa + n_1^2 C_{22}) - L_{21} \left(G + \frac{n_1^2}{r_{22}} \right) \right].$$

If we take into account inequalities (26) to (30), the formula for the frequency of generated oscillations for the common-base circuit will be obtained exactly the same as the formula for the common-emitter circuit [formula (31)].

The condition for self-excitation in a common-base circuit has the following form

$$n_2 = \frac{n_1 \cdot \sqrt{n_1^2 - 4MN}}{2M}, \qquad (37)$$

where

$$M = 1 + \frac{r_{21}(r_0 r_{11} \omega^2 C_{11}^2 - 1)}{r_{11}(r_0^2 \omega^2 C_{11}^2 + 1)} - \frac{\omega^2 L_{21} C_{11}}{r_0^2 \omega^2 C_{11}^2 + 1},$$

$$N = r_{21} \left(G + \frac{n_1^2}{r_{22}} \right) - \omega^2 L_{21} (C_\kappa + n_1^2 C_{22}) + \frac{L_{21}}{L_\kappa}.$$

On frequencies of f ≥ 1 megacycle

$$M \approx 1 + \frac{r_{21}}{r_0} - \frac{L_{21}}{r_0^2 C_{11}}.$$

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On frequencies of f 100 kilocycles

$$M \approx 1$$

It should be noted that in the case of inductive coupling

$$n_2 = \frac{M}{L_K}.$$

In the case of capacitive coupling

$$n_2 \approx \frac{C_K}{C_{SV}},$$

where C_{sv} stands for the capacitance from which feedback voltage is taken.

At autotransformer coupling

$$n_2 = \frac{L_1'}{L_K},$$

where $L_1' = L_1 + M$ stands for inductance of coupling.

Experimental verification of eqs.(31), (32) and (37) has proved satisfactory.

In computations of the self-oscillation frequency the verified error did not exceed 10%, and when verifying the condition for self-excitation the error did not exceed 20%.

The table below cites Y-parameter data for three Soviet-produced transistors of the P6G type:

U_k = - 5 volts, I_e = 1 milliampere

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Parameters	Components	P6G No 1			P6G No 2			P6G No 3		
		f in kilocycles			f in kilocycles			f in kilocycles		
		100	500	1000	100	500	1000	100	500	1000
	r_o ohms	73.5	73.5	73.5	78.5	78.5	78.5	72	72	72
Y_{11}	C_{11} .000 micro-microfarads	9.8	7.8	7.5	9.4	7.4	7.25	9.55	7.4	7.2
	r_{11} kiloohms	1.1	1	-	1.5	1.3	-	1.2	1.1	-
$-Y_{12}$	C_{12} micro-microfarads	15	-	-	18	-	-	14	-	-
	r_{12} kiloohms	570	-	-	640	-	-	650	-	-
Y_{21}	L_{21} microhenrys	20	18.1	17.5	17.8	17.6	17.4	17.2	16.8	16
	r_{21} ohms	30.4	27.8	1.92	29	22.2	1.97	30	27	3.49
Y_{22}	C_{22} micro-microfarads	69	35	26	89.5	43	38	87	43.5	32
	r_{22} kiloohms	32	12.2	9.1	38.2	10.6	8.4	41.2	10.9	7.65

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CONCERNING THE SYNTHESIS OF AMPLIFYING CIRCUITS

by

S.V.Samsonenko

Description of a new mathematical procedure for analysis of transient processes in amplifiers. Proposal to employ this procedure as the basis for synthesis of multistage systems according to various standards of signal distortion.

A particularly interesting factor in the designing of amplifiers for pulsed devices is the synthesis of their parameters according to some given standards of pulse distortion. Here, as a rule, the various standards of signal distortion determine various technical characteristics of the pulsed device. For example, steepness of pulse front characterizes the accuracy and resolving power of the pulsed range finder, overshoot characterizes the contrast of television image, distortions corresponding to the optimal signal-to-noise ratio yield the real sensitivity of the pulsed receiver, and so forth.

It is obvious that a definite form of pulse-form distortion is important for every type of radio line, with the allowable value of such distortion depending on the concrete conditions of operation of the pulsed device. The thus resulting distortion standards are to be regarded as initial values in computing a given amplifier circuit.

One fundamental difficulty in the synthesis of amplifying circuits is the complexity of the mathematical apparatus which, as a rule, is based on operational calculus. The application of orthogonal polynomials considerably simplifies a solution of this problem.

The present paper employs the method of orthogonal polynomials to furnish rating formulas underlying the synthesis of any amplifying circuit. In particular, this

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paper furnishes the formulas of relationships between the values characterizing amplifier parameters and such standards of signal distortion as front rise time according to the interdecimal interval, amplitude of the first overshoot, and front rise time according to the maximum steepness of transient process. With the aid of this method it is facile to obtain rating formulas for making such synthesis also according to the other standards of signal distortion that are of practical interest.

The application of orthogonal polynomials makes it also possible to simplify considerably the computation of the transient process, especially in multistage amplifiers.

The paper begins by citing the foundations of the aforementioned mathematical apparatus and also some definitions which will be used hereafter.

1. Standardized Parameters of the Transient Mode and the Transmission Factor of a System

Let us assume that a single break shock acts upon the linear system depicted in Fig.1, with a transmission factor F (p). The signal E (t) which is thereupon obtained on the output will be hereafter referred to as the transient function of the system. As is known, transmission factor F (p) represents a transient function converted according to Carson

$$\frac{1}{p}\Phi(p) = \int_0^\infty E(t)\, e^{-pt}\, dt,$$

where p is complex variable and t is elapsing time.

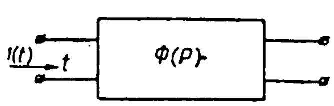

Fig.1

Let us introduce the standardized transient function

$$U(t) = \frac{E(t)}{E(\infty)}.$$

and also the steepness of the standardized transient function

$$f(t) = \frac{dU(t)}{dt}$$

Let us thereupon investigate only the systems for which the E (∞) equals a constant value termed stationary level. It is clear that for such systems U (∞) = 1 and f (∞) = 0. Let us examine only the systems (amplifiers) with zero initial conditions. for which U = 0 at t = 0. In this case, the following expression applies to the standardized parameters

$$\int_0^\infty f(t)\,dt = \int_0^\infty dU(t) = 1. \qquad (1)$$

Let us also introduce the standardized transmission factor of the system, K (p), as a standardized transient response converted according to Carson's law

$$\frac{K(p)}{p} = \int_0^\infty U(t)\,e^{-pt}\,dt. \qquad (2)$$

In accordance with the theorem of differentiation in the range of the real variable, the following formula applies to systems with zero initial conditions

$$K(p) = \int_0^\infty e^{-pt} f(t)\,dt. \qquad (3)$$

The purpose of the analysis is, as is known, to determine the formulas for U (t) or f (t) as functions of the parameters of a system, which is effected with the aid of operational calculus by means of converting eqs.(2) or (3) at a given formula for K (p). If the system is of the multistage type and is determined by an equation of a high order, the computations become greatly involved; the formulas thus obtained for some simple systems are so cumbersome that their computation, and the determination of the system parameters ensuring the necessary form of transient process, is an extremely arduous process. The present paper proposes a simpler way of analyzing transient responses, as based on dissociating them into Hermite polynomials.

2. Dissociation Into Hermite Polynomials

Threshold theorems of the probability theory (Bibl.1), as applied to the analysis of the asymptotic properties of the transient responses of multistage amplifiers, were used by L.A.Meyerovich and G.P.Tartakovskiy (Bibl.2) to demonstrate that at a number of stages equal to $n \to \infty$ the form of the curve of the transient mode on the output of a system strives, at the presence of break shock, to approximate Kramp's function, while the steepness of the transient process tends to approximate the Hermite function. Hence, it is natural to examine the transient responses of a multistage system from the viewpoint of their approximation to Kramp's and Hermite's functions, by representing them accordingly in the form of functional series whose coefficients should be functions of the parameters of the system.

As demonstrated in a previous paper (Bibl.3), such a dissociation, as applied to the steepness of the standardized transient response, can be represented as follows

$$f(t) = \sum_{\nu=0}^{\infty} \frac{B_\nu}{\nu!} \frac{\varphi^{(\nu)}(x)}{\sqrt{2\pi}} \qquad (4)$$

Here $\varphi^{(\nu)}(x) = (-1)^{\nu} e^{-\frac{x^2}{2}} H_\nu(x)$ is the ν^{th} derivative of Ermit's function, $H_\nu(x)$ is Hermite polynomial of the ν^{th} order,

B_ν are dissociation coefficients,

x is the standardized variable $x = \frac{t - A_1}{A_1}$, where t stands for elapsing time and A_1 stands for a constant value which can be subsequently determined (see eq.(9) at $\nu = 1$).

In order that dissociation (4) have sense, the coefficients B_ν should exist. It can be demonstrated that coefficients B_ν exist for all systems whose transmission factor has separate points on the left semi-plane. Without giving proof of this statement at the time let us pass over to determining the formulas for the dissociation coefficients B_ν. For this purpose, let us multiply both parts of eq.(4) by

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H_h (x) and carry out an integration ranging from $-\infty$ to ∞. Proceeding from the requirement of orthogonality of hermitian polynomials weighing $e^{-\frac{x^2}{2}}$ we obtain the following formula on the two-sided infinite interval

$$B_\nu = (-1)^\nu \int_{-\infty}^{\infty} f(t) H_\nu(x)\, dx. \quad (5)$$

Substituting hermitian polynomials into eq.(5) and carrying out integration within the abovementioned limits, we obtain

$$\left.\begin{aligned} B_0 &= \frac{1}{A_1}; \quad B_2 = \frac{M_2}{A_1^2} - 1 \\ B_3 &= \frac{M_3}{A_1^3}; \quad B_4 = \frac{M_4}{A_1^4} - \frac{6M_2}{A_1^2} + 3 \\ B_5 &= \frac{M_5}{A_1^5} - 10\frac{M_3}{A_1^3} \end{aligned}\right\}, \quad (6)$$

where M_1, M_2, . . . M_ν stand for the central moments of the standardized transient response of a system, determined by the formula

$$M_\nu = \int_{-\infty}^{\infty} (t - A_1)^\nu f(t)\, dt. \quad (7)$$

On removing the parentheses and integrating, we obtain

$$\left.\begin{aligned} &M_2 = A_2 - A_1^2, \quad M_3 = A_3 - 3A_1A_2 + 2A_1^3 \\ &M_4 = A_4 - 4A_1A_3 + 6A_1^2A_2 - 3A_1^4 \\ &\ldots\ldots\ldots\ldots\ldots \\ &M_\nu = A_\nu - \nu A_{\nu-1}A_1 + \frac{\nu(\nu-1)}{2!}A_{\nu-2}A_1^2 + \ldots (-1)^\nu A_1^\nu \end{aligned}\right\}. \quad (8)$$

where A_1, A_2, A_3 . . . A_ν stand for simple moments of the standardized transient response

$$A_\nu = \int_{-\infty}^{\infty} t^\nu f(t)\, dt. \quad (9)$$

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Obviously A_ν is the parameter of our system, which characterizes it in a definite manner. Another previous paper (Bibl.4) demonstrates that the moments of the system A_ν can be expressed by moments of separate stages α_ν^i, where

$$\alpha_\nu^i = \int_0^\infty \varphi_i(t)\, t^\nu\, dt, \tag{10}$$

and $\varphi_i(t)$ stands for the steepness of the transient response of the separate i^{th} stage. This steepness is determined through the standardized transmission factor of the i^{th} stage, $K_i(p)$ with the aid of the Laplace transform, by the following formula

$$K_i(p) = \int_0^\infty e^{-pt}\, \varphi_i(t)\, dt. \tag{11}$$

If a system consists of n different stages and its transmission factor is

$$K(p) = \prod_{i=1}^{n} K_i(p),$$

then, by introducing the logarithmic transmission factor (Bibl.2, 4), the connection between A_ν and $\alpha_\nu^{(i)}$ can be presented in this form

$$\left.\begin{aligned}
A_1 &= \sum_{i=1}^{n} \alpha_1^i \\
A_2 &= \sum_{i=1}^{n} (\alpha_2^i - \alpha_1^{i^2}) + \Big(\sum_{i=1}^{n} \alpha_1^i\Big)^2 \\
A_3 &= \sum_{i=1}^{n} (\alpha_3^i - 3\alpha_1^i\alpha_2^i + 2\alpha_1^{i^3}) - 3\sum_{i=1}^{n}\alpha_1^i \Big[\sum_{i=1}^{n} (\alpha_2^i - \alpha_1^{i^2}\Big] + \Big(\sum_{i=1}^{n}\alpha_1^i\Big)^3 \\
A_4 &= \sum_{i=1}^{n} (\alpha_4^i - 3\alpha_2^{i^2} - 4\alpha_1^i\alpha_3^i + 12\alpha_1^i\alpha_2^i - 6\alpha_1^{i^4}) + 3\Big[\sum_{i=1}^{n} (\alpha_2^i - \alpha_1^{i^2}\Big]^2 + \\
&\quad + 4\Big(\sum_{i=1}^{n}\alpha_1^i\Big)\Big[\sum_{i=1}^{n} (\alpha_3^i - 3\alpha_1^i\alpha_2^i + 2\alpha_1^{i^3}\Big] + \\
&\quad + 6\Big[\sum_{i=1}^{n}\alpha_1^i\Big)^2\Big[\sum_{i=1}^{n} (\alpha_2^i - \alpha_1^{i^2})\Big] + \Big(\sum_{i=1}^{n}\alpha_1^i\Big)^4
\end{aligned}\right\} \tag{12}$$

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On the other hand, in the event of identical stages $K(p) = K_1^n(p)$

$$\left.\begin{aligned} A_1 &= n\alpha_1 \\ A_2 &= n(\alpha_2 - \alpha_1^2) + n^2\alpha_1^2 \\ A_3 &= n(\alpha_3 - 3\alpha_1\alpha_2 + 2\alpha_1^3) - 3n^2(\alpha_2 - \alpha_1^2)\alpha_1 + n^3\alpha_1^3 \\ A_4 &= n(\alpha_4 - 3\alpha_2^2 - 4\alpha_1\alpha_3 + 12\alpha_1^2\alpha_2 - 6\alpha_1^4) + 3n^2(\alpha_2 - \alpha_1^2)^2 + \\ &+ 4n^2\alpha_1(\alpha_3 - 3\alpha_1\alpha_2 + 2\alpha_1^3) + 6n^3\alpha_1(\alpha_2 - \alpha_1^2) + n^4\alpha_1^4 \\ &\ldots\ldots\ldots\ldots\ldots\ldots\ldots \end{aligned}\right\} \quad (13)$$

Analogous formulas can be obtained for central moments:

$$M_2 = \sum_{i=1}^{n} \mu_2^{(i)}, \quad M_3 = \sum_{i=1}^{n} \mu_3^{(i)}. \quad (14)$$

If the stage moments are known, the substitution of eqs.(8) and (12) into the formulas for coefficients B_ν, and hence into series (4), will yield a solution to the posed task.

When using this series it is to be considered that, for practical purposes, the petty details of transient processes usually are of no interest; therefore, on being satisfied with an accuracy of the order of 5 to 10%, it is possible to discard members in the dissociations whose absolute value is below 5 to 10% of the modulus of the first member. Inasmuch as the value of functions $\varphi_1(x), \varphi_2(x) \ldots$ is of one and the same order, and $B_o = 1$, we can ignore the members for which

$$B_\nu < (0,1 \div 0,5)\,\nu! \quad (15)$$

Proceeding from this assumption, it can be presupposed that a definite sum of the series, consisting of four members, will yield the general formula for the steepness of the transient process of a system

$$f(t) = \frac{1}{A_1}\left[\varphi(x) + \frac{B_2}{2}\varphi''(x) - \frac{B_3}{3!}\varphi'''(x) + \frac{B_4}{4!}\varphi''''(x)\right]. \quad (16)$$

It is easy to make the transition from steepness f (t) to transient process by

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means of integration of members. In effect, considering that

$$\int_{-\infty}^{x} f(t) A_1 dt = U(x)$$

and also considering that

$$\int_{-\infty}^{x} \varphi(x) dx = \Phi(x) \quad \text{and} \quad \int_{-\infty}^{x} \varphi^{\nu}(x)\, dx = \varphi^{\nu-1}(x),$$

where $\varphi(x)$ is Kramp's function, we obtain the following formula for the transient process

$$U(x) = \Phi(x) + \frac{B_2}{2!} \varphi'(x) - \frac{B_3}{3!} \varphi''(x) + \frac{B_4}{4!} \varphi''' \qquad (17)$$

For the multistage systems consisting of identical stages the formulas for coefficients B_ν can be still more simplified. In this case

$$\left. \begin{array}{llll} B_2 = \frac{a}{n} - 1; & B_3 = \frac{b}{n^2}; & B_4 = \frac{c}{n^3} + \frac{d}{n^2} - \frac{6a}{n} + 3 \\ a = \frac{\mu_2}{\alpha_1^2}; & b_3 = \frac{\mu_3}{\alpha_1^3}; & c = \frac{\mu_4 - 3\mu_2^2}{\alpha_1^4}; & d = 3a^2 \end{array} \right\} \qquad (18)$$

Considering that the values a, b, c, and d, have one and the same order, only members of the order of $\frac{1}{n}$ should be considered in formulas for B_ν, and the members of a higher power, $\frac{1}{n^2}$ and $\frac{1}{n^3}$ can be ignored. Thereupon, the coefficients will be as follows

$$B_0 = 1; \quad B_2 = \frac{a}{n} - 1; \quad B_3 = 0; \quad B_4 = 3 - \frac{6a}{n}. \qquad (19)$$

Analogous simplifications can be obtained for an amplifier with non-identical stages.

The computation of transient processes according to eqs. (16) or (17) is much simpler than if done according to the formulas obtained by the operational calculus method. In effect, the multipliers determining the evolution of the process with

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time – the derivatives of Hermite and Kramp functions – are tabulated, while the coefficients B_ν are constant numbers at given parameters of a system. Moreover there is no increase in the difficulty of such computing of a system consisting of differing stages, while the analysis of such a system by the operational method would be very complicated.

In order to conduct computations according to eqs.(17) and (18) it is necessary to know the formulas for B_ν as a function of system parameters; for this purpose it is necessary to establish a relationship between stage moments and stage parameters.

3. The Calculation of the Moments of Transient Responses of Amplifying Stages

If the set stage transmission factor pertains to K_i (p), then $\alpha_\nu^{(i)}$ can be computed by applying conversion formulas to the formula for K_i (p) and thereupon computing the integrals according to eq.(10). But this procedure leads to very cumbrous calculations at even a small increase in the complexity of a stage circuit, and therefore we are presenting below a simpler method of solving this task and eliminating the need for resorting to conversion formulas. In effect, the general form of the transmission factor for most amplifying circuits should be represented by a **fractional-rational function of this form**

$$K_i(p) = \frac{1 + a_1 p + a_2 p^2 + \ldots a_m p^m}{1 + b_1 p + b_2 p^2 + \ldots + b_n p^n}, \qquad (20)$$

where $n > m$.

On dissociating the formula for K_i (p) into a series according to positive powers of p, and on substituting into eq.(11), we obtain

$$1 + \sum_{\nu=1}^{\infty} C_\nu p^\nu = \int_0^\infty \varphi(t)\, e^{-pt}\, dt, \qquad (21)$$

where C_1, C_2 . . . C_ν are determined by means of identical conversions. It is to be noted that the left-hand part of this equality converges within a circle with a

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radius extending to the first separate point, and with center at point p = 0; as for the right-hand part, it converges virtually for all p's located in the right-hand semi-plane: thus both parts of the equality have large areas of convergence on the plane of the complex variable.

On representing e^{-pt} in the form of an exponential series

$$e^{-pt} = 1 - pt + \frac{pt^2}{2!} - \frac{(pt)^3}{3!} + \dots$$

and substituting into the right-hand part of eq.(21), we obtain

$$1 + \sum_{\nu=1}^{\infty} C_\nu p^\nu = \sum_{\nu=1}^{\infty} a_\nu (-1)^\nu \frac{p^\nu}{\nu!},$$

where $\alpha_\nu = \int_0^\infty t^\nu f(t)\,dt$ is the moment of the ν^{th} order of the transient function of stage. On making equal the coefficients in this formula at equal powers of p, we obtain the simple formula

$$\alpha_\nu = (-1)^\nu \nu! C_\nu. \qquad (22)$$

Table 1 cites values of α_ν computed for some forms of the transmission factor according to eq.(22). Moreover, considering that the connection between central and simple moments for individual stages is expressed by a formula analogous to eq.(8), Table 1 can be used as the basis for computing the central moments of stages. The results of the calculations for some forms of the transmission factor are given in a general form in Table 2.

An example is provided in Table 3 which cites the values of moments α_ν and μ_ν for specific stage circuits computed according to the formulas in Tables 2 and 1.

Table 1

Table of Moments d_ν

$K(p)$ \ α_ν	α_1	α_2	α_3	α_4
$\frac{1}{1+b_1 p}$	b_1	$2b_1^2$	$6b_1^3$	$24b_1^4$
$\frac{1+a_1 p}{1+b_1 p+b_2 p^2}$	b_1-a_1	$2(b_1^2-b_1a_1-b_2)$	$-6(2b_1b_2+b_1^2a_1-b_1^3-a_1b_2)$	$24(b_2^2+2b_1b_2a_1+b_1^4-b_1^3a_1+3b_1^2b_2)$
$\frac{1+a_1p+a_2p^2}{1+b_1p+b_2p^2+b_3p^3}$	b_1-a_1	$2[a_2-b_1(a_1-b_1)-b_2]$	$-6[b_1^2(a_1-b_1)+2b_1b_2-b_1a_1+ +b_2a_1-b_3]$	$24[b_1^2a_1-b_1^3(a_1-b_1)-b_2a_2+2b_2b_1a_1- -3b_1^2b_2+b_2^2-b_3a_1+2b_1b_3]$

Table 2

Table of Central Moments μ_ν

$K(p)$ \ μ_ν	μ_2	μ_3	μ_4
$\frac{1}{1+b_1p}$	b_1^2	$2b_1^3$	$9b_1^4$
$\frac{1+a_1p}{1+b_1p+b_2p^2}$	$b_1^2-a_1^2-2b_2$	$2b_1^3-2a_1^3-6b_1b_2$	$3(8b_2^2+3b_1^4-12b_1^2b_2-2b_1^2a_1^2+4a_1^2b_2-8a_1b_1^3-a_1^4)$

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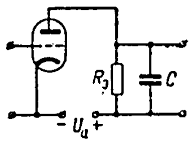
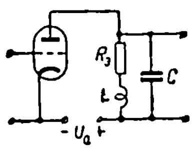
Table 3

Type of Stage Circuit	Standardized Transmission Factor	Stage Moments	Central Stage Moments
	$K(p) = \frac{1}{1+p}$ $p = i\omega R_a C$	$a_1 = 1;\ a_2 = 2$ $a_3 = 6,\ a_4 = 24$	$\mu_1 = 0;\ \mu_2 = 1$ $\mu_3 = 2;\ \mu_4 = 9$
	$K(p) = \frac{1+mp}{1+p+mp^2}$ $p = i\omega R_a C$ $m = \frac{L}{CR_a^2}$	$a_1 = 1 - m;$ $a_2 = 2(1-2m)$ $a_3 = 6(1-3m+m^2)$ $a_4 = 24(1-4m+3m^2)$	$\mu_1 = 0;\ \mu_2 = 1-2m-m^2$ $\mu_3 = 2(1-3m-m^3)$ $\mu_4 = 3(6-24m+18m^2+4m^3-m^4)$

4. <u>Examples of the Analysis of Transient Processes By Means of Orthogonal Polynomials</u>

a. Resistor-Coupled Amplifier Without Correction. In accordance with Table 3 and eqs.(18), we have

$$a = 1; \quad b = 2; \quad c = 6; \quad d = 3;$$

dissociation coefficients

$$B_2 = \frac{1}{n} - 1; \quad B_3 = \frac{2}{n^2}; \quad B_4 = \frac{6}{n^3} + \frac{3}{n^2} - \frac{6}{n} + 3$$

and the transient process will be, in accordance with eq.(17)

$$U(x) = \Phi(x) + \frac{1-n}{2n}\varphi'(x) - \frac{1}{3n^2}\varphi''(x) + \frac{1}{24}\left(\frac{6}{n^3} + \frac{3}{n^2} - \frac{6}{n} + 3\right)\varphi'''(x).$$

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In the particular case of a six-stage amplifier ($n = 6$), we have

$$U(x) = \Phi(x) - 0{,}416\,\varphi'(x) - 0{,}00925\,\varphi''(x) + 0{,}0875\,\varphi'''(x). \qquad (23)$$

The graph of this function is depicted in Fig.2

b. Resistor-Coupled Amplifier With Parallel Correction. In accordance with Table 3 and eqs.(18), we have

$$a = \frac{1 - m^2 - 2m}{(1-m)^2}; \qquad b = 2\,\frac{1 - 3m - m^2}{(1-m)^2};$$

$$c = 6\,\frac{1 - 4m + 2m^2 - m^4}{(1-m)^4}; \qquad d = 3\,\frac{(1 - m^2 - 2m)^2}{(1-m)^4}.$$

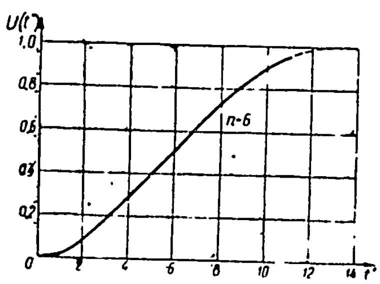

Fig.2

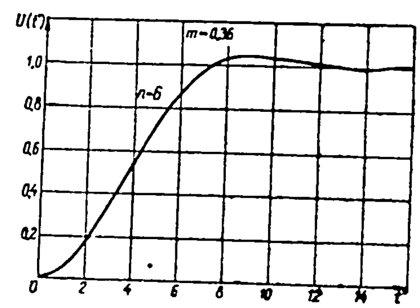

Fig.3

In the particular instance of a six-stage amplifier and $m = 0.36$, we have

$$a = 0{,}33; \quad b = -0{,}98; \quad c = -7{,}05; \quad d = 0{,}402;$$

$$B_2 = -0{,}94; \quad B_3 = -0{,}0272; \quad B_4 = 2{,}613.$$

In this case the transient process formula will be

$$U(x) = \Phi(x) - 0{,}47\,\varphi'(x) + 0{,}00456\,\varphi''(x) + 0{,}1\,\varphi'''(x). \qquad (24)$$

The graph of this function is depicted in Fig.3.

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5. Synthesis of Amplifying Circuits

It can be concluded from an examination of eqs.(23) and (24) that multistage amplifiers are not greatly affected by the third and fourth members of the dissociations; therefore, if a computing accuracy of the order of 10% be recognized as sufficient, the corresponding formula for the transient response can be represented in the form of a binomial

$$U(x) = \Phi(x) + \frac{B_3}{2}\varphi'(x), \tag{25}$$

and the steepness, in this form

$$f(t) = \frac{1}{A_1}\left\{\varphi(x) + \frac{B_3}{2}\varphi''(x)\right\}. \tag{26}$$

Thereupon the problem of the synthesis is reduced to investigating the applications of eqs.(25) and (26) to various types of signal distortion. Let us conduct syntheses of amplifiers.

a. Synthesis According to a Given Steepness of Transient Response (Pulse Front). Considering that the steepness of the front is variable, it can be set in various ways. In a number of practical cases it is important to know the maximum steepness of pulse front, which is the most characteristic point on the pulse front and as such is most easily recognizable by station operators. Therefore, in the given case we will adopt the maximum steepness of transient process at the output of a system as the initial parameter for the synthesis.

The coordinates of the point of maximum steepness are determined from the following equality

$$\frac{df_1(t)}{dx} = 0. \tag{27}$$

On substituting the Hermite functions into eq.(27) and taking the derivatives, we obtain, after reduction

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$$x\left[\frac{B_1}{2}(3-x^2)-1\right]=0.$$

The root x = 0 is the sought-for root, because it corresponds to the point of maximum steepness in the middle of the transient process. In effect, upon discarding in eq.(25) the member with the coefficient B_3 determining the asymmetry of the transient process, we, of course, obtain the symmetrical transient process in which the time coordinate of the maximum steepness coincides with the time coordinate of the center of the rise front.

On substituting the value x = 0 into eq.(26) we obtain the value of maximum steepness as a function of circuit parameters

$$f(A)=\frac{0,4}{A_1}\left[1-\frac{B_2}{2}\right]. \quad (28)$$

At a set steepness, the derivation of this equality in relation to circuit parameters presents no difficulties.

Inasmuch as all the above-expounded reasoning applies to standardized transient responses, a value opposite to steepress represents the front rise time τ_f

$$\tau_f=\frac{A_1}{0,4\left(1-\frac{B_2}{2}\right)}.$$

Usually, all problems of the transient process are solved in infinite time t = βt', where t' is time in seconds and β is the dimensional coefficient, which depends on circuit parameters. Therefore the front time determinable by eq.(2) is also dimensionless, and the front time in seconds will be

$$t'=\frac{\tau_f}{\beta}=\frac{A_1}{0,4\left(1-\frac{B_2}{2}\right)\beta}.$$

This formula can serve for computing the duration of the front of the transient

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process on the output of any multistage system upon substituting into it the values of A_1, B_2 and β corresponding to the given circuit.

At a practical synthesis it is also necessary to consider the resultant amplification factor of the system, and then the problem of synthesis has to be differently formulated. In this case, the set values are: a) time (or steepness) of transient process; b) parameters of amplifier tube S and C_k; and c) required amplification factor. The solution of this problem by means of eq.(26) is easily applicable to any circuit.

b. Synthesis of Amplifier According to Overshoot of Transient Response. In this case, the initial value for the calculations is represented by the amplitude of the first overshoot of transient response Δ. The amplitude of overshoot for the standardized transient response will be, in accordance with eq.(25)

$$\Delta = \Phi(x_1) + \frac{B_2}{2}\varphi'(x_1) - 1, \qquad (29)$$

where x_1 is the time coordinate of overshoot, determined from the following requirement

$$f(t) = 0, \text{ i.e., } \varphi(x) + \frac{B_2}{2}\varphi''(x) = 0.$$

On substituting here the formulas for Ermit functions and carrying out the reductions, we obtain

$$x^2 + \frac{2}{B_2} - 1 = 0.$$

We are interested in the positive root of this equation, which will be

$$x_1 = \sqrt{1 - \frac{2}{B_2}}.$$

At substitution of the value of this root into eq.(29) it can be seen that the amplitude of overshoot is a function of the value of B_2. A graph of this relationship for larger values of B_2 (B_2 . 1) is depicted in Fig.4, and for smaller values

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of B_2 (B_2 1), in Fig.5.

It is clear from these graphs that the relationship $\Delta = \psi(B_2)$ has a linear character at larger values of B_2, and is approximated by the formula

$$\Delta\% = 9{,}5\,(B_2 - 0{,}63).$$

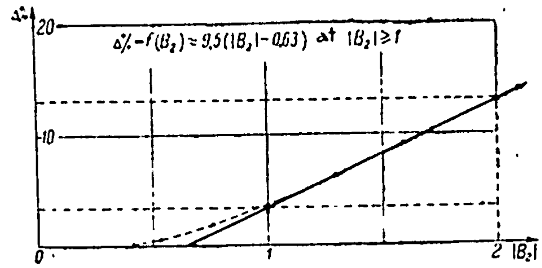

Fig.4

At small values of B_2 this relationship has a non-linear character, but for practical purposes it can be approximated with sufficient accuracy by two straight lines

$$\Delta = 6{,}4\,(B_2 - 0{,}41) \quad \text{at} \quad 1 > B_2 > 0{,}5,$$
$$\Delta = 2{,}1\,(B_2 - 0{,}27) \quad \text{,,} \quad 0{,}5 > B_2 > 0{,}3.$$

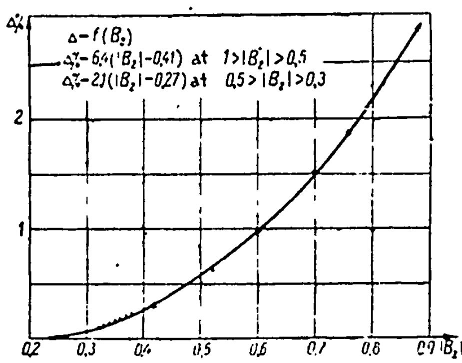

Fig.5

As for the coefficient B_2 for an amplifier with identical stages, it is determined by the following equation

$$B_2 = \frac{\mu_2}{n\,a_1^2} - 1. \qquad (30)$$

This formula can serve, together with the graphs in Fig.4 and 5, both for computing amplifier parameters according to a given value of overshoot and for determining the amplitude of overshoot according to known parameters of multistage circuit. On substituting the value of moments for this or that amplifier circuit into eq.(30) we obtain rating formulas for a specific multistage circuit schematic. The conduct of the related computations presents no difficulties.

c. Amplifier Synthesis According to Rise Time Corresponding to the Interdecimal Ordinate Drop (0.1 - 0.9) of the Transient Response. The rise time corresponding to the interdecimal ordinate drop is determined by the following formula

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$$\tau_{\varphi 0,1-0,9} = t_{0,9} - t_{0,1} = A_1 (x_{0,9} - x_{0,1}). \qquad (31)$$

where $x_{0.9}$ and $x_{0.1}$ are determined from eq.(25) by deriving the following equations

$$U(x_{0,9}) = \Phi(x_{0,9}) + \frac{B_2}{2} \varphi'(x_{0,9}),$$
$$U(x_{0,1}) = \Phi(x_{0,1}) + \frac{B_2}{2} \varphi'(x_{0,1}).$$

On deriving these equations graphically we obtain the relationships $x_{0.1}$ and $x_{0.9}$ as functions of B_2 (Fig.6), on the basis of which the sought-for relationship is determined.

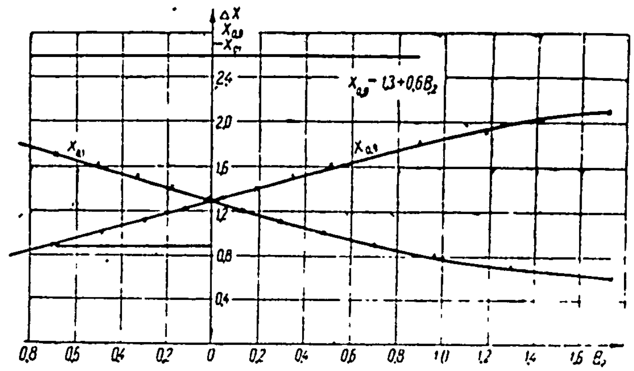

Fig.6

As can be seen from the above graph, relationships $x_{0.1} = \varphi_1 (B_2)$ and $x_{0.9} = \varphi_2 (B_2)$ have identical inclination, but only $x_{0.1}$ varies in the field of negative values of x; in this case, the value of $\Theta (B_2)$ is determined by the sum of values of $x_{0.9}$ and $x_{0.1}$, and not by the difference in these values. On carrying out the summation (Fig.6), we find that $\Theta (B_2)$ does not depend on B_2 and equals a constant value of $\Theta = 2.6$. Then the dimensionless duration of the interdecimal drop will be

$$\tau_{0,1-0,9} = A_1 \cdot 2,6.$$

This formula also furnishes a solution to the problem of the interdecimal-interval synthesis.

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RIPPLE FILTERS OF LOW-POWER RECTIFIERS

by

L.L.Dekabrun

Some specific proposals about computing the elements of the pi filters widely applied in low-power feed sources.

The methodology of the analysis of rectifier processes is fairly well developed by now. On the basis of a series of rational postulates all the information necessary in rectifier designing has been reduced to suitable graphs and formulas (Bibl.1), which are cited in student handbooks and manuals. But there is much less information available about the designing of ripple filters; moreover, the available information leaves us with a feeling of dissatisfaction. Thus for instance, some works (Bibl.2) cite empirical choke computing formulas which have to be considerably deviated from. Other works (Bibl.3) make attempts at analyzing filter processes (more exactly, filter-choke processes). It would be worthwhile to review briefly these attempts in order to get an idea of the situation in this field as a whole.

In Bibl.3 it is assumed that the maximum dynamic inductance for a given choke is achieved when the choke core displays the maximum dynamic permeability $(\mu_d)_{max}$, i.e., when the magnetic state of the core corresponds to a bend in the magnetization curve. The value of the air gap in the magnetic circuit is selected such as to attain precisely this state of the core at a set direct current component in the winding. No importance need be attached to the error committed by the author of Bibl.3 in the determination of induction at the bending point of the magnetization curve at divers air gaps in the magnetic circuit (in Bibl.3 every value of air gap is corresponded by a specific value of induction at the bending point of the magnetization curve, whereas actually this bending of the curve expresses merely the properties of the core and is always present at one and the same induction in steel, regardless of the value of the air gap). It is incorrect to relate choke computations to a definite coincidence between initial magnetizing of the core and the bend in the magnetization curve, because the dynamic inductance of the choke,

$$L_d = \frac{0{,}4\pi S_c w^2}{\frac{l_c}{\mu_d} + l_v}, \qquad (1)$$

determining the smoothing properties of the filter, is chiefly affected by the number of choke turns w, and not by permeability of steel μ_d, because the cited length of core remains lesser than or comparable in value with the length of the normal air gap l_v up to an induction of 10 to 12 kc. Moreover, none of the existing handbooks cites a complete curve of magnetization of transformer steel which would make it possible to establish reliably the induction at the bending point. The tables cited in the existing handbooks can merely be used to determine induction at 25 ampere-turns per centimeter, i.e., far behind the bending point which ranges at 3 to 4 kc. (Incidentally, from these figures it can be seen how profitable is it to use at the bending point a steel with a saturation induction of 15 to 18 kc.).

Experience shows that the basic characteristics of the feed source, that is to say, its internal resistance and value of alternating voltage background, depend very considerably on the proper choice of ripple filter elements and on the mode of operation of these elements. Therefore, it is relevant to devote serious attention to these matters.

The dynamic inductance of a filter choke at an initial current of I_o in its winding is thus expressed

$$L_d = \frac{E_b}{\left(\frac{dI}{dt}\right)_{I_0}}, \qquad (2)$$

where E_b is the change in choke-winding voltage caused by a change in initial current I_o by a value of dI during a time dt.

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A direct cause for the appearance of voltage is the variation in magnetic flux Φ_o in the magnetic circuit of the choke

$$E_d = w\left(\frac{d\Phi}{dt}\right)_{\Phi_0} \cdot 10^{-8}. \qquad (3)$$

It is clear that Φ_o stands for magnetic flux conditional on winding current I_o, and $d\Phi$ stands for the change in the flux caused by a change in magnetizing current by the value of dI.

Considering that

$$\Phi = B_c S_c, \qquad (4)$$

where B_c is induction in the steel of the magnetic circuit, and S_c is cross-sectional area of the magnetic circuit.

Equation (3) can be transformed

$$E_d = wS_c\left(\frac{dB}{dt}\right)_{B_0} 10^{-8} = wS_c \frac{dB}{dI}\left(\frac{dI}{dt}\right)_{I_0} \cdot 10^{-8} = = w^2 S_c\left[\frac{dB}{d(Iw)}\right]_{B_0}\left(\frac{dI}{dt}\right)_{I_0} \cdot 10^{-8}. \qquad (5)$$

Hence the dynamic inductance of choke, determining the smoothing properties of a filter, will be

$$L_d = w^2 S_c\left[\frac{dB}{d(Iw)}\right]_{B_0} \cdot 10^{-8}. \qquad (6)$$

It is convenient hereafter to deal with inductance per unit of cross-sectional area of magnetic circuit

$$L_{do} = \frac{L_d}{S_c} = w^2\left[\frac{dB}{d(Iw)}\right]_{B_0} \cdot 10^{-8}. \qquad (7)$$

To determine L_{do} the following formula is necessary

$$B = f(Iw) \qquad (8)$$

at divers values of the air gap l_v in the magnetic circuit. This formula can be obtained only in the case when the length of magnetic line in steel l_c is known, i.e., when subsequent analysis cannot remain general any more and has to be specific and to pertain to some definite transformer steel forging selected for a given choke. However, before treating of specific chokes with numerical formulations of corresponding conclusions, let us review the problem from its qualitative aspect, which will assist us in establishing some general laws.

Figure 1 depicts the method of plotting a magnetization curve for a magnetic circuit without indicating the scale of the axes of the coordinates. The air gap in this circuit has a definite value. In order to establish some induction B_1 in this circuit, it is necessary, on the one hand, to have the following ampere-turns

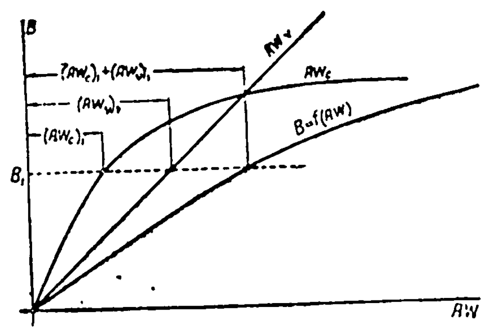

Fig.1

$$(aw)_{c1}=(aw)_{c01}l_c, \qquad (9)$$

creating a magnetic flux in the core and, on the other hand, to have these ampere-turns

$$(aw)_{v1}=\frac{B_1}{0,4\pi}l_v, \qquad (10)$$

for creating a magnetic flux in the air gap.

The curve

$$B=f(aw)$$

in Fig.1 represents the magnetization characteristic of a given magnetic circuit with a given air gap. It is obvious that such curves can be plotted for any value of the air gap l_v so long as this gap is not such as to cause induction in steel and induction in air to differ substantially from each other. The equality of inductions is fairly reliably guaranteed by the following formula

$$\frac{S_r\,[cm^2]}{l_b\,[cm]} \geqslant 100\,[cm]. \qquad (11)$$

This inequality remains true in all practical cases, and therefore the magnetization curve as plotted in Fig.1 does not have to be subsequently corrected. This plotting, and the subsequent graphical differentiation, yield the following formula

$$\frac{dB}{d\,(Iw)} = f\,(Iw)$$

at divers values of the air gap, which is illustrated in Fig.2 where it is obvious that

$$l_{v2} > l_{v1}.$$

These curves illustrate very demonstrably the effect of the air gap: at sufficiently large magnetizing ampere-turns the air gap increases choke inductance. However, the formula for choke inductance includes not only the determined value $\frac{dB}{d(Iw)}$ but also the number of choke winding turns w.

Every value of ampere-turns plotted on the axis of abscissas in Fig.2 is corresponded by a specific value of the number of turns which depends on the value of initial winding current I_o. In Fig.2 there are plotted curves 1, 2, and 3, expressing the relationship $w = f\,(Iw)$ at three different values of I_o. In this connection, it is obvious that

$$(I_0)_1 < (I_0)_2 < (I_0)_3.$$

Thus in Fig.2 there are concentrated all the data necessary for explaining by eq.(7) the specific dynamic inductance of choke, L_{do}. In this connection, we can plot a chart for any value of the current I_o, as illustrated in Fig.3 which demonstrates very graphically that at a given current I_o it is advisable to take a great number of turns in a choke introducing a commensurate air gap and magnetic circuit. It appears that the increase in L_{do} will be steady; however, at sufficiently large air gaps the leakage fluxes will begin to exert a considerable influence and the above-cited chart will cease to be valid.

It is evident that the number of choke turns cannot increase indefinitely, because, owing to the limited area of the port of the choke core S_o the cross-sectional

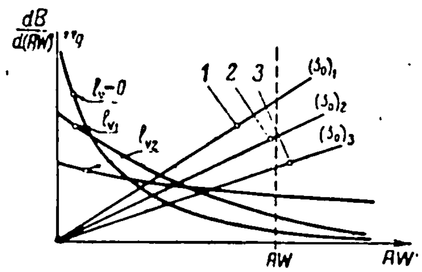

Fig.2

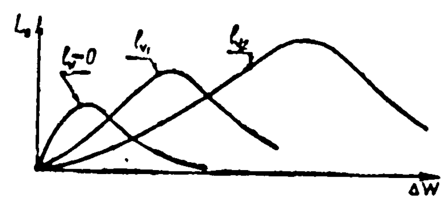

Fig.3

area of the wire will sooner or later prove inadequate for ensuring a safe thermal mode of operation of the choke. Accordingly, in choke designing it is necessary to set a limit on the number of turns which will enter the choke port without endangering the safety of the thermal aspect of operation of the cross-sectional area of the winding wire for a given current. For chokes with enameled-wire windings the winding current density should not exceed 2 to 2.5 a/mm^2. A proper value of the air gap l_v is chosen for the thus obtained number of turns, and then the transformer steel forging selected for the choke is utilized to the maximum degree.

The foregoing is illustrated by the calculation of a specific choke with an open Sh-40 core having a port area of $S_o = 20 \times 60 = 1200\ mm^2$. At the customary "pile-up" reeling the cross-sectional area of the copper in the winding takes up 25 to 50% of port area. At the allowable current density of 2 a/mm^2 the number of initial magnetizing ampere-turns aw_o varies from 600 to 1200. Upon plotting the magnetization curves for the given core for divers values of the air gap l_v and upon determining $\frac{dB}{daw}$ at the abovementioned value of magnetizing ampere-turns, we will obtain the following formula

$$\frac{dB}{d\,aw} = f(l_v),$$

illustrated in Fig.4.

The curves in Fig.4 indicate that the optimum air gap for the Sh-40 core meas-

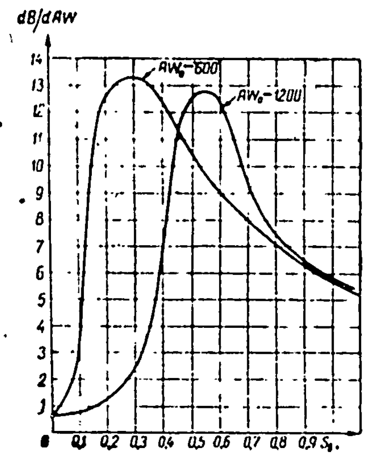

Fig.4

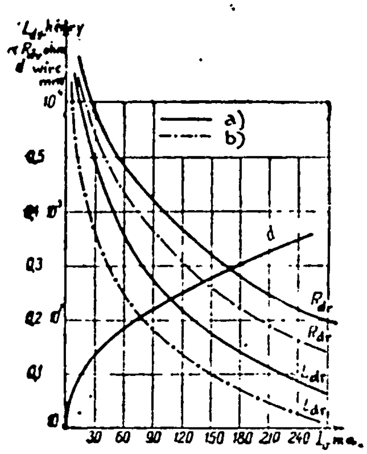

Fig.5

a) Sh40 × 40 steel; b) Sh32 × 32 steel

ures 0.5 mm, when the core port is properly utilized for the winding (here we have in mind the air gap in every leg of the magnetic circuit; thus the full air gap measures 1 mm). At such an air gap the dynamic inductance of the choke is approximately 20 times higher than in the case of a completely closed magnetic circuit.

The specific diameter of wire d and, consequently, the specific number of winding turns, depends on the value of the current I_o which should be passed by the choke. Fig.5 depicts the full response curve of a filter choke with a Sh40 × 40 core at a circuit-leg air gap of 0.5 mm and a port-filling coefficient of 0.5. For comparison, Fig.5 also depicts the curves for a choke with a Sh32 × 32 core at the same port-filling coefficient.

These curves indicate that the choke increases considerably the internal resistance of the feed source, which is undesirable no matter how one looks at it. It is therefore natural to pose the question of how to cause the feed source to have the least internal resistance at a given value of output voltage pulsation.

The available comparative data on all forgings of the transformer steel applied at present in low-power feed sources lead us to select the parameters of the Π-filter (Fig.6) in the following order of sequence:

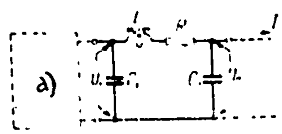

Fig.6

a) Rectifier; b) Toward stabilizer or load

1. Capacitor C_1 and the resistance of transformer and kenotron R_1 are selected by calculating the maximum peaks of charge current allowable for the rectifier's kenotron; all the related necessary information is provided in the aforecited book by E.P.Terent'yev. At such values of C_1 and R_1 we can ensure the minimum amplitude of pulsation of the filter input voltage U_1.

2. Choke inductance L_d should correspond to this formula

$$m\omega L_d \geqslant (10\div 12)\frac{1}{m\omega C_1}, \qquad (12)$$

where m is the number of phases of rectified voltage; ω is the frequency of this voltage. In a case to the contrary the kenotron of the rectifier would be overloaded by current pulses. The type of core and the choke-winding data are selected according to the value of choke inductance found from eq.(12), the known value of the current I_o and the comparative characteristics of various forgings analogous to the curves depicted in Fig.5.

3. The necessary value of the smoothing factor for the fundamental harmonic of rectified voltage

$$\gamma_1 = \frac{U_{1\sim}}{U_{2\sim}} = \sqrt{(1 - m^2\omega^2 L_d C_2)^2 + m^2\omega^2 R^2 C_2^2} \approx m^2\omega^2 L_d C_2 \qquad (13)$$

is used to calculate the value of capacitance C_2 of the filter. In eq.(13) $U_{1\sim}$

stands for the alternating component of the filter input voltage, which is computed by the methods described in Bibl.1; and $U_{2\sim}$ stands for the alternating component of the filter output voltage, whose value is determined by the technical characteristics of the consumer.

Such a ripple filter has the minimum dimensions and increases most minimally the internal resistance of the feed source.

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10

SIMULTANEOUS OSCILLATIONS OF TWO FREQUENCIES IN A SELF-OSCILLATOR WITH SELF-BIAS

by

G.M.Utkin

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A previous paper by the writer (Bibl.2) furnished a general theory of two-circuit self-oscillators with multiple circuit frequencies. The present paper discusses an analogous self-oscillator on taking into account self-bias from grid current. At a low inertiality of the self-bias cell in a two-circuit self-oscillator (Fig.1) it is possible to have stable oscillations of two frequencies, be they multiple or asynchronous (Bibl.3). To demonstrate this, let us investigate the stability of the modes of synchronization and of beats at a low-inertia self-bias from grid current.

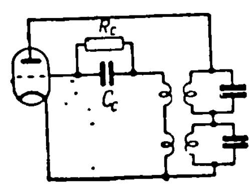

Fig.1

The equations for the self-oscillator shown in Fig.1, with self-bias for the frequency multiples of $n = 3$ and with a linear-broken approximation of the plate current characteristic, have been formulated in Bibl.1. During the derivation of these equations, the compound form of plate and grid currents, having the aspect of a series of higher-frequency pulses periodically repeated with a lower oscillation frequency, has been approximated by a cosinusoid inscribed between the plate-current pulse envelope and the axis of abscissas, as illustrated by broken line in Fig.2 which depicts the form of plate current and its approximation when two multiple-frequency voltages are acting on the grid. As a result, the necessary plate current harmonics prove to be functions of the angle of cutoff Θ of the plate current envelope, and the equations for the self-oscillator assume this aspect:

$$
\begin{aligned}
&T_1\dot{U}_1 = \left(\frac{1}{2} S R_1 \gamma_1 - 1\right) U_1,\\
&T_2\dot{U}_1 = \frac{1}{2} S R_2 \gamma_0 U_1 - U_2,\\
&T\dot{E}_c| = \frac{1}{2} S_c R_c \gamma_0 (\Theta_c) U_1 - |E_c|,\\
&T_2\dot{\varphi} = \Delta\omega T_2 - S R_2\left[\frac{U_1}{2U_2}\gamma_n + \frac{1}{4}\frac{\delta_1}{\delta_2}\cdot\frac{R_1}{R_2}(\gamma_{n-1} - \gamma_{n+1})\right]\sin\varphi,\\
&T_1\dot{\psi}_2 = \frac{1}{4} S R_1 \gamma_{n-1} - \gamma_{n+1})\sin\varphi,
\end{aligned}
\qquad (1)
$$

where U_1, U_2 and φ are the amplitudes and general phase difference in the grid voltages of the fundamental frequencies; E_c is the bias voltage; $\Delta\omega = \omega_2 - n\omega_1$ is the general circuit detuning;

ψ_1 is the correction for oscillation frequency in the first circuit in relation to its natural frequency; $\gamma_1(\Theta)$ are the coefficients of dissociation of the cosinusoidal pulse into the cutoff angle Θ; R_1, R_2 are the control resistances of the circuits; S, S_c are the transconductances of the idealized characteristics of plate and grid circuits $T_1 = \frac{2}{\omega_1\delta_1}$; $T_2 = \frac{2}{\omega_2\delta_2}$; $T_c = R_c C_c$ are the circuit time constants and self-bias cells.

The equations for the synchronization mode are obtained from (1) at $\dot{U}_1 = \dot{U}_2 = \dot{E}_c = \dot{\varphi} = 0$, and have this aspect

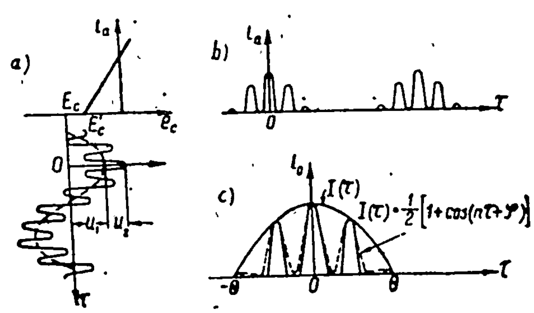

Fig.2

$$
\left.
\begin{aligned}
&S R_1 \gamma_1 = 2,\\
&S R_2 \gamma_0 U_1 = 2U_2,\\
&|E_c| = \frac{1}{2} S_c R_c \gamma_0 (\Theta_c) U_1,\\
&\Delta\omega = (\Delta\omega' + \Delta\omega'')\sin\varphi,\\
&\dot{\psi}_1 = \frac{\Delta\omega''}{n}\sin\varphi
\end{aligned}
\right\} \qquad (2)
$$

The last two equations use the first two equations and introduce the designations:

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$$\left.\begin{aligned}\Delta\omega' &= \frac{\omega_2 d_2}{2}\frac{\gamma_n}{\gamma_0};\\ \Delta\omega'' &= \frac{\omega_2 d_1}{2}\frac{\gamma_{n-1}-\gamma_{n+1}}{2\gamma_1}\end{aligned}\right\} \quad (3)$$

Let us investigate the stability of the modes of simultaneous oscillations of two frequencies on the assumption that the angle of cutoff of grid current is small, i.e., that R_c is sufficiently large. In this case, the bias voltage equals the sum of the voltage amplitudes of the fundamental frequencies and, at a low inertiality of the self-bias cell ($T_c \approx 0$), it continually varies with them. In this connection, it is found that the angle of cutoff of the plate current envelope is not affected by the voltage amplitude of the higher frequency (actually, there is some relationship between them, but a weak one). The justice of the above statements can be easily verified by substituting $E_c = -(U_1 + U_2)$ into the formula for the cosine of the angle of cutoff

$$\cos\Theta = -\frac{E_c - E_c' + U_2}{U_1},$$

whereupon we obtain

$$\cos\Theta = 1 - \frac{|E_c'|}{U_1}.$$

Considering that the first two equations in the initial set (1) are not affected by phase φ*, the study of the stability of the system at variations in voltage amplitudes is reduced to a study of these first two equations in set (1). On composing the corresponding first-approximation formulas it is not difficult to ascertain that the conditions for system stability in face of variations in amplitudes correspond to the satisfaction of this inequality

$$\frac{\partial\gamma_1}{\partial U_1} < 0. \quad (4)$$

* This result is the more probable the higher is the multiplicity of frequencies.

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The above inequality is always satisfied at the assumptions adopted: an increase in U_1 causes a decrease in θ and hence also in the coefficient γ_1.

When a system is stable in relation to the variations in amplitudes, it is possible to have simultaneous oscillations of either the multiple or nonmultiple frequencies corresponding to the synchronization and beat modes. The detuning on the threshold of the synchronization mode is determinable from the requirement for system stability toward variations in phase φ. The requirement for the "phase" stability of a system is easily obtained from the fourth equation in set (1) and has this form

$$(\Delta\omega' + \Delta\omega'')\cos\varphi > 0.$$

As depending on the plus or minus sign of the expression in brackets, the range of stable values of phase φ is $-\frac{\pi}{2} < \varphi < \frac{\pi}{2}$ or $\frac{\pi}{2} < \varphi < \frac{3}{2}\pi$. The extreme values correspond to the thresholds of the synchronization mode. The formula for the detuning on the threshold of the synchronization mode is obtained on substituting $\varphi = \pm\frac{\pi}{2}$ into the fourth equation in set (1), whence it follows that

$$\Delta\omega_{thresh} = \Delta\omega' + \Delta\omega''.$$

The present article will not expatiate upon the peculiarities of the synchronization and beat modes, because these have already been discussed in detail in Bibl.1 and 3.

A conducted experiment has corroborated the conclusion about the possibility of the existence of simultaneous oscillations of two frequencies outside as well as inside the synchronization zone in a self-oscillator with self-bias. These simultaneous oscillations of two frequencies continue uninterruptedly at a smooth change in circuit detuning from one synchronization zone to another.

Article received by Editors on 1 October 1955

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A.S.POPOV SOCIETY NEWS

CONFERENCE ON THE PROBLEMS OF FORWARD SCATTER OF ULTRA SHORT WAVES

In January of this year a conference on the problems of forward scatter of ultra short waves has been jointly convened by the A.S.Popov Scientific-Technical Society for Radio and Communication Engineering, the All-Union Scientific Council for Radio Physics and Radio Engineering of the USSR Academy of Sciences, and the Institute for Radio Engineering and Electronics, USSR Academy of Sciences.

The conference members listened to and discussed fifteen addresses devoted to the theoretical and experimental studies of tropospheric and ionospheric forward scatter of ultra short waves. As is known, during the last six or seven years many studies of the forward scatter of ultra short waves have been carried out in a number of countries. The investigation of this new form of propagation of ultra short waves is of great practical importance to radio communications, long-distance television transmission, radio navigation, etc.

In his prefatory address Professor A.G.Arenberg, Doctor of Engineering Sciences, surveyed the state of theoretical and experimental studies of the related problems and outlined the principal tasks of the conference.

Five papers were devoted to problems of tropospheric scatter.

P.P.Biryulin's paper discussed an integral equation for the vector potential of scatter field in an environment with a fluctuating permittivity. The sequence of approximations obtained when deriving that equation takes into account the multiple scatters of various degrees. Unfortunately, the related theory has not yet been perfected to a point where it can yield numerical results and can be compared with experimental data. The physical premises of the theory are not either well founded as yet.

V.A.Zverev described in his paper the methods of computing the mean intensity of the scatter of radio waves on random irregularities upon taking into account the width and form of the radiation pattern of the receiving antenna. In this connection, it has been found that the computations of scatter intensity published in literature are correct only when the dimensions of the receiving antenna exceed considerably the correlation scale. The lecturer also reviewed the opportunities for an experimental determination of the correlation function.

D.M.Vysokovskiy's paper was concerned with a critical analysis of the derivation of the general formula for the effective area of the scatter of ultra short waves in the troposphere. The paper furnished a derivation of the formula for power at the reception point in the event of broad antenna radiation patterns, and it described some aspects of reception. With respect to narrow antenna radiation patterns, the paper evaluated the influence of the irregularity of turbulence on the magnitude of the loss in antenna gain. The paper also contained formulas derived for taking into account the influence of refraction on the diffusion propagation of ultra short waves, and demonstrated that it is possible to consider that influence by introducing the effective radius of the Earth into the appropriate formulas. The influence of multiple scatter on the magnitude of field attenuation was appraised, and it was demonstrated that this influence can be ignored in a great majority of cases.

The paper by A.A.Semenov and G.A.Karpeyev described the results of an experimental investigation of rapid fluctuations in the amplitude of signals reflected from two fixed reflectors and received by two spaced receivers. The thus obtained scale of instantaneous values of signal amplitudes proved approximate to the logarithmically normal law, which clashes with the theory of multiple reflectors on the Earth's surface, which was originally adopted as the working hypothesis. This indicates the predominance of the influence of the irregularities in the troposphere itself.

The paper by L.Ya.Kazakov and A.N.Lomakin was devoted to a survey of the principle of performance and design of a radiorefractometer used for measuring the irregularities in permittivity. It also described the problems encountered in the application of measuring methods and the evaluation of data. Preliminary experiments have demonstrated the feasibility of measuring the intensity and dimensions of irregularities. The measurements also have demonstrated the varying character of the distribution of intensities as depending on altitude, and the presence of intensive stratified irregularities on divers altitudes, and also sharp changes in intensity on the threshold of the cloud ceiling.

Ten papers and reports were devoted to the problems of ionospheric forward scatter.

A.N.Kazantsev surveyed the materials of the VIII Plenary Conference on Ionospheric Forward Scatter of Meter Waves, convened by the Comite Consultatif International pour le Radio, and the program of scheduled research in this field, and also he briefly described the principles of the new form of communication constituted by utilizing the reflections of meter waves from the traces of meteors.

Ya.L.Al'pert described in his paper the results of a theoretical analysis of ionospheric scatter of radio waves at a correlation factor having the form of

$$\rho(r) \approx \exp\left[-\frac{r}{l}\right]^2,$$

where l is the scale of irregularity.

Utilization of these results makes it possible to employ the analysis of experimental data for determining the fluctuations of electron density and the optimum dimensions of the scattering irregularities.

B.N.Gershman's paper derived the effective area of ionospheric scatter upon taking into account the turbulent shifting of the ionized gas. The obtained results are comparable with the data of the theory of Buker-Gordon and Villars-Veiskopf.

M.V.Boyenkov surveyed the problems of the use of the forward scatter of 6 - 10 meter radio waves by reflection from ionospheric layers, and the future perspectives

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of this form of communication.

The papers by S.F.Mirkotan and L.A.Drachev, and also by Yu.V.Berezin, described the methods of investigating the irregular structure of the ionosphere at frequency-spaced reception, by means of recording the variations of the phase path of the reflected pulse.

The reports by V.A.Bubnov, A.I.Khachaturov and S.I.Sotnikov described divers cases of long-distance reception of meter waves, reception of programs from foreign television stations, etc. Emphasis was placed on the possible importance of bringing the related studies to the attention of radio clubs and radio amateurs so as to gain en masse actual experimental data about the forward scatter of ultra short waves.

The resolution adopted by the Conference noted the great theoretical and practical importance of developing omnilateral studies of the forward scatter of ultra short waves in the troposphere and ionosphere, and it also formulated concrete proposals concerning a number of scientific and organizational problems.

The resolution pointed out the desirability of coordinating further research in this field by the USSR Academy of Sciences, and of bringing these matters to the attention of the institutes of the said Academy, and especially the Institute of Atmospheric Physics, and the Main Directorate of the Meteorological Service, ministries of communications and radio-engineering industry, and the faculties and laboratories of higher schools.

The Conference approved a motion for publishing a collection of the papers read there.

SEMINAR ON TRANSISTOR ELECTRONICS

From March 14 to March 24 of this year the A.S.Popov Scientific-Technical Society for Radio and Communication Engineering (Section of Transistor Devices and Small Parts), in collaboration with the Polytechnical Museum, convened a ten-day

All-Union seminar on transistor electronics, in Moscow. The seminar was attended by over 600 experts working on electronic matters in plants, designing bureaus, and scientific-technical institutes of various ministries and agencies, including about 300 scientist and scientific-technical workers sent to the seminar from 43 cities.

The lectures read at the seminar were aimed to explain the foundations of the theory of transistor devices and their applications in the radio engineering circuits. These lectures also surveyed the physics of the operation of transistors, methods of computing low-frequency, high-frequency and pulsed circuits, and aspects of operation of these circuits at changes in temperature.

The participants in the seminar observed that the conduct of such undertakings is favorable to the realization of the directives of the XX Congress of the Communist Party of the Soviet Union with regard to the introduction of transistor devices into the Nation's economy. At a time when relations between the enterprises are not yet well organized the conduct of this seminar is doubtless conductive to an exchange of experience.

It is necessary to print as soon as possible the lessons read at the seminar and to send a copy to each participant. Only then can the task posed to the seminar be regarded as completely fulfilled.

In view of the previous experience in convening such seminars, the Governing Board of the A.S.Popov Society has resolved to convene during October-November 1957 a special seminar on the problems of popularizing the use of printed circuits, small parts and ferrites.

A new seminar on transistor electronics is planned to be convened in the next year. The program of that seminar should be so scheduled as to reflect to a great degree the problems of the design and computing of transistorized equipment and of the omnilateral utilization of printed circuits and small parts. The program should envisage a date for the exchange of experience among the participants in the seminar and for the organization of visits to neighboring enterprises.

VIII^th PLENARY ASSEMBLY OF THE INTERNATIONAL RADIO CONSULTATIVE COMMITTEE (CCIR) IN WARSAW*

August 9 - September 13, 1956

Study Group 4

Study Group 4 is concerned with investigating the problems of the propagation of surface radio waves. At its first session the following two Subgroups have been formed: 4A, under the chairmanship of Millington (England); and 4B, under the chairmanship of Herbstreyt (USA).

The area of studies of Subgroup IVA was as follows: 1) Study program and recommendations on "Propagation of Surface Waves over Mixed Routes"; and 2) Study program and paper on "Propagation of Surface Waves over Uneven Terrain".

The area of studies of Subgroup 4B included: 1) Study program on "Influence of Tropospheric Refraction on Frequencies Below 10 mc"; 2) Study program on "Time Variations in the Field Strengths of Surface Waves"; 3) Resolution on "Plotting Curves of Propagation for Frequencies Below 300 kc"; 4) Recommendation on "Presentation of Data on Antenna Radiation"; and 5) A novel question broached by the Czechoslovak delegation, "Determination of the Electrical Parameters of the Earth's Surface".

The recommendation on "Propagation of Surface Waves Over Mixed Routes" has been modified in the sense that the use of theoretical methods is currently recommended in all possible cases and,whenever this is not possible empirical methods may be employed upon considering the limits of their suitability.

A study of the "Propagation of Surface Waves Over Uneven Terrain" served as the subject for a paper describing briefly the results of various theoretical investigations. Due note was taken of the theoretical work on the solution of the problem of the propagation of radio waves over mixed routes. The paper included presentation

* Continuation of report published in Radiotekhnika, Vol.12, No.3, 1957.

of documents containing the solution of this problem upon taking into account the sphericity of the Earth (Godzinskiy, Furutsu), even though this was not presented as yet in a form suitable for practical computations.

The paper on "Influence of Tropospheric Refraction on Frequencies Below 10 mc" appraised the difficulties related to applying the methods of geometrical optics to these frequencies and also to the influence of troposphere on fairly low frequencies.

The paper concerning the resolution on "Plotting Curves of Propagation for Frequencies Below 300 kc" pointed out the necessity for a cautious approach by the CCIR toward the use of these curves for determining field strengths on frequencies below 300 kc, owing to the influence of the ionosphere, which is not considered in these curves.

The recommendation on "Presentation of Data on Antenna Radiation" has been completely modified so as to recommend that data on antenna radiation be presented in the form of field strength or power relationships and also by means of introducing the concept of "simomotive" force.

The novel question of "Determination of the Electrical Parameters of the Earth's Surface", presented by the Czechoslovak delegation, has met with endorsement by the whole Study Group and, after evaluation, was approved.

Study Group 4 had also adopted two proposals for new resolutions. The first resolution expresses the wish that the Secretariate of the CCIR should complement its published atlas of curves for the determination of field strengths of meter waves by adding thereto an appendix explaining the method of computing field strength at equivalent-radius values and soil parameters differing from those for which the curves in the atlas are plotted. The second resolution expresses the wish that the Secretariate of the CCIR should draft and publish a new atlas of curves for determining field strengths at higher frequencies (up to 10,000 mc) and higher antenna heights (up to 20,000 m), which is needed by aviation services. The resolution contains tentative data for the related calculations.

Study Group 5

Study Group 5 is concerned with investigating the problems of the tropospheric propagation of radio waves. At its first session these two Subgroups have been formed: 5A, under the chairmanship of Allen (USA), and 5B, under the chairmanship of Rauden (England).

The area of studies of Subgroup 5A comprised: 1) Study program on "Measuring the Field Strengths of Radio Signals"; 2) Problem "Measurement of Field Strength at Direct Proximity of Obstacles"; 3) Recommendation on "Optimum Methods of Expressing Field Strength at Pulsed Transmission"; 4) Recommendation on "Field Strength Measurements. Types of Receiving Antennas and Equipment for Every Frequency Band"; and 5) Recommendation on "Field Strength Measurements. Influence of Local Conditions on Interpretation and Accuracy of Field Strength Measurements".

The area of studies of Subgroup 5B comprised: 1) Study program on "Curves of Tropospheric Propagation for Distances Much Greater than Line-of-Sight Distances"; 2) Study program on "Tropospheric Propagation of Radio Waves; 3) Problem "Radio Wave Propagation Data Necessary for Wide-Pand Radio Systems"; 4) Recommendation on "Data Presentation at Studies of the Tropospheric Propagation of Radio Waves"; and 5) Recommendation on "Curves of Tropospheric Propagation for Distances Much Greater than Line-of-Sight Distances".

The study program and recommendation on "Curves of Tropospheric Propagation for Distances Much Greater Than Line-of-Sight Distances" led to the adoption of a resolution for postponing the revision of these curves until after the VIIIth session of the CCIR, and not during that session, because of the enormous scope of the related work. Also there was formed an international working group which should within a year or a year and half complete the work on the revision of tropospheric curves upon utilizing all the currently available data. The revision of these curves will consist in the plotting of new curves for higher frequencies and higher time percentages, in determining the corrections to be introduced for propagation over sea surfaces, and in an appraisal of the spread of field strength values stemming from the diversity of climatic conditions and antenna heights.

The study program on "Tropospheric Propagation of Radio Waves" has been subjected to some modifications. Considering that irregularities of the troposphere cause the forward tropospheric scatter of ultra short waves, the newly modified program turns attention to the necessity of a broad study of these irregularities (their intensity, form, dimensions, etc.) with the aid of special sensitive, low-inertia radio-engineering and meteorological instruments. Attention is also turned to the necessity of investigating the correlation between meteorological conditions and the conditions of the tropospheric forward scatter of ultra short waves. Considering that at such scatter the value of field strength is substantially affected by meteorological conditions, the program proposes the compilation of charts of isolines of the vertical gradient of air permittivity, based on meteorological data obtained by means of radiosondes, as the first step in the development of radioclimatology.

The question of "Wave Propagation Data Necessary for Wide-Band Radio Systems" in its original version pertained only to the systems operating on the basis of the propagation of ultra short waves approximately within the line-of-sight limits.

On the last session of Study Group 5 the United States delegation introduced a proposal for a new, important study program and recommendation under the overall name of "Radio Transmissions Utilizing the Irregularities of the Troposphere". The new study program includes the investigation of problems of great practical importance to the design of communication systems utilizing the phenomena of the tropospheric forward scatter of ultra short waves; the investigation of fading and of the influence of meteorological conditions, determination of the maximum band of transmitted frequencies, antenna efficiency and gain at the use of spaced-antenna reception.

Study Group 6

Study Group 6 is concerned with investigating the ionospheric propagation of radio waves. The chairman of this Group, Dellinger, was absent, and Bailey (USA) deputized for him.

Five Subgroups were formed at first, and afterward a sixth one was added.

Subgroup 6A (Chairman: Smith-Rose, England), was entrusted with the following issues:

1) Study program and paper on "Selection of Basic Index of Ionospheric Propagation"; 2) Recommendation on "Forecasting the Solar Activity Index"; 3) Study program, paper. recommendation, and resolution on "Short-Term Ionospheric Prognoses"; 4) Study program and paper on "Basic Information for Prognoses of Ionospheric Propagation"; and 5) Paper on "A Centralized Organization for Rapid Exchange of Information on Propagation".

Subgroup 6B (Chairman: Millington, England), investigated the following problems:

1) Study program on "Radio Wave Propagation on Frequencies Below 1500 kc"; 2) Study program and paper on "Ionospheric Propagation on Frequencies of 30 - 300 mc"; 3) Study program on "Pulsed Transmission Experiments at Inclined Incidence"; 4) Recommendation on "The Study of Absorption in the Ionosphere"; and 5) Study program on "Non-Linear Effects in the Ionosphere".

Subgroup 6C (Chairman: Crichlow (USA)), reviewed a number of problems relating to atmospheric interference: 1) Study program and recommendation on "Measurements of Atmospheric Radio Interference"; 2) Recommendation on "Review of Data on Atmospheric Interference; and 3) Recommendation on "Counters of Local Lightning Flashes".

Subgroup 6D (Chairman: Grosskopf, West Germany) was concerned with the problem of fading at ionospheric propagation (study program).

Subgroup 6E (Chairman: Roberts, International Frequency Registration Board) investigated chiefly the issues raised by the International Frequency Registration

Board, and also the question of protecting the frequencies applied in radio astronomy.

Finally, Subgroup 6F investigated a question shared with Study Group No.7: "The Utilization of Modulated Transmissions on Standard Frequencies for Evaluating the Reliability of the Prognoses of Radio Wave Propagation".

Subgroup 6A

England and New Zealand proposed that the basic index for ionospheric prognoses should be constituted by the so-called "ionospheric number of sunspots", i.e., by some given number of sunspots determined by measuring the critical frequencies of the F_2 Layer (index J_{F_2}). However, the small working group formed after discussion (consisting of delegates of England, USA and the Soviet Union) adopted a revision proposed by the delegate of the Soviet Union. This revision consists in that the R index (Wolf number) should continue to be used for long-term prognoses, because it is the simplest and most homogeneous index in the circumstances, but the possibility of a future application of J_{F_2} and other indexes should at the same time be explored.

In the revised version of the lecture paper the index of countries and institutions furnishing long-term prognoses has been expanded by adding the Soviet Union (NIZMIR, Ministry of Communications) thereto. A like addition has also been made with regard to short-term prognoses and rapid exchange of information on propagation.

A recommended method of improving ionospheric prognoses consists in the compilation of world charts of critical frequencies of the F_2 Layer for specific hours of the day (every 2 hours, round the clock). Study Group 6 should investigate the possible advantages of such charts. Basically no new recommendations have been made arent the problem of short-term ionospheric prognoses.

The problem of solar index prognoses has been investigated chiefly by the Director of the CCIR, Professor van der Pol. However, it was stated that "the methods explored by the Director of the CCIR and based on the technique of autocorrelation have not led to a completely satisfactory method of forecasting solar activity".

In a technical circular issued by van der Pol, the Director notes that the shortly expected maximum solar activity (which perhaps will occur as soon as in the middle of 1957) will be very high and will apparently exceed all hitherto observed maximum limits.

Subgroup 6B

Subgroup 6B investigated in detail the results of a multilateral experimental study conducted by the European Broadcasting Union (With the participation of England, France, Holland, Finland, Yugoslavia, and other countries) and concerning the propagation of medium and long waves. The obtained experimental curve of field strength at right is shaped much lower than the curve accepted on the Cairo Meeting but it coincides satisfactorily (at distances up to 2000 km) with the curve accepted by the International Frequency Registration Board. The latter curve was plotted on the Copenhagen Meeting by a study group under the chairmanship of Professor V.N.Kessenikh (USSR).

The principal questions investigated by Subgroup 6B pertained to the long-distance propagation of meter waves - especially by means of their forward scattering from ionospheric irregularities. The study program points out that the scatter is successfully used for radio communications at distances of 1000 - 2000 km, especially in the arctic and subarctic areas. The program postulates future investigation of the mechanism of such scatter - the seasonal, diurnal and momentary changes in field strength, the influence of solar and geomagnetic activity, the most suitable types of modulation, etc.

The Chairman of Study Group 6, Bailey, who leads studies on forward scatter of radio waves in the USA, delivered a special lecture on this subject, on the conference. He pointed out the special importance of this form of communication in polar areas, and he emphasized that the worse are the conditions of ordinary propagation of radio waves in the polar aurora zone, the better are the conditions for effecting communication by means of the wholly reliable forward scatter of meter waves. Bailey noted that this kind of radio communication, which extends over distances of the order of many thousands of kilometers, has a good perspective of development, and he pointed out the actual existence of an experimental line between America and England via Iceland. The paper of the Czechoslovak delegation mentioned the possibility of super-long-distance radio communication on frequencies much higher than the minimum applicable frequencies, by propagating radio waves between the lower boundary and the maximum ionization level of a single ionized layer.

The program of research on return-inclined sounding ("return scatter") actually constitutes a new study program containing a number of interesting proposals: investigation of divers forms of scatter (from the Earth, from the E Layer); determination of the scatter coefficient as a function of frequency; nature of the scattering surface and the angle of incidence of the ray; and investigation of the change in antenna radiation pattern at considerable distances from the transmitter, etc.

Subgroup 6C

On the VIIth Plenary Assembly of the CCIR in London the Soviet Delegation did not accept any of the resolutions pertaining to atmospheric radio interference, in view of the negative Soviet attitude toward the American charts of atmospheric interference.

In the interval of time between the VIIth and VIIIth Assemblies a special study group consisting of representatives of England and USA compiled new charts of atmospheric interferences based on a greater number of measurements, although with a broad application of interpolation. The Soviet delegation made the motion that these new charts "can be used, with great reservations". In the accepted compromise phrase the words "with great reservations" were changed to "with some reservations".

Subgroup 6D

This Subgroup issued a paper describing the studies conducted with regard to the topic, including also the studies conducted by the Soviet Union, and supplying a brief resume of the law of the distribution of field strength variations.

Subgroup 6E

Subgroup 6E has been principally concerned with questions posed by the IFRB to the CCIR. The first question was whether it is feasible to use the curves of maximum applicable frequencies plotted in Mexico City on the basis of Paper No.462 of the Bureau of Standards.

The delegate of the Soviet Union transmitted to Subgroup 6E a report furnishing a comparative analysis of some results of the computations of the IFRB as made by the American, Czechoslovak, and Soviet methods, and demonstrating that the novel version of the American method yields, as a rule, lower results. The first question posed by the IFRB remains pending.

The second problem was that of selecting a method of computing field strength. On the motion of the delegate of the Soviet Union, Subgroup 6E adopted unanimously a resolution - later ratified by the Plenum - for organizing a special study group consisting of representatives of six powers. including the USSR, Rumania, and Czechoslovakia, for investigating all the existing methods of computing field strength.

The labors of the CCIR and, especially, of the three study groups concerned with radio wave propagation, manifest tangibly the spirit of international cooperation (which in a number of cases was violated by the representatives of the United States alone /sic/). A preponderant majority of problems was solved cooperatively, which makes the Warsaw Conference much different from the London one. Naturally, this was largely owing to the active position of the Soviet delegation, which submitted a great number of papers and introduced a number of valuable resolutions.

Study Group 7

The labors of Study Group 7, which is concerned with the organization of the world service of standard frequencies and time signals and the techniques of the transmission and reception of these signals, have been somewhat advanced at the VIIIth Plenary Assembly.

The paper compiled on basis of the data on Question No.87 includes a tabulation of the characteristics of the stations transmitting standard frequencies and time signals, which comprises also data on the Moscow station. The number of the stations transmitting standard frequencies and time signals has risen from six to ten during the interval between the VIIth and VIIIth Plenary Assemblies. In addition, four stations are in the project state and should be set into operation between the end of 1956 and the spring of 1957. There are also four other stations transmitting standard frequencies and time signals with a high degree of stability and accuracy, but not operating on the frequencies set aside for this kind of transmission.

A new study program and two novel questions have considerably expanded the program of labors in this field.

The study program concerns the investigation of the possibilities for improving the transmission of standard frequencies and time signals for the purpose of reducing mutual interference among standard-frequency transmitters. This program undertakes a study of the possibility of reducing interference by means of utilizing special forms of transmission such as single sideband with suppressed carrier or two sidebands with suppressed carrier.

A novel question has also been posed: the investigation of the causes of the decrease in the stability and accuracy of the reception of standard frequencies and time signals. It is also recommended that research be initiated on determining the optimum forms of time signals and receiving-equipment characteristics that would ensure maximum accuracy of reception.

From the engineering viewpoint an interesting bit of information supplied by

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Study Group 7 concerned the frequency standard based on the utilization of the atomic resonance of cesium, which was developed in the USA. According to the advertised data, this standard ensures a stability of no less than $5 \cdot 10^{-10}$ in the course of the entire life of device.

Study Group 8

Study Group 8 is concerned with problems of the methodology and technical characteristics of measuring equipment of the stations of the International Control Service. The fundamental trend of the labors of this Group is dictated by the needs of the IFRB purporting to facilitate its work by perfecting the control of the stability of frequencies, occupation of the frequency spectrum, and width of radiation band.

The activities of Study Group 8 during the VII[th] Plenary assembly were expressed by two study programs and two questions.

The activities of Study Group 8 during the VIII[th] Plenary Assembly were conducted by two Study Subgroups: 8A, whose labors were centered on the question of "Automatic Control of the Radiation Spectrum" and on the study program on "Measurements of Radiation Spectra by Control Stations", and 8B, which examined the problems relating to the study program on "The Accuracy of Field Strength Measurements by Control Stations" and "Measurements of Frequencies Over 50 mc by Control Stations".

Eighteen documents were presented on the above problems by Study Group 8. The labors of the Group have become somewhat advanced, and a result of the VIII[th] Plenary Assembly was the issuance of 10 documents including a modification of the recommendation on "The Accuracy of Field Strength Measurements by Control Stations", which contains a table of the values of the accuracy of measurements which should be satisfied by the control devices used for measuring frequency.

The recommendation that was formulated concerning the automatic control of the occupancy of the radio frequency spectrum cites tentative technical features of automatic control equipment corresponding to the features recommended by Soviet organ-

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izations (State Inspectorate of Electrocommunications and the Scientific Research Institute of the Ministry of Communications).

A complement to the study program on "Frequency Measurements by Control Stations" refers to the necessity of comparing the station radiation spectra measured directly on transmitters with those that are measured by long distance or a control station, developing new equipment for control of wide radiation bands (about 10 mc in a range of over 30 mc), and conducting a preliminary examination of the possible measurements of wave-form characteristics on control stations for systems for which this is of basic importance (for instance, for television).

Study Group 9

Study Group 9 is concerned with studying the problems pertaining to radio relay lines. The CCIR has begun its labors in this field comparatively recently and no recommendations were adopted until the VIIth Assembly. The Group investigated ten problems and one research program. The Group was divided into four Subgroups and, moreover, each Subgroup created two or three working groups for drafting the projects of recommendations and reports. The Commission prepared 34 new documents, including: 23 recommendations, 6 reports, 1 resolution, 2 new questions, and 2 new research programs.

The principal contents of these documents are summed up as follows:

International linkage of frequency-multiplex radio relay lines and standardization of their basic features: This involved the determination of the characteristics of radio relay lines which should be standardized on the divers forms of international connection. Considering that such standardization has been deemed to be as yet premature, there have been adopted recommendations determining the advisable parameters of the basic characteristics of the following three forms of radio relay systems: group-channel, intermediate-frequency, and radio-frequency. These recommendations concern such characteristics as the number of channels, values of intermediate frequency, frequency deviation, frequency division or six trunks operating simultaneously on one line, and so forth. There was also adopted a recommendation for determining the allowable frequency instability on the transmitters of radio relay lines.

Quality of communications on radio relay lines: The adopted recommendation determines the so-called standard hypothetical circuits for radio relay lines. These circuits (for lines with a 12 - 60 channel capacity and for lines with capacity of over 60 channels), which have a definite length and structure, should provide guidance to the designers of radio relay systems.

Another recommendation to be adopted established the allowable noise level in telephone channel at the end of a 2500-km standard circuit. This recommendation determines the mean noise power per hour, and it appears to be temporary, because the allowable noise power per shorter units of time has not yet been determined, and the investigation of this problem should be continued.

All these recommendations are consonant with the proposal presented by the Soviet delegation. The recommendation determining the allowable noise level includes an appendix describing a Soviet-proposed method of computing the basic parameters of equipment.

Auxiliary equipment for radio relay lines: the adopted recommendations determine the methods of ensuring reserve equipment on radio relay lines, the application of special control currents at transmission of television signals and 600 telephone conversations, and a method for measuring the quality of radio relay line channels.

A new question has been approved for investigation, for the purpose of drafting a recommendation on the methods of effecting service communications and on the requirements posed to service channels.

Time-division-multiplex (pulse-modulation) radio relay lines: No agreement could be reached about the related problems and thus no recommendations on the basic characteristics of such lines could be drafted. Even the rough drafts of recommendations which were prepared at a meeting of Study Group 9 in September of 1954 were not approved either. However, the Group did compile a list of basic characteristics which should be standardized at international linkage of pulsed radio relay lines. It was proposed that at such international linkage the recommendation to be adopted should specify that the receiving party should accept the terms of the transmitting party.

Study Group 10

The labors of this Group were oriented toward two basic trends: radio-frequency broadcasting and sound recording for international exchange of programs. Accordingly, the following documents pertaining to radio-frequency broadcasting have been examined and approved.

1. Recommendation on ultra short wave FM broadcasting. The discussion ranged about the documents presented by the German Federal Republic, France and England, concerning the frequency deviation, protective ratios and minimum field strength that are necessary for satisfactory reception. The papers presented by the above-named countries proposed a frequency deviation of ± 75 kc. This value has also been supported by the US delegation. The Soviet delegation demonstrated the pertinency and some advantages of the application of a frequency deviation of ± 50 kc, and this value has been unanimously approved together with the deviation of ± 75 kc.

Also adopted were the Soviet-recommended field strength standards for low-noise zones - 250 mw; for towns - 1 mw; and for large cities - 3 mw.

2. The Polish administration proposed an antenna which makes it possible to reduce spurious radiation. At a session of the Group the antenna was approved and a report was drafted to recommend that this antenna be listed in the CCIR antenna book. Also, a program on practical research in spurious radiation of classic directional antennas had been formulated and unanimously approved.

3. Investigation of the problem of ensuring the reception zone with the nec-

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essary field strength by means of antennas operating at some angle to each other. Such an antenna system makes possible a more uniform covering of the reception zone owing to the expansion of the energy radiation lobe in the main direction, provided that the said zone have considerable scope and width. Also, this system makes it possible to obtain a more uniform field at the use of a single frequency.

With regard to sound recording for international exchange of programs, the following resolutions were examined and adopted: 1) adoption of the types of magnetic tape adapters proposed by the European Union and supported by the Soviet delegation; 2) during the examination of the problems of sound recording on magnetic tape, there was adopted a Soviet proposal that tape data be recorded on the side of the record representing the continuation of the unused part of tape.

The Group reached no specific agreement about the standards for the width and tolerances of magnetic tape.

As for movie-tape recording for exchange of television programs, the Group adopted a proposal made by the European Union that sound and image be both recorded on a single tape by means of magnetic or optical track.

Study Group 11

This Group (chairman: Esping, Sweden) was concerned exclusively with television problems.

Its basic labors were conducted by five Subgroups, concerned with the following problems: 11A - color television standards; 11B - formulation of requirements for long-distance transmission of television signals on communication channels; 11C - black-and-white television standards; 11D - quality of television images; and 11E - protective ratios at planning the distribution of television stations and frequency channel division.

Color Television

The sessions of the Group at the VIIIth Plenary Assembly of the CCIR in Warsaw were preceded by demonstrations of various color television systems in the USA, England, Holland and France in the spring of 1956.

The results of these demonstrations were cited in the chairman's report. It had been supposed that an All-European color television system would be settled upon in Warsaw. The discussion revealed that, at research on the development of a color television system, most countries gave preference to an adaptation of the NTSC system to European conditions.

The Soviet delegation advised that at present in the USSR principal attention is turned to developing a fully combined color television system with a single subcarrier and quadratic components, because this system appears to be the most thoroughly tested one.

The English and French delegations were categorically opposed to the selection of any specific television system on the present Plenary Assembly, basing their opposition on the insufficiency of the hitherto conducted research in this field. The French delegation insisted on the conduct of comparisons of the NTSC system with the systems developed in France.

The discussion continuing throughout the three sessions of the Subgroup yielded no definite results and was halted in view of the difficulties involved in a comparison of the qualitative and technical indices of the NTSC system, which is already in operation, with analogous indices of the French-developed systems which are still in the laboratory-experiment stage. The United States delegation did not participate actively in the discussion, upon advising that the USA does not propose to alter its viewpoint as to the selected NTSC system. It was pointed out that in this year several hundred thousand television sets will be manufactured, and that their output will continually increase.

In view of the absence of sufficient data for the adoption of definite color television standards, and also in view of the absence of an unanimity of opinion on this problem, the VIIIth Plenary Assembly of the CCIR adopted no definite resolutions as to an European color television system. It was merely resolved that this problem should be re-examined on the subsequent session of the Commission.

It is assumed that, after that session, and before 1958 is over, there will be convened the second European conference on a revision of the Stockholm plan for the division of frequency channels for television and ultra short wave FM broadcasting in the I, II and III bands. This conference should arrive at a new plan for frequency division in the IV and V bands. At that time, Study Group 5 of the CCIR should have ready some more precise data on radio wave propagation in these bands.

Requirements Posed to Communication Channels for Long-Distance Television Transmission. An evaluation of these requirements, conducted by Subgroup 11F, was based on the set of standards drafted in 1955 in Brussels, upon taking into account the new proposals presented by England, Holland, Switzerland, and the German Federal Republic. This included the formulation of new requirements as to the qualitative indices of the television channels of radio relay and cable lines. This document also included methods for measurement of non-linearity of the amplitude characteristic, amplification factor of communication channel, and transient-response tolerances in the range of low and medium video frequencies, and so forth. All these indices were established only with regard to transmission of black-and-white television signals.

Upon a proposal by the Soviet delegation and the representatives of the CCIF, there was adopted a resolution for regarding the above requirements as a desirable aim and not as standards which should be maintained on the existing or projected communication systems. The possibility of adopting these requirements as standards for a hypothetical control line should, in accordance with that resolution, be investigated by a combined CCIR-CCIT team.

Black-and-White Television Standards

Subgroup 11C examined a report including information on the black-and-white television standards established in various countries. This also included evaluation of the basic parameters of the OIR, and the Subcommission adopted a resolution for including these parameters in the CCIR's report on black-and-white television standards. Also, this Subcommission drafted a paper complementing the recommendation (on television standards) by the point on the gamma-characteristic coefficient of television transmitters which, on taking into account the modulation characteristic of the receiving tube, should be below unity.

The new research program envisages an investigation of problems relating to correction of the distortions of television signals at single sideband transmission (quadratic distortions, phase distortions in transmitters and receivers, etc.).

Quality of Television Images

Subgroup 11D appraised the data presented by various countries with regard to evaluating the quality of television images. The Subgroup approved the resolution of the Soviet delegation that this problem be reviewed anew on considering that the methods of evaluating the quality of television images should not depend on the television standards used.

As for the protective ratios at planning of the distribution of television stations and frequency channel division, Subgroup 11E examined a report on protective ratios in television. It adopted the proposal that the standards specified in this report can be applied only at planning the distribution of black-and-white television stations. Moreover, it was pointed out that the determination of protective ratios should take into account the frequency responses of television receivers, because the curve of protective ratios cited in the report was plotted without considering these responses.

The new research program envisages the investigation of problems relating to

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Study Group 12

Study Group 12, which is concerned with tropical-zone broadcasting, and whic[h] has Mr. Balig (India) as its chairman, reviewed Recommendation No.87 (London, 195[...] and established new power limits for transmitters operating on tropical broadcast frequencies (below 5060 kc): not over 10 kwt for distances of up to 0 km, and not over 30 kwt for distances of up to 800 km.

For transmitters operating in the tropical zone on frequencies of over 5060 (i.e., in the customary radio-frequency broadcasting bands), the Group cancelled power limitation and recommended that the power applied there be the same as that established on the Mexico-City conference on radio-frequency broadcasting.

Furthermore, the Group approved three reports on: noise in radio broadcastin[g] bands; improved methods of computing the field intensity of the space wave on tropical-zone broadcast transmitters; and the design of receiving antennas for tropical-zone broadcasting. The Group also approved the study of the new problem of feeding tolerances in tropical-zone broadcasting.

Study Group 13

The problems investigated by this Group include:

1. Publication of service codes for the international telegraph service. 2. Identification of radio stations. 3. Marine equipment for identification. 4. Classification of bearings for short wave and ultra short wave radio-direction fi[nd]ing. 5. Technical characteristics of marine ultra short wave FM equipment. 6. Testing of an emergency radio-telegraph 500-kc receiver on marine vessels. 7. L[oud]speaker equipment for ship and coastal stations. 8. Preventing of noise in radi[o]

determining the gain in protective ratios at the use of the system employing the offsetting of the carrier frequencies of television transmitters at a considerable difference between the carrier frequencies of offsetting stations.

Study Group 12

Study Group 12, which is concerned with tropical-zone broadcasting, and which has Mr. Balig (India) as its chairman, reviewed Recommendation No.87 (London, 1953) and established new power limits for transmitters operating on tropical broadcasting frequencies (below 5060 kc): not over 10 kwt for distances of up to 0 km, and not over 30 kwt for distances of up to 800 km.

For transmitters operating in the tropical zone on frequencies of over 5060 kc (i.e., in the customary radio-frequency broadcasting bands), the Group cancelled the power limitation and recommended that the power applied there be the same as that established on the Mexico-City conference on radio-frequency broadcasting.

Furthermore, the Group approved three reports on: noise in radio broadcasting bands; improved methods of computing the field intensity of the space wave on tropical-zone broadcast transmitters; and the design of receiving antennas for tropical-zone broadcasting. The Group also approved the study of the new problem of feeding tolerances in tropical-zone broadcasting.

Study Group 13

The problems investigated by this Group include:

1. Publication of service codes for the international telegraph service. 2. Identification of radio stations. 3. Marine equipment for identification. 4. Classification of bearings for short wave and ultra short wave radio-direction finding. 5. Technical characteristics of marine ultra short wave FM equipment. 6. Testing of an emergency radio-telegraph 500-kc receiver on marine vessels. 7. Loud-speaker equipment for ship and coastal stations. 8. Preventing of noise in radio

reception on ships. 9. Radio-telephone observation of the 2182-kc distress frequency, and equipment for that frequency. 10. Alarm signal for use on the 2182-kc distress frequency.

Five Study Subgroups were formed. The Soviet delegation was able to participate in the work of only two of these Subgroups.

1. With regard to the problem of identification of radio stations, five documents were presented at the session. The Soviet delegation attempted to reject the documents recommending the methods of identification not acceptable to the USSR, without opposing the problem itself. The Soviet proposal was not seconded, and the Soviet delegation reserved its opinion as to three of the documents while voicing in advance its positive attitude as to the question of further research.

2. With regard to the problem of the equipment and classification of short wave and ultra short wave radio-direction finding, these two reports were approved: "Short Wave and Ultra Short Wave Direction Finders", and "Marine Identification Devices". The nature of the issued documents (reports) speaks for itself: the session could not develop any concrete recommendations on the given problem, because the documents presented differ in their appraisals of the accuracy of radio direction finding, their observations were not based on unified methods, and the results they obtained are not identical.

Group 13 examined and adopted the following: recommendation on common technical characteristics of ultra short wave FM marine equipment, and the new question of spurious radiations of ultra short wave FM equipment. Document No.761 introduced the new question of investigating the degree of necessity of applying the international selective-call system in mobile marine stations using the ultra short wave band, and the advantages of this system.

Study Group 14

1. The principal labors of this Group were centered on compiling a vocabulary

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of radio engineering terms in accordance with a recommendation of the CCIR (London) Assembly. The chairman of the Commission, Professor Tullio Gorio, submitted to it a draft of the vocabulary he compiled, which contained both the definitions formulated by the CCIR and the definitions formulated by the International Electrical Engineering Commission, and also definitions taken from the dictionaries of various national organizations. After examination, the Commission resolved, on basing itself on the draft of vocabulary compiled by Professor Gorio, to divide all terms into the following four groups: Group B - terms requiring no definition; Group C - terms whose definitions are cited in the draft of the dictionary of the International Electrical Engineering Commission; Group D - terms whose definitions are taken from the dictionaries of various national organizations; and Group R - terms whose definitions will be supplied by the CCIR (that is to say, by the CCIR's vocabulary).

For purpose of information, the terms in Groups C and D will be published in the form of an appendix to the CCIR's vocabulary. All investigating commissions of the CCIR have been enlisted in the work on compiling the definitions of the CCIR vocabulary. The completion of the work of the vocabulary was undertaken by the "national correspondent" of the French Administration, who will enlist "national correspondents" of England or the USA for the purpose of hastening the English-language version. The Group approved the report on this matter.

With regard to the adoption of the universal decimal classification for the standard classification of documents and articles concerning radio, the Group composed and approved a new report which pointed out that there is not enough related data for approving this system at the present Plenary Assembly, but the matter should be resolved at the next Plenary Assembly of the CCIR.

Commission for Technical Assistance to Technologically Underdeveloped Countries

The Administrative Council of the International Telecommunications Union proposed at its XI^{th} Session (Geneva, 1956) that the next Plenary Assembly of the CCI should explore the ways and means for providing technical assistance to countries in which communications engineering is poorly developed.

In compliance with the proposal of that Administrative Council and the letter of the CCIR Director, the VIII[th] Plenary Assembly set up a temporary commission for provision of technical assistance. The leader of the Soviet delegation, Z.V.Topuriya, was elected as the Commission's chairman. The Commission includes representatives of the delegations of 23 countries.

The agenda of the Commission includes, in accordance with the resolutions of the XI[th] Session of the Administrative Council of the International Electrocommunications Union, examination of the ways and means of providing technical assistance to technologically underdeveloped countries for the purpose of a harmonious development of the internal and international electrocommunications of these countries, and compilation of resolutions on this problem.

All work on provision of technical assistance is organized by the Secretariate General of the International Electro-Communications Union. The Secretariate of the CCIR (and also of the CCIT and CCIF) confines itself only to consultation and inquiries posed to organizers, who are members of the Administrative Council of the International Telecommunications Union. The role of the International Telecommunications Union in the work of providing technical assistance is a purely consultative one.

The Soviet delegation drafted and submitted to the Commission for Technical Assistance several proposals on the forms and organization of technical assistance within the framework of the CCIR.

The Soviet delegation recommended that a permanent committee be set up and attached to the International Telecommunications Union for the purpose of coordinating the work on the provision of technical assistance. In making this proposal the Soviet delegation proceeded from the assumption that the character of the participation of the ITU in the work on provision of technical assistance is an unsatisfactory one.

because the CCIR has been making consultations only after all basic resolutions had already been approved by the committees of the UN.

During discussion of the problems of providing technical assistance, the Commission approved the proposals of the Soviet delegation with regard to the forms of technical assistance which can be provided by the CCIR. However, there arose a difference in opinions as to the organization of technical assistance work in the system of the International Telecommunications Union, and no agreement could be reached in this respect. The Soviet delegation agreed with the opinion of a number of other delegations that, in view of the inadequacy of data on this problem submitted to the Commission for examination, further investigations are necessary. The Commission unanimously entrusted its revision group with the task of drafting a resolution on the provision of technical assistance upon considering: a) necessity of establishing a temporary joint CCIR-CCIT commission for investigating the problems of providing technical assistance within the framework of the International Telecommunications Union; and b) formulation of a program for study of this problem by the said temporary joint commission.

The revision group has drafted the resolution of the Commission for Technical Assistance and the program for study of the problems of provision of technical assistance by the IEU.

As drafted by the revision group and approved on the Plenary session of the Commission for Technical Assistance, the resolution contains proposals on the establishment of a temporary committee consisting of seven members of the CCIR and CCIT for the purpose of exploring the ways and means of providing technical assistance and drafting, for the benefit of the Administrative Council, a recommendation on improving this assistance with regard to electrocommunications. A complement to this resolution describes the related study program.

The VIIth Plenary Assembly of the CCIR adopted the resolution on the formation of a temporary joint commission consisting of four representatives of the CCIR and

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CCIT and two directors, with the aim of continuing the studies of the ways and means of improving the technical assistance of the ITU. The Assembly also had confirmed the program for the study of the problems of providing technical assistance as based on the proposals of the Soviet delegation concerning the forms of technical assistance within the framework of the CCIR.

Final Plenary Sessions

The work of the CCIR sessions reached its apogee at the period when the mass of documents drafted by the study groups was submitted for examination to the Assembly. A part of the documents was not totally approved by the study groups themselves, and the differences in opinions were reviewed by the Assembly.

Altogether, over 1000 documents were presented to the Assembly. The great extent of activities of the study groups and the increased number of documents and the related considerable overburdening of the work of the Secretariate during the convention of the Assembly, have necessitated the raising of the question of reorganizing the activities of the CCIR. In this connection, the resolutions adopted propose a system of work similar to that applied by the committees of the CCIF and CCIT, where the study groups ply their activities uniformly throughout the three-year intervals between plenary assemblies instead of concentrating their activities in the period directly preceding a plenary assembly.

For the purpose of greater economy in expenditures, it was proposed that the study groups should convene their meetings not separately but jointly, to comprise mutual problems. The proposal recommended that the following experimental order of presentation of documents should be introduced: the contributions of the participants of the study group should be sent to the chairman of the corresponding study group interested in examining them (one copy), and to the Director of the CCIR for translation, printing and transmission to the interested members of the study group (three copies). The result of this experiment is to be communicated at the next, IXth Plenary Assembly of the CCIR. Henceforth, the plenary sessions of the CCIR would be concerned only with the reports of the representatives of investigating commissions, while all preliminary documentation should be dealt with on intermediate sessions of the commissions and distributed only by participants in these commissions. Therefore, to obtain all documents, it is necessary to inform the CCIR of the commissions in which a given national administration will participate.

Also, there was adopted a special resolution on reducing preliminary documentation and size of all documents.

The documents that are of theoretical interest only and bear no direct relation to the undertaken problems and research programs, and the papers containing detailed original material, should not be submitted to the CCIR. Instead, only brief annotations of such materials should be sent to the CCIR for purposes of translation and publication. Original-language copies of such documents can be distributed directly by a given administration to those expressing a desire to receive them. These documents should contain a minimum amount of special mathematical formulas or designations and experimental data. Most of the documents approved at the VIIIth Plenary Assembly were also approved by the Soviet delegation, which has reserved its opinion only with regard to several documents that were not acceptable to it for various reasons.

On the concluding plenary session the head of the United States delegation invited the IXth Plenary Assembly to convene its sessions in the USA.

The head of the Soviet delegation invited the XIth Plenary Assembly, which is to be held in 1958, to convene in the Soviet Union.

NEW BOOKS

Arenberg,A.G. - Propagation of Decimetric and Micro Waves. Sovetskoye Radio Publishing House. Moscow. 1957. 304 pages. Price 17 r. 80 k.

This book surveys in a quite accessible form the fundamental problems relating to the propagation of decimetric and micro waves in various conditions, and it also cites some formulas and information of practical nature. The book sheds light on the latest explorations of this field. The explanations are made without any complex mathematical procedures.

The book is chiefly designed for readers who are actually working with decimetric and micro wave equipment. It will be also useful for students and instructors in the proper institutions of learning.

Grodiyev,I.I., Lakernik,R.M., and Sharle,D.L. - Foundations of the Theory of Communication Cables and Their Production. Gosenergoizdat Publishing House, Moscow-Leningrad, 1956. 480 pages. Price 10 r. 25 k.

A survey of the problems of computing, design and production of coaxial and balanced-line cables. Separate chapters are devoted to a popular exposition of the theory of the transmission of energy along cables and to an explanation of the electrical processes occurring in cable circuits.

Kulikovskiy,A.A., Boloshin,I.A., and Potryasay,V.F. - Fundamentals of the Designing of Radio Receivers by Students. Edited by A.A.Kulikovskiy. Gosenergoizdat, Moscow-Leningrad, 1956. 328 pages. Price 7 r. 45 k.

An exposition of the fundamentals of the design of radio receiving devices for various wave bands. The book is approved as a student manual for higher schools and faculties.

Diffraction of Electromagnetic Waves on Some Rotating Bodies. A collection of articles. Sovetskoye Radio Publishing House, Moscow, 1957. 176 pages. Price 4 r. 85 k.

The collection contains following articles:

1. Fok,V.A. - Theory of Diffraction from a Rotating Paraboloid; 2. Belkina,M.G., and Vaynshteyn,L.A. - Radiation Characteristics of Spherical Antenna Surfaces; 3. Belkina,M.G. - Radiation Characteristics of a Stretched Rotating Ellipsoid; 4. Belkina,M.G. - Diffraction of Electromagnetic Waves on a Disk.

These articles are concerned with the rigorous theory of the diffraction of electromagnetic waves on conducting bodies.

The book is designed for radio physicists and radio engineers concerned with superhigh frequencies.

Zvorykin,V.K., and Morton,D.A. - Television: Electronic Problems of the Transmission of Color and Monochrome Images. Translated from the English; edited by Professor S.I.Katayev. Foreign Literature Publishing House, Moscow, 1956. 780 pages, + 1 inset. Price 46 r. 10 k.

An exposition of the physical foundations of television, including the description of the principal units and circuits of television equipment; a survey of the problems of the theory and practice of color television and of the optimum utilization of television set-ups.

Some Problems of the Theory and Computing of the Elements of Radio Reception Circuits. A collection of articles edited by A.P.Beloyusov. Oborongiz Publishing House, Moscow, 1956. Transactions of the Sergo Ordzhonikidze Order-of-Lenin Moscow Institute of Aviation. Bulletin 65. Price 7 r.

This collection contains articles by:

Beloyusov,A.P. - Sorting Out the High Frequency Signal and Noise at Detection; and The Calculation of the Complete Ultra Short Wave Autotransformer Input Circuit; Protopopov,A.S. - Energy Relationships at Combined Detection of Signal and Noise; and The Calculation of a Single Amplifier Circuit Under Matched Line Load; and Vol'pyan,V.G. - A Diagram of the Input Circuit of a

Receiver.

Fradin, A.S. - Superhigh Frequency Antennas. Sovetskoye Radio Publishing House, Moscow, 1957. 648 pages. Price 17 r.

This book surveys wireless diffraction antennas of the horn, lens, mirror, slot, and dielectric types. A description of the design of such antennas, and an explanation of the principles of their operation, as well as of the theory and methods of their computing. There is also an appended bibliography which can be useful for antenna designing.

The book is designed for students of higher schools, degree candidates and radio engineers.

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