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DESIGN PROBLEMS IN A MODERN HIGH-LEVEL MODULATION SYSTEM

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[Numbers in brackets in the text refer to the bibliography. Figures referred to are appended.]

In the present-day high-power transmitter employing plate modulation, the quality and, to a considerable extent, the power characteristics are governed by the modulation system. Recently, this has consisted of a multistage push-pull audio-frequency amplifier, the last stage of which is the modulator, which works in a Class B system using very high voltages and currents in the tubes. The modulation system embraces a feedback system consisting of the primary circuit of the output or modulation transformer.

Many questions concerning the theory, the schematic layout, and peculiarities of the system have already been dealt with previously [1,2]. In the course of recent practical work on the construction and regulation of high-level modulation systems, many of these questions have received further attention. New problems also arose, necessitating the introduction of a number of changes in the layout of individual stages. This article gives a survey of some of the constructional problems in the present-day modulation system.

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1. METHODS FOR SUPPRESSION OF DYNATRON OSCILLATION IN THE MODULATOR'S GRID CIRCUIT

It is known that the presence of a downward drop in the dynamic characteristic curve of the grid current of modulation tubes is the cause of parasitic dynatron oscillation. It originates in the dynamic system, and lasts during that part of the voltage-amplification period when its instantaneous values run into the downward bend of the curve. In this case, the frequency of parasitic oscillations is equal to the fundamental frequency of the grid circuit which is determined by the parasitic capacity of the system, the distributed inductance, and the distributed capacity of the submodulation [driver] transformer.

To eliminate the possibility of dynatron oscillation it was necessary that the resistance in the modulator grid circuit at absolute resonance frequency was less than the equivalent negative resistance of the space between the grid circuit and the tube filament in the zone of the drop in the curve.

This was previously achieved by two methods:

1. The secondary winding of the submodulator transformer was shunted (in each arm) with a small active resistance. However, although the drop in the dynamic characteristic of the grid circuit is comparatively small in extent, it may have a very large grid-plate transconductance (for tube G-433 in particular, it reaches 3 - 5 mA/V). Consequently, a very low resistance shunt is required, which greatly increases the submodulator load. It therefore follows that this method is irrational.

2. The space between the grid circuit and the filament of the modulator tube is shunted with an antidynatronic kenotron. The equivalent conductivity of the kenotron is equal to ghe grid-plate transconductance of the modulator tube. In practice, triodes are frequently used, with the grid and anode connected together. The total tube current in this connection (i.e., where $e_a = e_g$) is determined by the equation,

$$i_a = S\left[e_g + D\left(e_a - E_{a0}\right)\right] = S(1+D)\left[e_a - \frac{E_{g0}}{1+D}\right], \qquad (1)$$

where $E_{g0} = DE_{a0}$. (E_{a0} is the driver-plate voltage).

The equivalent tube resistance will be

$$R_1 = \frac{1}{S(1+D)} = \frac{R_i}{1+\mu}$$

i.e., $\frac{1}{(1+\mu)}$ of its internal resistance in the triode connection.

It is obvious that the shunting affect of the antidynatronic kenotron is most effective when the section of greatest transconductance of its characteristic curve conforms with the drop in the grid-current characteristic. The kenotron's unilateral conductivity and the limited value of its saturation current are favorable factors here, because they reduce the energy which it requires outside the dynatronic zone. In practice, however, the characteristic of the kenotron is much closer to the beginning of the coordinate than is the drop in the modulator's grid current, and the latter can be said to correspond to the kenotron's saturation zone where the transconductance is extremely small (Figure 1). Consequently, it was necessary to increase the filament [voltage] in the kenotrons and to resort to including several tubes in parallel.

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For antidynatronic valves it was intended to use copper-oxide rectifiers whose advantages included a large transconductance and the absence of a filament. However, it was not possible to use copper oxide because the absence of saturation in it meant that the peak of the total current in the grid circuit was sharply raised, and consequently the load on the modulator was increased.

As an antidynatronic measure, S. V. Person proposed an active low-resistance shunt, and the inclusion in series with it of a reactance with a small impedance in the frequency zone of probable oscillation and with a sufficiently large impedance in the fundamental band of the amplified frequencies [3]. Such an arrangement would naturally only be used where the resonance frequency of the grid circuit lies outside this fundamental band. Thus, for example, if the resonance frequency lies beyond its upper limit, a sufficiently small capacitive load would have to be included in series with the active shunt resistance (Figure 2).

It is evident that such an antidynatronic shunt would not absorb a large amount of power (the further the frequency of possible oscillation from the amplified band, the smaller it would be). It should be noted that if the resonance resistance of the modulator's grid circuit is large, the ultrasonic [sic] harmonics of the grid current reaching it may have very large amplitudes (considerably exceeding the excitation amplitude at the fundamental frequency). The use of the shunt, which greatly lowers the resonance impedance, reduces the danger of arc-overs and increase the stability of the modulation system's operation.

The efficiency of the kenotron's antiparasitic action can also be increased if its characteristic is moved to the right, thus causing it to operate with the minimum angle of intersection (Figure 1, dotted curve). To achieve this, a supplementary negative grid bias, automatic and using internal current, should be applied to the kenotron's plate (Figure 3a). In cases where a triode is used as an antidynatronic tube, the grid bias can, according to N. G. Fr[illegible]'s suggestion, be obtained by using the current of one grid, for which purpose the automatic grid bias can be included between the plates and the grid (Figure 3b). In this case, the antidynatronic tube's instantaneous voltage value at the grid will be:

$$e_g = e_a - E_{ag} \quad ,$$

where $E_{ag} = R_g I_{a=}$ (R_g is the automatic grid bias' resistance; $I_{a=}$ is the constant component of the antidynatronic tube's grid current).

Similarly, for the tube's total current we get

$$i_{\Sigma} = S(1+D)\left[e_a - \frac{E_{\Sigma o}}{1+D} - \frac{E_{ag}}{1+D}\right] \qquad (2)$$

A comparison of this expression with equation (1) shows that the presence of a supplementary grid bias E_{ag} between the grid and the plate transfers the equivalent diode's characteristic to a zone of greater positive potentials at its plate; the slope of the characteristic curve is not altered. The equation for a family of ideal grid-current characteristics [4] can be used in the calculation of the grid bias E_{ag}, t

$$i_g = S(e_g + \mu_g e_a - E_{go}), \qquad (3)$$

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where $\mu_g = \frac{\partial e_g}{\partial e_a}$ for a constant i_g. The instantaneous voltage at the antidynatron tube plate is shown by

$$e_a = E_{mg} \cos \Omega t + E_g ,$$

where U_{mg} and E_g are, respectively, the excitation amplitude of the modulator tube, and its grid bias. Having established the value of e_a in equation (3) and considering that $e_g = e_a - E_{ag}$, we get a system of two transcendental equations in which the unknowns are E_{ag} and θg,

$$\cos \theta_g = \frac{E_{g0} + E_{ag}}{(1+\mu_g) E_{mg}} - \frac{E_g}{E_{mg}} \qquad (4)$$

and

$$E_{ag} = S(1+\mu_g) R_g E_{mg} \alpha_{0g} (1 - \cos \theta_g). \qquad (5)$$

Here, θ_g is the angle of intersection of the grid-current impulse; $\alpha_{0g} = f(\theta_g)$ is the coefficient of the constant component of this impulse. The solution of this system of equations can be obtained graphically.

Equation (3), and consequently also the calculation of E_{ag}, is true if

$$e_{a\ max} \leqslant \frac{e_g - E_{g0}}{-\mu_g} , \text{ i.e., } E_{ag} \leqslant (E_{mg} + E_g)(1+\mu_g) - E_{g0} .$$

The time constant of the automatic grid-bias circuit must satisfy the ratio,

$$C_g R_g < \frac{1}{F_H} ,$$

where F_H is the lowest modulating frequency.

Only the grid current of the antidynatronic tube, which is less than 1/5 of its total current, flows through the grid-bias resistance in the layout described above, and therefore this resistance must be larger than it is when it is included directly in the plate circuit (Figure 3a); similarly, the capacity C_g can be reduced. The possibility of making a substantial reduction of this capacity is one of the advantages of this layout.

In practice, when constructing a high-level modulation system containing two G-433 tubes in each arm, the most effective way to combat antidynatronic oscillation is to introduce a capacity-resistance shunt and an antidynatronic tube with a supplementary grid bias into the circuit between the grid and the plate. The frequency of dynatron oscillation in this layout lies in the 30 - 40-kc band.

The oscillographic representation of the modulator's grid-current impulses at different signal levels (i.e., different modulation coefficients) is shown in Figure 4a.

The use of a purely active shunt was quite impossible owing to the submodulator's heavy overload. When using simple antidynatronic kenotrons, four B-8-500 tubes (or N-600 in a diode connection) with an increased filament voltage (up to 20-20.5 v) should be included in the arm in order

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to reduce oscillation. The inclusion of a shunt, formed in each arm by a series resistance of 250 ohms and a capacity of 0.05 μF, permitted the retention in the arm of two M-600 tubes with a filament voltage of 18.5 - 19 v. The impulse diagram for the total grid current (together with the current in the antidynatronic tubes) which conforms with this case is shown in Figure 4b. Eventually, when the automatic grid bias circuit (with a resistance of 30,000 ohms and a capacity of 4 μF) was included between the grid and each M-600 tube, it was possible to reduce the filament voltage of these tubes to 17 v. The grid-current impulse took the form shown in Figure 4c.

It should be noted that when the gas [sic] in the modulator tube is ignited, a large part of the plate voltage is applied to the grid. An air discharger [sic] is included in parallel with the space between the grid and the modulator-tube filament in order to protect the apparatus. However, in normal operating conditions, the precise regulation of a discharger to DC voltages of the magnitude of several kilovolts is difficult, and therefore a tube with a high electrical stability should be used as an antidynatronic measure. The electrical stability of the component parts of the grid circuit should not be lower than the plate supply voltage.

II. PHASE COMPENSATION IN A SUBMODULATION TRANSFORMER

It is known that in order to ensure a deep and stable feedback over a broad frequency band, the phase displacement in various components of the circuit involved must be reduced as far as possible.

In the layout of the present-day modulation system, the submodulation transformer is the basic source of large phase displacements at high frequencies. Considering the simplified equivalent circuit of the transformer and taking into account the leakage inductance and load capacitance, it might be expected that the submodulator would give a reverse phase of 180 degrees. In fact, however, due to the sharply defined effects of resonance peaks in the transformer, the resultant phase displacement at high frequencies is considerably in excess of this amount. A considerable improvement in the phase characteristics of the submodulation transformer can be achieved by the direct coupling of the primary and secondary windings in the high-frequency zone. For this purpose, the primary and secondary windings are connected through an impedance which offers a very high resistance at medium and low frequencies, and a correspondingly low resistance at high frequencies. This impedance can of course, be put in series with the combined capacity C_0 and resistance r_0 (Figure 5).

If the required symmetry of windings is ensured (i.e., the condition where the connected ends have the same potential with reference to the earth), the limiting phase displacement which the submodulation transformer can give is 90 degrees (in cases where the ends are cross connected it may reach 180 degrees).

However, this layout does permit the use of a well-known phase compensation effect for the submodulator at certain frequencies. An equivalent circuit for one arm of the submodulator under conditions described above can be seen in Figures 6a and 6b. The following symbols are used:

$\bar{R}_i$ is the internal resistance of the submodulator tubes,
$\bar{Z}_0$ is the mutual impedance between the transformer's primary and secondary windings,
$\bar{Z}_2$ is the load impedance in the secondary winding.

Kirchhoff's equation for this layout is

$$E_1 = i_1(R_i + i\omega L_1) - i_2 i\omega M - i_3(i\omega L_1 - i\omega M),$$
$$0 = -i_1 i\omega M + i_2(\bar{Z}_2 + i\omega L_2) - i_3(i\omega L_2 - i\omega M),$$
$$0 = -i_1(i\omega L_1 - i\omega M) - i_2(i\omega L_2 - i\omega M) + i_3(\bar{Z}_0 + i\omega L_1 + i L_2 - 2i\omega M)$$

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In addition, we calculate the voltage at the output $E_2 = i_2 Z_2$; the full leakage inductance pertaining to the transformer's primary winding $L_{s1} = L_1 - \frac{M^2}{L_2}$; the coefficient of transformation $n = \sqrt{\frac{L_2}{L_1}}$.

In a zone of sufficiently high audio frequencies, where the impedance $\bar{Z}_0$ is of the same order as the leakage impedance, the ratios $\frac{L_{s1}}{L_1}$ and $\frac{Z_0}{\omega L_1}$ can be disregarded since they are quantities of another order.

Thus, after a number of modifications we get an equation for the subpodulator's frequency characteristic

$$\frac{E_2}{nE_1} = \frac{\bar{Z}_0 + i\omega L_{s1} n}{R_i(1-n)^2 + \bar{Z}_0 + R_i n^2 \frac{\bar{Z}_0}{\bar{Z}_2} + i\omega L_{s2} n^2 \left(1 + \frac{\bar{Z}_0 + R_i}{\bar{Z}_2}\right)} \qquad (6)$$

This expression is not true for medium and low frequencies in the audio spectrum. For frequencies not exceeding 30 - 40 kc, the influence of the modulator's full input capacity can be disregarded, and it may be considered that the lead in the submodulation transformer's secondary winding is produced for the most part by the capacity-resistance antidynatronic shunt formed by the capacity C_2 and the resistance r_2.

Having established in formula (6) the values $\bar{Z}_2 = r_2 + \frac{1}{i\omega C_2}$ and $\bar{Z}_0 = r_0 + \frac{1}{i\omega C_0}$, we get the following expression for the frequency characteristic

$$\frac{E_2}{nE_1} = \frac{1 - \omega^2 L_{s1} C_0 \left(n + \frac{C_2}{L_{s1}} r_0 r_2\right) + i C_0 r_2 \left(\frac{r_0}{r_2} + \frac{C_2}{C_0} - n\omega^2 L_{s1} C_2\right)}{1 - \omega^2 L_{s1} C_2 \left[\frac{r_2 R_i}{L_{s1}} C_0 (1-n)^2 + n^2\left(1 + \frac{C_0}{C_2}\right) + \frac{r_0 r_2}{L_{s1}} C_0 \left(1 + n^2 \frac{R_i}{r_2}\right)\right] + i\omega C_0 r_2 \left[\frac{r_0}{r_2} + \frac{C_2}{C_0}\left(1 + n^2 \frac{R_i}{r_2}\right) + \frac{R_i}{r_2}(1-n)^2 - \omega^2 L_{s1} C_2 n^2 \left(1 + \frac{r_0}{r_2} + \frac{R_i}{r_2}\right)\right]}$$

As a basis for this equation, a family of frequency-amplitude and frequency-phase characteristics were worked out for the following corresponding parameters:

$$\frac{R_i}{r_2} = 1.5; \quad r_2 \sqrt{\frac{C_2}{L_{s1}}} = 0.45; \quad n = 0.34.$$

The calculation was made

The calculation was made for different ratios $\frac{r_0}{r_2}$ and $\frac{C_2}{C_0}$. The curves in Figure 7 were obtained. The values of the relevant parameters lie within the limits of quantities normally found in practical use. In particular, they satisfy a submodulator with the following data:

$$R_i = 375\,\Omega; \quad L_{s1} = 0.0154\,H; \quad r_2 = 250\,\Omega; \quad C_2 = 0.05\,\mu F.$$

In this case, $\omega\sqrt{L_{s1} C_2} = 3$ corresponds to the highest frequency in the audio range $F_{max} = 10\,Kc$.

The frequency-amplitude and frequency-phase characteristics of a similar submodulator without phase compensation is shown by the dotted curves in Figure 7. These curves only show the general tendency of the continued accumulation of phase displacement with increase of frequency. In fact, the reversal of phase takes place considerably more rapidly on account of resonance peaks. Curves 2 and 4 show that in the layout described above, the phase angle increases for frequencies of up to 12 - 14 kc, after which it gradually decreases. If the series resistance is reduced (curves 3 and 5), the submodulator can induce a compensating phase angle in frequency bands above 20 kc; the negative phase displacement at frequencies of 14 - 15 kc is somewhat increased. When a phase-compensating system is introduced, the submodulator's resultant frequency-amplitude characteristic is very satisfactory. Frequency distortion not exceeding 2 decibels (curves 2. 3 and 4) can be obtained at the highest frequencies in the audio spectrum, whereas at frequencies of 20 - 30 kc a considerable restricting effect may result.

III. SUBMODULATOR OPERATION IN THE VARIABLE DISPLACEMENT SYSTEM

Until recently, the Class A system was generally accepted for the operation of submodulator tubes. The power deficiencies of this system are well known, and its use in the submodulator noticeably reduces the overall efficiency of the station. Another disadvantage of the Class A system is the impossibility of making good use of power tubes owing to the consumption and dissipation, at their plates, of an amount of power exceeding the maximum permissible quantity. Nevertheless, this is particularly necessary for submodulators in high-level modulation systems where the modulator grid current impulses are very large. The above-mentioned disadvantages also limit the reduction of direct current in the modulator tubes inasmuch as they would require an increase in excitation, i.e., a further increase in the number of submodulator tubes.

For these reasons, it was essential to transfer the submodulator to a system which would permit the wide use of its power tubes with a low power consumption at zero modulation. One of the possible solutions for this problem was the use of the so-called "variable displacement" system [5], the essence of which is that the operating point in the dynamic characteristic of a tube is changed in accordance with the signal level, so that at a maximum value of the signal, the operating point is located along the linear section, and at zero modulation, in proximity to the lower bend of the curve.

With such a system, the average power consumption by the submodulator is substantially reduced. The position of the operating point corresponding to the maximum modulation level can be selected much higher up than when using Class A, because in this case only that part of the power which supplies the submodulator is dissipated at the plate. If there should be no protracted maximum modulation and the maximum signal level is momentary, an even wider use of power tubes is possible. It should be noted that in this case the submodulator works continuously in Class A, i.e., its internal resistance, as a source of electromotive force, does not increase.

The adjustment of the operating point on the characteristic of the tube is achieved by means of introducing a supplementary displacement, proportional to the amplitude of the signal, into the submodulator's operating circuit. With this aim in view, a special detector [sic] should be introduced into the circuit, similar to that used in the modulation of an FM oscillator with variable carrier signals. Time-constant requirements for an increase of voltage in the supplementary displacement and for its pulsations could, however, be considerably easier. Indeed, when the system is regulated by a push-pull Class A amplifier, voltage pulsations in the supplementary displacement are fed in phase into the grid circuit of both arms, and thus can only produce distortion when there is a considerable amount of asymmetry in the amplifier. The time taken to increase the supplementary displacement when there is a sharp increase in signal level is only the time taken to change over the amplifier from the Class B system (in which it rapidly gains its new level) to the Class A system.

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For submodulators in high-level modulation systems, the required supplementary change in displacement may reach a sizeable amount (300 - 500 v). In calculating this amount, the drop in voltage at the plate limiting resistance of the modulator tubes should be taken into account. In this case, the signal voltage should be supplied to the detector by means of a special separating series amplifier (Figure 8). The grid voltage of this amplifier can be supplied by means of a voltage divider from any preceding stage of audio amplification (including the grid circuit of the submodulator itself, see Figure 8).

The load resistance R of the detector can be much greater than the internal resistance of the tube, and therefore the voltage amplitude at a position halfway along the secondary winding of the transformer T is in practice equal to the value of the supplementary displacement (Figure 8). In this case, the transformer is loaded by a resistance equal to 2R. The supplementary displacement voltage is supplied to the submodulator tube grid through a unique filter. The series element of this filter is the grid leak resistance R_g of the submodulator, and the shunt element is the blocking capacity C_g in series with the internal resistance Ri of the tubes of the preceding stage; the resistance R_1 is in parallel with the plate load R_a of this stage. Thus, the time constant for the increase of supplementary displacement, even in the absence of capacity at the detector output, cannot be made less than the amount,

$$C_g\left(R_g + \frac{R_i R_a}{R_i + R_a}\right) \approx C_g R_g,$$

which may reach 100 or more milliseconds. This however does not impair the operation of the system.

Experimental verification confirmed this principle. Measurement of distortion with sinusoidal tone modulation showed that in practice, when reducing the direct current six to eight times, the variable displacement system does not give an increase in the nonlinear coefficient in comparison with the normal Class A system. Oscillograms of the input and output voltages of the modulation system (Figure 9), taken when supplying a manually controlled [interrupted] tonal signal, showed the complete absence of any sort of additional nonstatic effects governed by the time constant of the increase in supplementary displacement (the final reading was 0.02 sec). Lastly, a lengthy experiment on the operation of a high-level modulator in a variable-displacement system showed that the performance of this system is completely reliable and satisfactory for the tubes, and allows a substantial economy in power.

Disadvantages in the system include the additional requirements in connection with the alternating-current power supply from the rectifier to the filter of the submodulator's plate feed (if it has a separate filter) and the necessity for the inclusion of a fixed grid-bias voltage tap. However, these requirements must also be satisfied in the other system with a Class B modulator. This latter system has in addition many other defects as for example: an increase in the modulator tube's equivalent internal resistance and consequently an increase in distortion governed by the tube load; the appearance of additional distortion at high frequencies, produced by nonstatic effects in the plate circuit; the necessity for increasing the excitation amplitude for the submodulator, etc.

IV. SYSTEM OF PRELIMINARY STAGE AMPLIFICATION

A special feature of a preliminary-stage amplification system with feed-

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back, already described in our work [1], is the necessity of having a supplementary supply in the stages for undistorted amplification. This requirement is caused by the possibility of overloading these stages with in-phase components of even harmonics contained in the feedback voltage at both high and low frequencies. In the latter case, the appearance of even harmonic voltage in the reverse-feed circuit is caused by the filter impedances in the modulators plate supply.

In addition, the nonlinear character of the submodulator's load produced by the modulator tubes' grid current causes a characteristic distortion in the shape of the output signal. The feedback compensates for this distortion by supplying to the stages a preliminary amplification which corresponds to the compensating impulse; the excitation curve for these stages is derived, for deep modulation, from this extended and peaked form (Figure 10). A case occurred in practice, where the ratio of the peak excitation value at 100 percent modulation to the corresponding value at 50 percent modulation was 3:1. Where a preliminary amplification stage failed to ensure undisturbed amplification of the sharply peaked excitation curve, the nonlinear distortion sharply increases, and a further increase in modulation level cannot usually be obtained.

The impossibility of calculating this peak excitation value beforehand with sufficient accuracy made it necessary to design preliminary stages with a supplementary feed for amplification. When constructing the resistance circuit of these stages, the good use of voltage amplifier tubes necessitated an increase of the plate supply voltage. If this is undesirable, the layout shown in Figure 11 will be extremely useful. The maximum phase displacement which it gives at low frequencies is determined by the expression

$$\varphi_{max} = \operatorname{arctg} \sqrt{\frac{r_1}{r_2}} - \operatorname{arctg} \sqrt{\frac{r_2}{r_1}},$$

where $r_1 = \frac{(R_i + R_1)R_2}{R_i + R_1 + R_2}$, $r_2 = \frac{R_1 R_2}{R_1 + R_2}$.

(e.g., when $\frac{R_1 + R_2}{R_i} = 8$ and $\frac{R_1}{R_2} = 1$, we get $\varphi_{max} \approx 3^{\circ}$)

Thus, a considerable increase of voltage at the tube plates is in practice obtained in the absence of supplementary-phase reversal. By using this circuit in one of the stages in a high-level modulation system, the limits of undistorted amplification were extended by 60 – 70 percent.

V. THE INFLUENCE OF FEEDBACK IN THE MODULATION SYSTEM ON THE BACKGROUND OF A MODULATED OSCILLATOR

In conclusion, the reader's attention is drawn to the influence of feedback in the modulation system on the background created by the oscillator tubes' supply sources.

Owing to the fact that the feedback is usually produced from the primary circuit of the modulation transformer, it was considered until recently that it

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was not involved in the parasitic modulation set up by the modulated oscillator's supply sources. (This remark does not apply to certain layouts in new transmitters where there is a supplementary selective feedback for the suppression of the most outstanding background components.) This theory is, however, not entirely true, inasmuch as it is evident that the modulation level of this parasitic modulation chiefly depends on the size of the equivalent resistance of the modulator tubes in relation to the secondary circuit. The latter, as is known, is in its turn determined by the feedback level.

Figure 12 shows an equivalent circuit which permits the evaluation of the oscillator's parasitic modulation set up by voltage ripples in the plate supply. Besides the main L-type filter $L_f C_f$ the layout includes a second filtering unit, formed by the inductance of modulator choke coil L_m and the oscillator's equivalent resistance. The oscillator's resistance is in parallel with the circuit formed by C_{bl} (modulator's blocking capacity) and R''_{im} (equivalent internal resistance of modulator tubes in relation to the secondary circuit) in series. Usually, the capacitive reactance of C_{bl} is much less than R"im at ripple frequencies of Ω_0 ; in the absence of feedback, or with a deep feedback of one order [sic] or less, R"im is much greater than R_g. Thus, the presence of deep feedback may increase the filtration of the second unit two or three times (because $\Omega_0 L_m \gg R_g$).

However, calculations show that this is only true when there is a very large rectifier pulsation. With the usual six-phase rectifier circuit and a normal filter (taking nonstatic processes, demodulation etc. into account) the pulsation at the output of the latter is not great, and the supplementary filtration produced by the modulator choke coil is more than sufficient.

Figure 13 shows a circuit which illustrates the influence of the modulator's equivalent internal resistance on an oscillator's background, created by the AC supply to the filaments or by the parasitic modulation of the excitation voltage. It is evident from this circuit that for audio frequencies, the oscillator plate circuit impedance which reduces parasitic modulation is formed by two parallel branches. The first, consisting of the modulator choke coil and the plate supply filter joined in series, has an exceedingly large reactive resistance at background frequencies. The second is formed by the series connection of R"im and C_{bl}, and as was shown earlier, its impedance is determined by the size of R"im. Thus, at parasitic modulation frequencies the plate circuit resistance chiefly depends on the size of R"im. When it is small, i.e., when there is deep feedback, the background caused by the grid circuit or the oscillator filament is much stronger than when there is no feedback.

It is particularly important to bear these remarks in mind during the practical analysis of the causes of background in transmitters employing plate modulation. An investigation of the background produced by the oscillator was carried out by extinguishing or excluding the modulator tubes and including a resistance, equal to the equivalent internal resistance of these tubes at zero modulation, in parallel with the modulation transformer. This was verified during regulating work on the high-level transmitter.

We will examine briefly the effect of the modulator and feedback tubes' internal resistances on the nonlinear distortions in transmitters at low audio frequencies. As a result of the modulator working in the Class B system, sizeable even harmonic voltages of low modulation frequencies are formed in the filter of the plate supply (usually common to moderator and oscillator). These voltages, which in practice reach 10 - 15 percent of the rectified voltage modulate the oscillator, and the transmitter's nonlinear coefficient at low frequencies would therefore attain at least the same quantity. However, owing to the presence, mentioned above, of the supplementary filtering unit (L_m, R_g, C_{bl}, and R''_{im}) the quantity of even harmonic voltage actually reaching the oscillator plate is several times smaller. This is true even

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for very high-level transmitters whose modulation choke coils have a correspondingly small inductance. Naturally the presence of a feedback in the modulation system considerably increases filtration of these harmonics.

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[Appended figures follow]

Figure 1

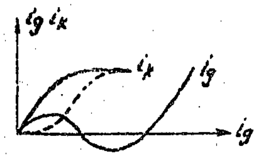


Figure 2

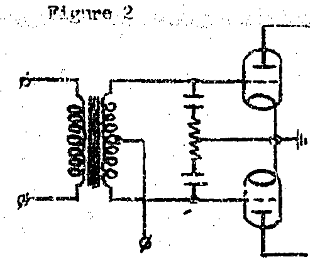
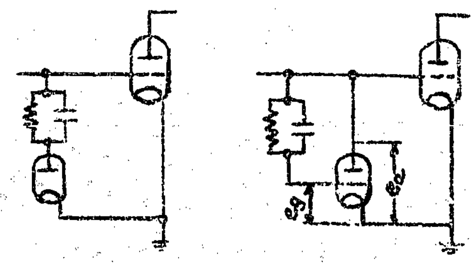
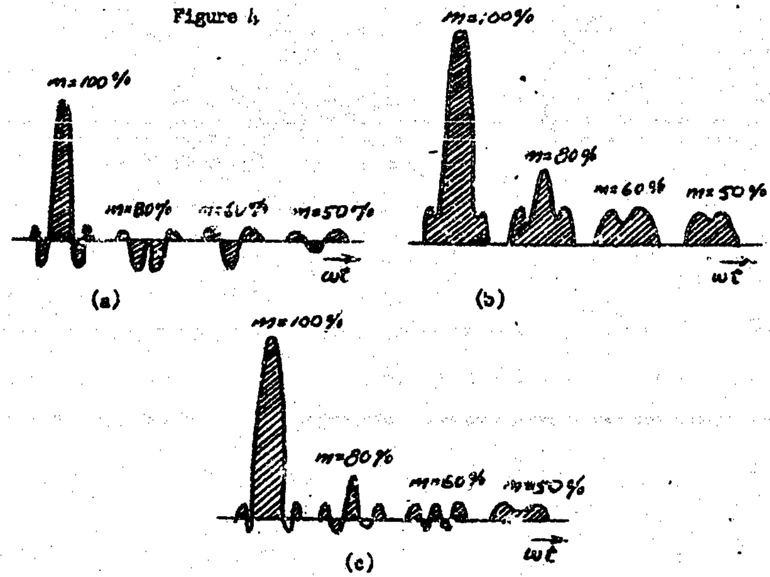


Figure 3



(a) (b)

Figure 4



(a) (b)

(c)

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Figure 5

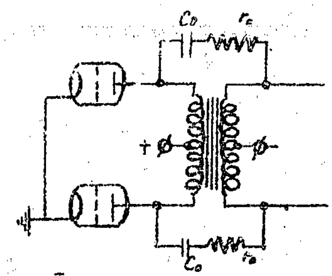
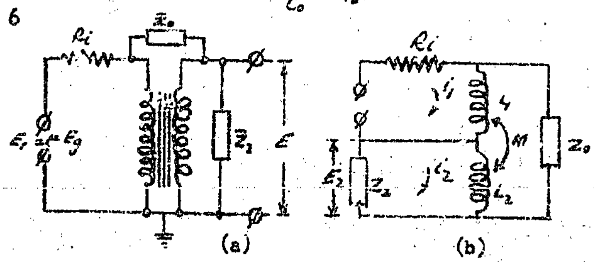


Figure 6



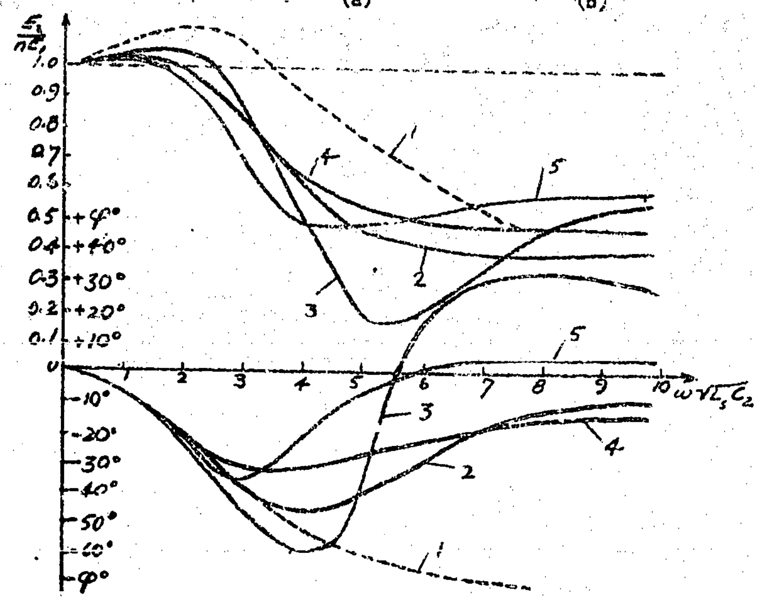


Figure 7 Submodulator's frequency-amplitude and frequency-phase characteristics
where $n = 0.34$; $\frac{R_i}{r_2} = 1.5$; $r_2\sqrt{\frac{C_2}{L_s}} = 0.45$

1) $r_0 = \infty$ $C_0 = 0$ 4) $\frac{r_0}{r_2} = 4$ $\frac{C_2}{C_0} = 5$

2) $\frac{r_0}{r_2} = 4$ $\frac{C_2}{C_0} = 10$ 5) $\frac{r_0}{r_2} = 2$ $\frac{C_2}{C_0} = 5$

3) $\frac{r_0}{r_2} = 1$ $\frac{C_2}{C_0} = 10$

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Figure 8

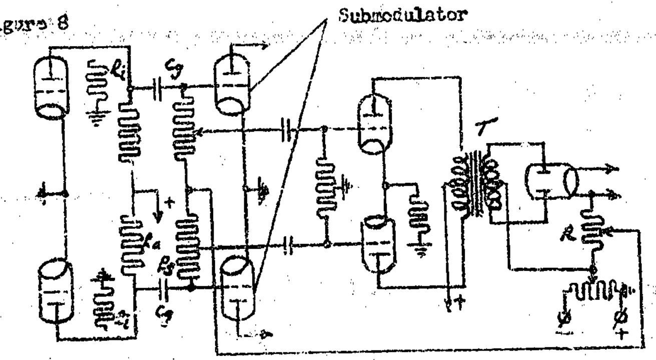
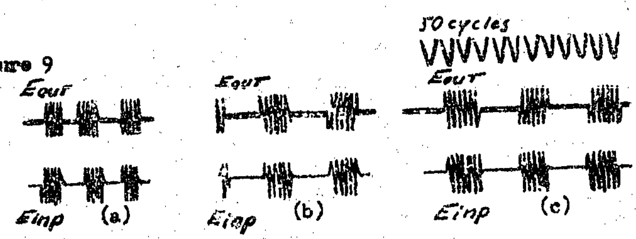


Figure 9



(a) Class A, direct current 70 mA, modulation frequency 200 cycles

(b) Variable displacement system, direct current 12 mA, modulation frequency 200 cycles

(c) Variable displacement system, direct current 6 mA, modulation frequency 200 cycles

Figure 10

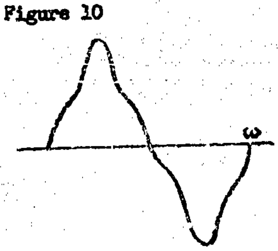


Figure 11

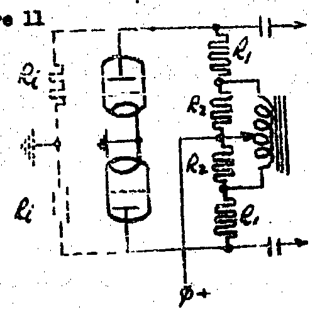


Figure 12

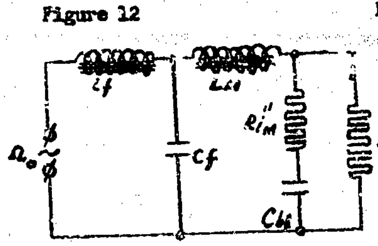
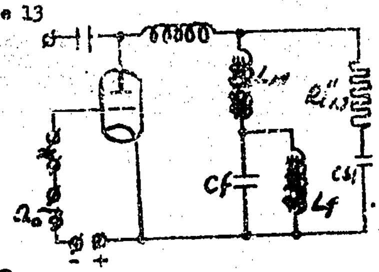


Figure 13



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