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MODULATION CIRCUIT ARRANGEMENT

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Fig. 1

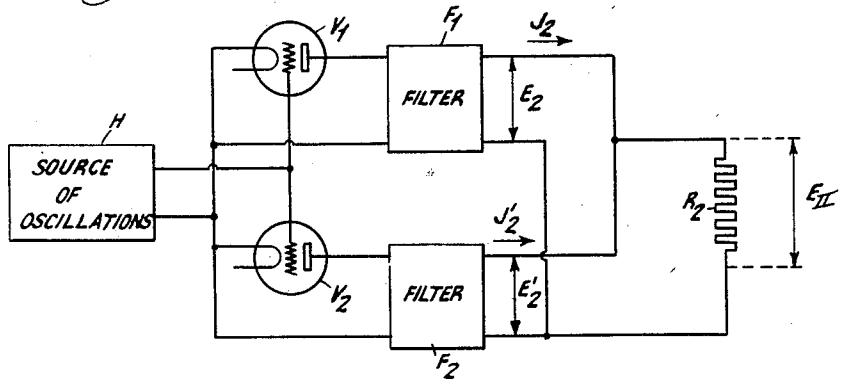
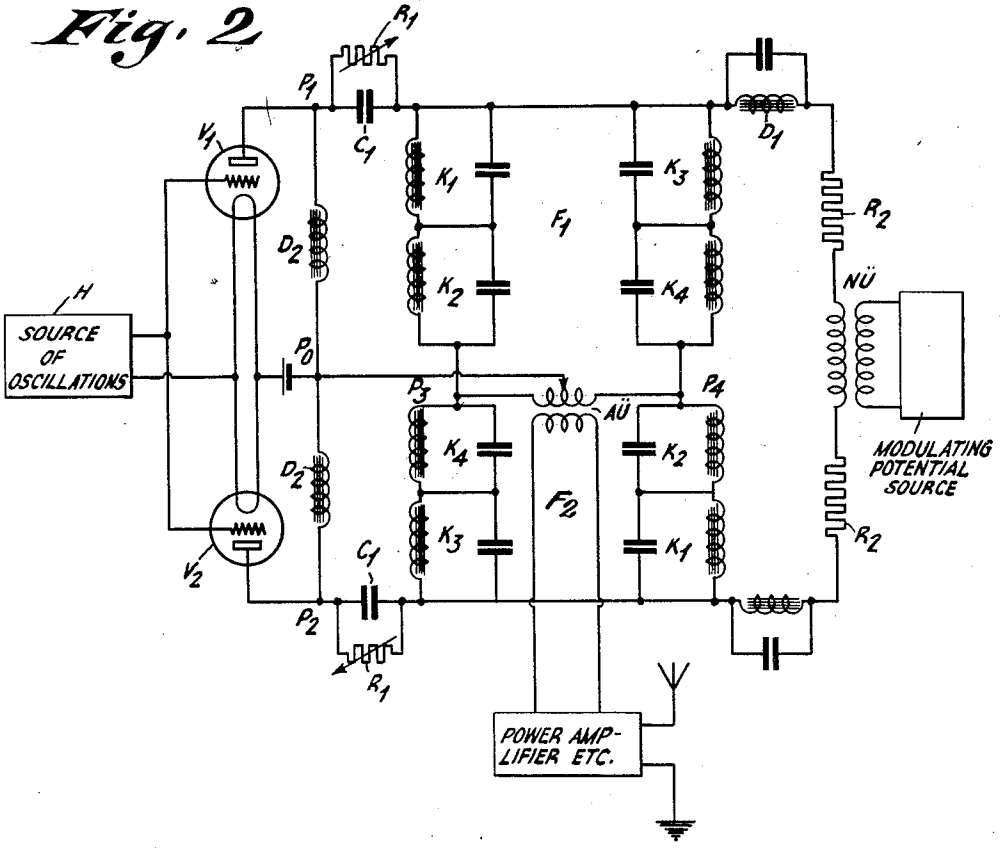



Fig. 2




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## MODULATION CIRCUIT ARRANGEMENT

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5 Claims. (Cl. 179—171)

The invention is concerned with a circuit scheme adapted to amplitude modulation. It is known in the art to superpose two waves of like frequency for the purpose of modulation, and to shift the phases of the two waves in reference to each other by a certain amount at the rhythm of the modulation wave or oscillation. What then results in the superposing circuit is an amplitude-modulated wave of the frequency of the superposed waves. By the ways and means known in the prior art for changing the phases of the two waves at the rhythm of the modulation frequency it is, however, impossible to insure linear modulation.

Now, according to the present invention, a perfectly straight modulation characteristic is insurable with the said modulation methods if, for the production of the requisite phase displacement, recourse is had to ways and means which will change the cosine of the phase angle roughly in direct proportion to the modulation current.

For further explanation of the basic idea of the invention the following considerations may be useful. The sum total of two alternating current voltages $e_1$ and $e_2$ whose form and development is expressible by the equations

$$e_1=e_0 \cdot \sin(\omega t-\phi) \quad (1)$$

$$e_1=e_0 \cdot \sin(\omega t+\phi) \quad (2)$$

is

$$E=e_1+e_2=2 \cdot e_0 \cdot \cos\phi \cdot \sin\omega t \quad (3)$$

The amplitude of the aggregate voltage, seeing that the amplitudes of the partial potential $e_0$ are by presupposition constant, is proportional to the cosine of the phase angle between the two partial or constituent potentials. Inasmuch as for the desired straight-line amplitude modulation the amplitude of the modulated voltage should change in proportion to the modulation current $i_m$ it follows that cos $\phi$ must be altered in proportion to the modulation current, in other words, there must be

$$\cos\phi=k \cdot i_m \quad (4)$$

or else

$$\phi=\text{arc cos}\ (k \cdot i_m) \quad (5)$$

$i_m$ which denotes the instantaneous modulation current is given by the equation

$$i_m=i_0 \cdot \sin nt \quad (6)$$

Equation (5) may be fulfilled, for instance, if the phase shift is accomplished by the aid of a band-pass filter. For small relative channel widths the phase measure of a univalent band-pass filter obeys this formula

$$a=\text{arc cos}\ \frac{2(\omega-\Omega)}{\Delta\Omega} \quad (7)$$

if $\omega$ is the operating frequency, $\Omega$ the mid-channel, and $\Delta\Omega$ the absolute channel width. The shift in the transmission range of the band-pass filter, i. e., the shift in the middle of the band $\Omega$, with constant value of the channel width $\Delta\Omega$ is realizable if the various elements of the band-pass filter are detuned in like sense. Such detuning is accomplishable by altering the inductances or capacities. A change in the inductances is obtainable by that the iron-cored inductances are subjected to a bias magnetization and by altering such bias magnetization at the rhythm of the modulation current and in accordance with the amplitude thereof. In view of the slight changes which are here dealt with, the existence of linear dependence of the change in inductance upon the change in magnetization may be assumed.

The presupposition shall be made that the relative detuning of the inductances is less than, or at the most equal to, 0.1, in other words, that there prevails a relation:

$$\lambda=\frac{\Delta L}{L}\leqq 0.1. \quad (8)$$

If, then, the mid-channel $\Omega$, in non-detuned state of the filter, coincides with the working frequency $\omega$, this equation holds good:

$$\frac{\Omega-\omega}{\omega}=\frac{\lambda}{2} \quad (9)$$

In other words, the relative shift of the middle of the channel is equal to one-half the relative change in inductance. From Equations 7 and 9 there follows for the phase "measure":

$$a=\text{arc cos}\cdot\left(-\lambda\frac{\omega}{\Delta\Omega}\right) \quad (10)$$

Inasmuch as the working frequency $\omega$ and the channel width $\Delta\Omega$ are constant, and, since, in the light of the presuppositions made, the relative detuning is proportional to the modulation current, Equation (10) may be expressed also in this form:

$$a=\text{arc cos}\ (K \cdot i_m) \quad (11)$$

In other words, the condition laid down initially by Equation (5) will be fulfilled, provided care is taken so that the phase measure $a$ coincides with sufficient accuracy with the phase angle $\phi$, a con-

dition insurable by that monovalent band-pass filters possessing as constant as feasible a surge impedance inside its transmission range are employed.

A fundamental circuit is shown in Figure 1 which illustrates a modulation circuit scheme according to this invention. The carrier frequency H is impressed upon the grid circuits of the two amplifier tubes $V_1$ and $V_2$ being paralleled at the input ends. United with the output circuits of the two tubes are the two filters $F_1$ and $F_2$, the latter being adapted to the internal resistance of the tubes, i. e., the surge impedance $z$ of the filters has been made equal to the internal tube resistance $R_i$. Presupposing perfect symmetrical construction, the said two amplifier tubes at their output ends result in two co-phasic electromotive forces of like amplitude. Between these electromotive forces and the voltages $E_2$, $E_2'$ and currents $J_2$, $J_2''$, at the output end of the filters there prevail the following relations:

$$E_0=(E_2+z\cdot J_2)\cdot(\cos\phi+j\cdot\sin\phi) \quad (12)$$

$$E_0=(E_2'+z\cdot J_2')\cdot(\cos\phi'+j\cdot\sin\phi') \quad (13)$$

If the phase displacement angles of the two filters, $\phi'$ and $\phi$, are made equal to each other, and if care be taken so that the two filters will cause a phase shift in contrary directions, then $\phi'$ may be put equal to $-\phi$. Because of the parallel connection at the output ends of the two filters there hold good the following relations:

$$E_2=E_2'=E_{II} \text{ and } J_2+J_2'=J_{II} \quad (14)$$

From Equations (12) to (14) there follows this relationship between the output potential and the electromotive force set up by the generators:

$$E_{II}+\frac{z}{2}\cdot J_{II}=\cos\phi\cdot E_0 \quad (15)$$

The external load resistance is suitably chosen

$$R_2=\frac{z}{2},$$

and, since $R_2\cdot J_{II}=E_{II}$, in the light of Equation (10), there is obtained

$$E_{II}=-\frac{\omega\lambda}{2\Delta\Omega}\cdot E_0 \quad (16)$$

The crest value for

$$E_{II}=\frac{E_0}{2}$$

is obtained for

$$\frac{\omega\cdot\lambda}{\Delta\Omega}=1 \text{ or } \frac{\omega}{\Delta\Omega}\cdot\lambda=1 \quad (17)$$

What flows from Formula (17) is that the complete modulation of the potential is realizable with so much smaller relative inductance changes $\lambda$, the lower the relative channel width

$$\frac{\Delta\Omega}{\omega}$$

of the band-pass filter.

The absolute channel-width $\Delta\Omega$ is governed by the maximum modulation frequency that shall be transmitted in a satisfactory manner and which is to be passed by the filter. In order that cut-offs in the transmitted frequency band by "initial oscillation actions" in the filter may be minimized, the transmission range of the band-pass filter should be chosen at least twice as large as the maximum modulation frequency. It will be noted that the circuit scheme becomes so much more sensitive, the higher the carrier frequency, for a given channel width. The channel width, or band-width, of the filter should be chosen as large and ample as feasible also for this reason that the damping inside the channel or band can in this instance be kept within limits by the use of simple means.

The modulation scheme here disclosed exhibits extremely great sensitiveness. For instance, in the case of a carrier wave of 1000 kilocycles and a filter band width of 30 kilocycles, perfect and complete modulation is attainable even by means of a change of inductance of only three percent from the normal inductance. By the use of suitable kinds of iron it is possible in this case to make the damping less than 0.1 Neper. One hundred percent rectilinear modulation will be feasible even if the phase shift in the marginal parts, due to coil losses and to imperfections in adaptation, does not fully correspond to the theoretical value, inasmuch as it is possible to insure rectilinear modulation and control down to the zero point, resultant voltage equal to 0, and since the scheme, if desired, need not be used under conditions where perfect phase coincidence between the partial voltages does not obtain. All that would happen as a result is that the efficiency is somewhat impaired.

Since the modulation scheme of the invention insures a filter action at the same time, it follows that undesirable harmonics are automatically cut off. Hence, the same may be used directly before the antenna circuit. The quality of modulation does not depend upon the degree of modulation chosen for the two tubes $V_1$ and $V_2$ of Figure 1. Indeed, these tubes may be modulated to the extent that is desirable from insuring energy transmission under conditions as economical as feasible, without an incidental reduction or impairment of modulation. Such harmonics of the carrier wave as may result from a further modulation of these tubes fail to affect the modulation process, in fact, they will be precluded readily from the output circuit by the band-pass filter.

The modulation, as here disclosed and being based upon the use of two monovalent band-pass filters whose range of transmission is shifted in contrary senses, is a pure amplitude modulation as clearly borne out by Equation (3). In cases where a certain phase modulation is not harmful, the circuit arrangement could be simplified by that, in lieu of the two monovalent band-pass filters, one bivalent filter is used whose output potential has superposed upon it the input voltage in a suitable manner. The bivalent filter, as is known in the art, for similar shifts of the transmission range, experiences twice as large a phase shift as a monovalent filter with the result that the phase angle of the shifted partial voltage, compared with the input voltage, will always agree with the phase angle of the two partial voltages in the instance above described. The aggregate resultant potential evidences also in this case a linear dependence upon the modulation current, though the phase angle of the aggregate voltage varies at the rhythm of the modulation frequency. Hence, phase modulation occurs conjointly with rectilinear amplitude modulation.

An exemplified embodiment of the basic idea of the invention is shown in Figure 2. The radio frequency wave H to be modulated is impressed upon the grid circuits of the two amplifier tubes $V_1$ and $V_2$ under conditions of phase equality. Hence, alternating voltages of the carrier frequency of like phase and like amplitude are set

up at the input ends of the two symmetric bridge filters $F_1$ and $F_2$. Each of said bridge filters comprises four paralleled oscillation circuits $K_1$—$K_4$. The output ends of both filters are combined so that superposition in the output transformer AU of the two alternating voltages fed to the two filters is insured. By the variable resistances $R_1$ part of the plate direct current of the two tubes flows in the same direction through the two filters. The resistances are by-passed or bridged by the condensers $C_1$ for radio frequency energy. The modulation frequency is fed by the transformer NU to the two filters in such a manner that the same will flow therethrough in opposite senses. By the stopper circuits $D_1$ which are tuned to the carrier wave, radio frequency energy is kept away from the source of the modulation frequency. Radio frequency choke-coils $D_2$ serve for the supply of the plate direct current for the amplifier tubes. The ohmic resistances $R_2$ connected in series with the secondary winding of transformer NU are intended to prevent a condition of frequency dependence in the energy transmission from the source of the modulation frequency to the modulation circuit.

Since the biasing magnetization occasioned by the direct current of the filters is reinforced in one of the filters by the modulation current, while being weakened in the other filter, it follows that the two filters will be detuned in contrary senses and thus cause a shift of the phases of the potential fed to the input ends thereof in opposite senses. The modulated radio frequency energy is collected across the terminals of the secondary winding of transformer AU.

The practical application of the basic idea of the present invention is not restricted to the example hereinbefore described. Indeed, the band-pass filters here suggested may be replaced by dissimilar filter arrangements or else the carrier frequency to be modulated may be fed by way of the transformer AU to both filters jointly, and the two phase-shifted voltages, after separate amplification, may be superposed. To be sure, in that case the amplifier tubes must be perfectly free from non-linear distortions, for these could subsequently not be eliminated any more, as is true, for instance, of the case illustrated in the circuit scheme shown in Figure 2.

Having thus described my invention and the operation thereof, what I claim is:

1. Signalling means comprising, a source of high frequency oscillations, a pair of thermionic tubes, each having input and output electrodes, a circuit for applying oscillations from said source to the control electrodes of said tubes in phase, a filter circuit comprising a plurality of paralleled oscillation circuits connected between the output electrodes of each of said tubes, a circuit for applying modulating potentials in phase opposition to said filter circuits, and a work circuit associated cophasally with said parallel oscillation circuits.

2. An arrangement as claimed in claim 1 in which the surge impedance of each of said filter circuits is constant over a predetermined frequency range.

3. Means for impressing amplitude modulations on high frequency oscillations, which modulations are linear with respect to the modulating potentials including, a pair of thermionic tubes having their control grids energized in phase by the said high frequency oscillations, each of said tubes including an anode and a cathode, a pair of filter circuits, one of said filter circuits having its input terminals connected between the anode and cathode of one of said tubes and the other of said filter circuits having its input terminals connected between the anode and cathode of the other of said tubes, the surge impedance of each of said filter circuits being substantially equal to the internal impedance of the tube to which it is connected, said filter circuits having their output terminals connected cophasally, the elements of the filter circuits being so selected that the filters cause phase shift in the carrier passed in opposite directions, a circuit for applying modulating potentials in phase opposition to said filter circuits, and a load circuit coupled with the output terminals of said filter circuits.

4. A device as recited in claim 3 in which the modulating potentials applied in phase opposition to said filter circuits detune said circuits alternately at the modulation current frequency, thereby altering the phase of the carrier wave in said circuits in opposite directions at the frequency of the modulating potentials.

5. A device as recited in claim 3 in which each of said filter circuits includes a plurality of parallel resonant circuits including inductances and capacities, and in which the modulating potentials applied in phase opposition to said filter circuits alter the tune thereof at the frequency of the modulating potentials.

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