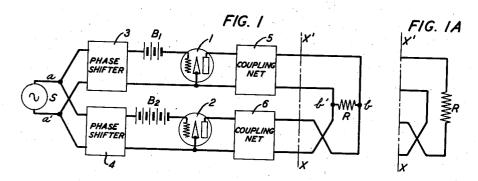
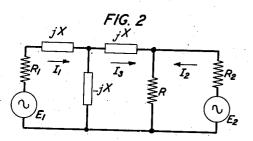
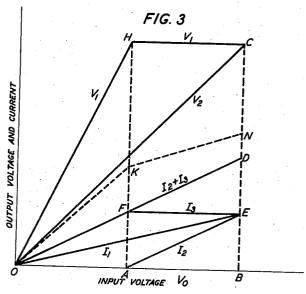
AMPLIFIER

Original Filed April 1, 1936

3 Sheets-Sheet 1







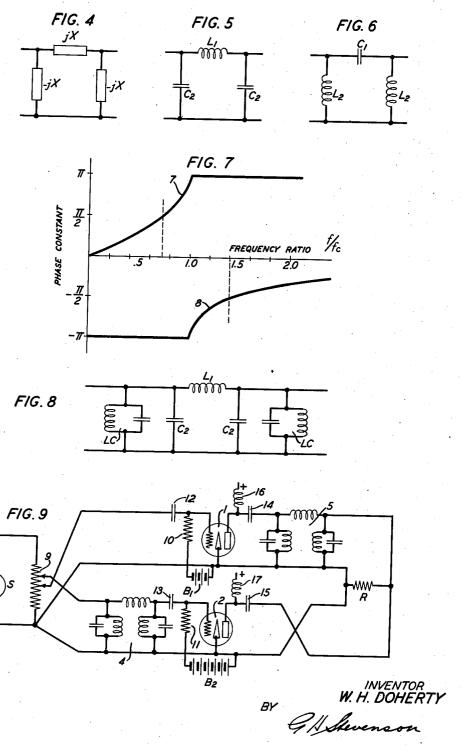
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3 Sheets-Sheet 2

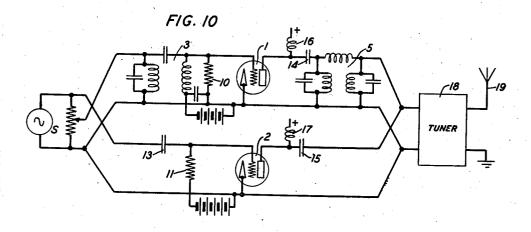


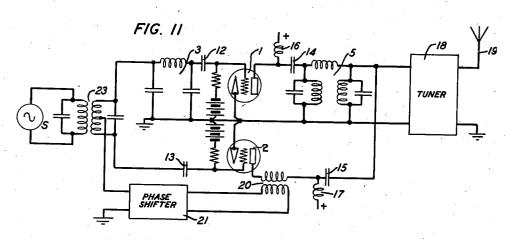
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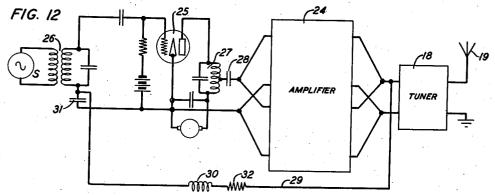
AMPLIFIER

Original Filed April 1, 1936

3 Sheets-Sheet 3







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UNITED STATES PATENT OFFICE

2,210,028

AMPLIFIER

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47 Claims. (Cl. 179-171)

This invention relates to vacuum tube amplifiers and more particularly to power amplifiers for signal modulated carrier waves.

The principal objects of the invention are to secure increased efficiency in the operation of power amplifiers and to provide for the maintenance of this increased efficiency under all conditions of modulation.

Power amplifiers for modulated carrier waves are usually class B or class C amplifiers, wherein the vacuum tubes are so biased that the flow of space current occurs in pulses of duration corresponding to one-half cycle or less of the impressed carrier oscillations. In amplifiers of this kind the plate circuit efficiency is substantially proportional to the ratio of the amplitude of the fundamental alternating output voltage to the value of the direct current plate voltage and is generally equal to about 85 or 90 per cent of this 20 ratio numerically. In practice, amplifiers may be operated satisfactorily at plate circuit efficiencies as high as from 65 to 75 per cent with constant high frequency input voltages.

When the input voltage is modulated the maximum efficiency, with the circuit arrangements heretofore used, is obtained only at the peak of the modulation cycle, the efficiencies at other instants in the cycle being less in proportion to the voltage amplitude. The average efficiency over a modulation cycle, for any given degree of modulation, is therefore always considerably less than the maximum efficiency obtainable under steady conditions, its value ranging from about 37 per cent for zero modulation to about 35 per cent for 100 per cent modulation.

By the present invention, the average efficiency may be increased to greater than 60 per cent for zero modulation and may be maintained at this high level for all degrees of modulation. Where amplifiers of large power are used, as in radio broadcast transmitters, the annual power saving represented by the increased average efficiency is very large, particularly since the average degree of modulation over any extended operating period is only about 20 per cent.

The increased efficiency is achieved in part by dividing the vacuum tubes of the amplifier into two groups, in one of which the tubes are continuously operative and in the other of which the tubes are operative only when the input voltage is greater than that of the unmodulated carrier. Since the tubes of the second group consume no power during one-half of a modulation cycle and reach their full power consumption only at the peak of the modulation, the net re-

sult is a reduction of the average power required.

Increased efficiency is obtained also by the use of coupling and driving circuits of novel character, the details of which are described hereinafter, whereby the tubes of the first group are enabled to supply their proper share of the output power at constant output voltage during the intervals when the tubes of the second group are also operative. This permits the tubes of the first group to be operated at their maximum 10 output voltage and therefore at their maximum efficiency for all input voltages in excess of the normal carrier voltage. The special circuit connections may be regarded as functioning to vary the effective load impedance into which the tubes 15 of the first group operate, the impedance diminishing as the impressed voltage increases so that the power output increases even though the output voltage, and therefore the efficiency, remains constant.

A characteristic feature of the invention is that the circuit coupling one of the groups of tubes to the load circuit is a circuit of the type having a phase constant of some odd integral multiple of

$\frac{\pi}{2}$ radians

Circuits of this type have certain properties which accomplish the functions described above. Among these are the following: A voltage at one terminal of the circuit is associated with a definite current at the other terminal irrespective of the terminating impedance; with a constant input voltage the output or load voltage is a linear function of the load impedance; and the input impedance of the circuit is inversely proportional to the load impedance. Because of this last characteristic, such circuits are sometimes referred to as impedance inverting networks.

The nature of the invention and its manner of operation will be more thoroughly comprehended from the following detailed description of a number of its embodiments and from the attached drawings of which:

Figs. 1 and 1a are schematics illustrating the general circuit configurations of the invention;

Figs. 2 and 3 are diagrams explanatory of the operation of the invention;

Figs. 4, 5 and 6 illustrate types of coupling 50 networks used in the invention;

Fig. 7 illustrates properties of the coupling networks:

Fig. 8 shows a modified form of coupling network:

Figs. 9, 10 and 11 are detailed schematics of two specific embodiments of the invention; and Fig. 12 discloses a modified form of the inven-

tion embodying feed-back stabilization.

Figs. 1 and 1a show in a generalized form the two fundamental species of the invention; Fig. 1, that in which the two amplifiers are connected in parallel to the load and Fig. 1a that in which they are connected in series. Figs. 9 to 11 show 10 specific embodiments of the species of Fig. 1. This parallel connected species which is illustrative of the fundamental principles of the invention will be described first.

The amplifying system, illustrated in Fig. 1. 15 comprises two parallel transmission paths interconnecting a source S of amplitude modulated carrier waves and a resistive load R, each path including a vacuum tube amplifier and its associated input and output networks. For the 20 sake of clarity the details of the networks and of the vacuum tube power supply circuits are omitted. The vacuum tube amplifiers are designated I and 2 and are shown as single tubes although parallel arrangements of tubes may be used. 25 Networks 3 and 4 are phase shifting networks for controlling the relative phases of the voltages impressed on the amplifiers. Networks 5 and 6 are tuned circuits for selectively transmitting the modulated carrier waves and are 30 preferably broad-band wave filters having transmission bands substantially wider than the frequency range occupied by the carrier and its attendant side-bands. Networks 3 and 4 should also, preferably, have relatively wide transmis-35 sion bands. B1 and B2 designate sources of negative grid bias voltages for the amplifiers.

Amplifier I is biased to about the plate current cut-off by the voltage of source B1. This amplifier is continuously operative for all amplitudes of the alternating voltage impressed on its grid. Amplifier 2 has a considerably greater negative bias on its grid, the bias voltage being such as to permit the space current to flow only when the voltage of the impressed oscillations 45 exceeds that of the unmodulated carrier.

Coupling network 5 in the output of the continuously operative amplifier is proportioned to have a phase constant which, in the general case,

is an odd integral multiple of

In practice it is preferable to have the phase constant of this network equal to

 $\frac{\pi}{2}$

since this requires a minimum of impedance ele-60 ments, but under special conditions, for example, where strong suppression of harmonics is desired, more complex networks giving greater discrimination may be used. Network 6 in the output of the intermittently operative amplifier is proportioned to have a phase constant equal to an even integral multiple of

70 at the carrier frequency. Preferably the phase constant of this network should differ from that of network 5 by

and when the latter network has a phase constant equal to

 $\frac{\pi}{2}$

the phase constant of network 6 may be zero. That is, the output of the amplifier may be connected directly to the load.

Phase shifting networks 3 and 4 are proportioned to produce phase changes of the impressed voltages complementary to those produced by coupling networks 5 and 6 to insure that the final output currents will combine in phase with each other in the load impedance. Generally these networks may be simple well-known phase splitting circuits for producing voltages in quadrature, for exciting the grids of the respective

The principles underlying the operation of the invention will be developed from a consideration 20 of the circuit of Fig. 2 which is representative of the output portion of the amplifier shown in Fig. 1 for the simplified case in which the network 5 has a phase constant

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The space path of amplifier I is represented by a resistance R1 with a wave source of voltage E1 connected in series and that of amplifier 2 by a corresponding resistance R2 and voltage E2. The voltages E_1 and E_2 differ in phase by the same amount as the respective input voltages differ from each other and are proportional to the input voltages in magnitude. Amplifier 2 is connected directly to the load resistance R, the coupling network 6 of Fig. 1 having been reduced to a direct connection. The circuit coupling amplifier 2 to the load network 5 of Fig. 1 is represented by a T-network of pure reactances, the shunt and series reactances being of opposite sign and all three having the same magnitude X. This network is an equivalent of any reactance network having a phase constant of

 $\overline{2}$

It is used here simply for theoretical purposes and is not intended to indicate a form that networks used in practice must have. Other structural forms are described later.

The equations for the mesh currents I1, I2 and I3 are the following:

$$R_1I_1+jXI_3=E_1 \ jXI_1+RI_3+RI_2=O \ RI_3+(R+R_2)I_2=E_2$$

From these are obtained

$$I_{1}=[E_{1}RR_{2}+jE_{2}XR]+\Delta I_{2}=[jE_{1}XR+E_{2}(R_{1}R+X^{2})]+\Delta I_{3}=-[jE_{1}X(R+R_{2})+E_{2}R_{1}R]+\Delta$$
 (2)

and

$$I_2+I_3=[E_2X^2-jE_1XR_2]\div\Delta$$

where

$$\Delta = R_1 R_2 R + X^2 (R + R_2)$$
 65

Current I1 is the output current of amplifier 1. I_2 is the output current of amplifier 2, and I_3 is the current delivered to the load from network

5. The total load current is the sum of I_2 and I_3 . 70 As expressed by Equations 2 each of these currents is made up of two components, one proportional to the electromotive force E1 and the other to the electromotive force E2.

If the voltages E1 and E2 were in like phase, 75

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each of the mesh currents and the load current would be made up of two quadrature components. By proper phasing of the voltages impressed on the amplifiers the relative phases of E1 and E2 may be made such that the two components of each current are in the same phase or are in phase opposition. In accordance with the invention the relative phases of the inputs are such that E_1 is in phase with jE_2 and hence that $-jE_1$ is in phase with E_2 . The two components of I1 defined by the first of Equations (2) are then of like sign and are in phase with E1, the components of the load current are also of like sign and are in phase with E_2 , while those of I_2 15 and I3 are of opposite sign. Uunder this condition, the output current of amplifier I is augmented when amplifier 2 is delivering current and the power output from amplifier I for a given input voltage is therefore increased.

20 In dealing with power amplifiers it is customary to describe the characteristics in terms of the output voltages of the vacuum tubes rather than the electromotive forces generated in the space paths. This results in a simplification of the equations for the mesh currents and in the case of the present invention serves also to bring out certain important operating characteristics.

The values of the mesh currents in terms of the voltages V₁ and V₂ at the output terminals of the vacuum tubes may be obtained from Equations (2) by making resistances R₁ and R₂ zero, the voltages E₁ and E₂ being then equal to the output voltages. Keeping in mind the condition stated above as to the relative phases of E₁ and E₂, the following magnitude relationships are obtained:

$$I_1 = \frac{V_2}{X}; I_2 = \frac{V_2}{R} - \frac{V_1}{X}$$
 (3)

$$I_3 = \frac{V_1}{X}$$
; $I_2 + I_3 = \frac{V_2}{R}$

From these equations may be obtained the further expressions for I_1 ,

$$I_1 = V_1 \frac{R}{\overline{X}^2} + I_2 \frac{R}{\overline{X}} \tag{4}$$

and

45

50

$$I_1 = V_1 - \frac{X^2}{R\left(1 + \frac{I_2}{I_2}\right)} \tag{5}$$

Equations (4) and (5) show how the current I₂ delivered from the intermittently operative amplifier 2 modifies the effective impedance into which the continuously operative amplifier I works, thereby permitting that amplifier to deliver an increased amount of power without a corresponding increase in its output voltage. Equation (5) also illustrates the impedance inverting characteristic of the quarter-phase network which couples amplifier I to the load.

It will be evident that so long as amplifier 2 is delivering no current the impedance into which amplifier I works is equal to

$$\frac{X^2}{D}$$

and is therefore inversely proportional to the actual load impedance. If the terminating impedance R is increased, the effective impedance presented to the tube of the amplifier I will diminish and vice versa. This impedance inverting property is a well-known characteristic of quarter wave-length lines and other reactance

networks operating at a frequency such as to make the phase constant equal to

$$\frac{\pi}{2}$$

radians or any odd multiple thereof. In multiple section networks such as wave filters the phase constant increases progressively with frequency in the transmission range and may pass through several odd multiples of

 $\frac{\pi}{2}$

at different frequencies. At each of such frequencies the impedance inverting property is found. At these frequencies also the output current is independent of the value of the load impedance so long as the voltage at the input terminals is held constant. This is indicated by the expression for I₃ in Equations (3) above.

The power delivered by amplifier i is equal to

$$\frac{V_1^2R}{X^2}$$

and hence is directly proportional to the load resistance. By keeping the output voltage of the amplifier constant it is, therefore, possible to increase the power output by increasing the resistance of the load; the opposite effect would take place if the amplifier were directly coupled to the load, or coupled to it through a line having a phase shift of any even multiple of

$$\frac{\pi}{2}$$

radians

When amplifier 2 comes into operation the apparent resistance of the load is equal to

$$R\left(1+\frac{I_2}{I_3}\right) 40$$

and consequently increases as the current I2 increases; i. e., as the amplifier 2 supplies a greater portion of the load current. Because of the impedance inverting property of the coupling network the effective load impedance presented to amplifier I is diminished as I2 increases and a larger power output is obtained from that amplifier without an increase in its output voltage. By proper adjustment of the amplifier input and biasing voltages, the rate at which the current I2 increases with the input voltage may be so regulated that the output voltage of amplifier I remains constant during the active half of the modulation period in which the amplifier 2 is operative. The additional power from amplifier I is thus obtained without reduction of its efficiency.

The use of the quarter-phase network in the output of amplifier I and the direct connection of the amplifier to the load whereby the above-noted results are obtained imposes a requirement that the input voltages of the two amplifiers shall be in phase quadrature in order that the two output currents shall combine additively in the load.

However, this exact circuit is only typical and as already pointed out, the impedance inverting network in the output of amplifier 1 may be a line or reactance network having a phase shift 70 of any odd multiple of

 $\frac{\pi}{2}$

radians and the coupling from amplifier 2 may be 75

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through a network having a phase shift of any even multiple of

 $\frac{\pi}{2}$

radians. In the general case, therefore, the networks 5 and 6 in the output circuit may have phase shifts differing by any odd multiple of

radians while the networks 3 and 4 in the input circuit have such complementary phase shifts as to make the total phase changes in the two amplifier paths equal or to bring the two load currents into phase. The essential feature is the use of the impedance inverting or quarter-phase network between one of the amplifiers only and the load and the compensation elsewhere of the phase displacement produced therein.

At the peak of the modulation cycle, when the total output power is a maximum, it is desirable that the load should be equally divided between the two amplifiers. This requires the proportioning of the values of the reactance X and the load resistance R with respect to each other as follows:

From Equation (3) the power outputs of the

amplifiers are found to be

$$V_1I_1 = \frac{V_1V_2}{X}$$

and

40 or

$$V_2 I_2 = \frac{V_2^2}{R} - \frac{V_1 V_2}{X} \tag{6}$$

35 At the peak values of the output voltages equality of V₁I₁ and V₂I₂ requires that

$$\frac{2V_1}{X} = \frac{V^2}{R}$$

$$\frac{2R}{2} = \frac{V_2}{R}$$
(7)

If the tubes of both amplifiers are of the same type, they will preferably be operated to deliver their maximum outputs at the same output voltages, in which case the required relationship between X and R becomes

$$X=2R$$
 (8)

In special cases it may be desirable to operate the two amplifiers at different maximum voltages or to share the load between them in other proportions. In each case an appropriate relationship of the reactance to the resistance may be found and the basic proportions of the coupling thereby established.

For linear amplification it is necessary that the voltage across the load impedance should vary in direct proportion to the voltage impressed 60 upon the system from the modulated wave source. In Fig. 1 the voltage from the input wave source is impressed on the system at terminals a a' and the load voltage is measured at terminals b b'. If these voltages are linearly related, the total 65 load current, being directly proportional to the load voltage, will also vary linearly with the input voltage. In the simplified system illustrated in Fig. 2, where the output terminals of amplifier 2 are connected directly to the load, the load voltage is the same as the amplifier output voltage V2. Due to the character of the coupling circuits and to the phase relationships of the impressed voltages the magnitudes of the several currents and voltages in the amplifier output cir-75 cuits are related in accordance with Equations

(3). To secure linearity of the output voltage characteristic further control of the magnitude variations of the input voltages on the amplifier grids is necessary. If it be assumed that this linearity of the output voltage and the total output current has been achieved, then the corresponding variations of the other voltages and currents in the output network are readily determined and from these the necessary characteristics of the input voltage control may be arrived at.

In Fig. 3 the required variations of the output current and voltage amplitudes are plotted as functions of the source voltage, denoted by V_0 . In the figure the values of V_0 are plotted as abscissae, OA representing the unmodulated carrier amplitude and OB, which is twice OA, representing the peak amplitude for 100 per cent modulation. Straight line OC represents the variation of voltage V_2 required for linear amplification, this voltage being also the voltage at the load terminals. The total output current I_2+I_3 is also linear and is represented by straight line OD.

Output current I₂ from amplifier 2 is zero for all input voltages less than OA and at higher voltages is represented by the line AE, rising to a value equal to half the load current at the modulation peak. Its variation from A to E may be assumed for the present to be linear. Current I₃ is equal to the difference between the total load current and I₂ and its required variation is therefore represented by broken line OFE. At input voltages greater than OA the amplitude of I₃ remains constant. Current I₁, which is proportional to V₂ is represented by straight line OE. 35 At the modulation peak it is equal to I₂ on the assumption that the two amplifiers have the same output voltage and power at this point.

The output voltage V_1 of amplifier I is directly proportional to I_3 and, since at the modulation peak it is assumed to be equal to V_2 its required variation follows broken line OHC. The maximum value is reached at the unmodulated carrier input and the voltage remains constant for larger inputs.

These curves show the voltage and current characteristics necessary for linearity of the output. The means for obtaining the required characteristics by control of the input voltage variations will be described later.

The efficiency of the amplifier when operating under the output voltage and current conditions defined by the curves of Fig. 3 may be determined approximately as follows:

It will be assumed that the tubes of both amplifiers are similar and are operated with equal D. C. plate voltages. It will also be assumed that the carrier wave is modulated by a sinusoidal signal oscillation producing a percentage modulation equal to 100m, where m denotes the ratio of the variation of the carrier oscillation amplitude to the amplitude of the unmodulated carrier. Over the modulation cycle the instantaneous value of the power delivered to the load varies in proportion to the square of the load current. 65 The average power depends upon the degree of modulation and has the value

$$W_{m} = \frac{I_{c}^{2}R}{2} \left(1 + \frac{m^{2}}{2}\right) \tag{9}$$

where W_m denotes the average output power and I_{c} the amplitude of the load current for zero modulation.

The plate current in amplifier I flows in a series of similar pulses, the amplitudes of which

vary directly with the amplitude of current I1. Its average value therefore remains constant for all degrees of modulation and the plate power has the average value

$$W_1 = E_b I_{b1} \tag{10}$$

where W1 denotes the plate power, Eb the D. C. plate voltage and Ib1 the average plate current for zero modulation. The plate current in amplifier 2 flows only during alternate half periods of the signal oscillation and occurs in the form of impulses which, due to the stronger grid bias, are of shorter duration than half cycles of the carrier current. The amplitudes of the impulses follow 15 the variations of the output current I2 but their duration is not constant and increases gradually with the amplitude. If the impulses were all of half wave duration the average value of the plate current would be equal to $2mI_{b1} \div \pi$, the 20 value being expressed in terms of the plate current in amplifier I since at the modulation peak the plate currents in both tubes would be equal. However, because of the shorter duration of the current pulses, amplifier 2 will operate at a slight-25 ly higher efficiency than amplifier i and the average value of the plate current will be somewhat less than that given above. To take account of this improvement in the efficiency the plate power consumed by amplifier 2 may be expressed by

$$W_2 = \frac{E_b I_{b1} 2mq}{\pi} \tag{11}$$

where W_2 is the plate power and q a variable factor which ranges from about 0.7 for zero modulation to 0.93 for full modulation.

30

60

From these expressions for the power output and power consumption, the average efficiency, denoted by η_m , throughout a modulation cycle is found to be

$$\eta_m = \eta_0 \frac{1 + \frac{m^2}{2}}{1 + \frac{2mq}{\pi}} \tag{12}$$

where η_0 is the efficiency for zero modulation. Since amplifier I alone is operative at zero modulation and is delivering power at its maximum voltage, its efficiency may be as high as 65 per cent or greater. As the degree of modulation increases the efficiency tends to diminish slightly in accordance with Equation 12 and at full modulation may be about 94 per cent of its value for zero modulation.

While the characteristics illustrated in Fig. 3 have been developed for the simplified circuit of Fig. 2 it may be shown readily that they also obtain in the general case where the network 5 has a phase constant equal to an odd integral multiple of

$$\frac{\pi}{2}$$

and network 6 has a phase constant differing therefrom by

$$\frac{\pi}{2}$$

The required form of the variation of voltage and current variations arises from the special phase relationships in the output networks regardless of the total phase angles and from the suppression of the output of the second amplifier at low input voltages. That this is so may be understood from the following consideration. If the two amplifiers were coupled directly to the load

or through networks producing phase changes that are even multiples of

$\frac{\pi}{2}$

the output voltages of both amplifiers would necessarily vary directly with the load voltage and hence, for linear amplification, would have to follow directly the variations of the voltage of the input source. Under such conditions it would not be possible to operate the first amplifier at constant output voltage and at the same time secure linear over-all amplification.

To obtain the required constancy of the output voltage V₁ for outputs greater than the normal carrier output it is necessary that the signal voltage impressed on the grid of the first amplifier should vary in an appropriate manner. For values up to the normal carrier amplitude the grid voltage should be directly proportional to the voltage Vo of the modulated wave source, but at larger amplitudes should increase more slowly than the wave source voltage. I have found that for most types of vacuum tubes the increase in the grid voltage between the normal carrier value and the value at the modulation peak should be about 30 per cent. The form of the variation of the impressed grid voltage is shown by the broken dotted line OKN in Fig. 3.

One method of securing the desired excitation 30 characteristic is to adjust the input circuit so that the grid of the first amplifier begins to draw current as soon as the input voltage amplitude exceeds the normal carrier value. When the grid becomes conductive, the amplifier input impedance is reduced and a diminution of the effective alternating grid voltage results. By the use of a grid leak, the direct component of the grid current may also be used to increase the negative bias on the grid thereby effecting a further reduction of the output voltage. Both effects may be utilized together, if desired, or each may be used separately. Further particulars of this method of excitation will be described later in connection with the subsequent figures.

The second amplifier is required to start delivering current when the source voltage exceeds the normal carrier amplitude and at the modulation peak, to deliver a maximum current equal to output current of the first amplifier. For this twofold adjustment of the output current, two independent variables of the input are available, namely, the grid biasing voltage and the amplitude of the voltage impressed on the grid from the source. By appropriate adjustment of these parameters the desired output characteristic may be obtained.

With regard to the structural form of the phase shifting and selective networks, many schematic forms including the T-arrangement shown in Fig. 2 may be used, but the equivalent schematic form shown in Fig. 4 is generally preferable in practice. This network, like that of Fig. 2, has a phase constant of

$\frac{\pi}{2}$

65

when the reactances are of the magnitudes and sign indicated. Changing the sign of each of the reactances changes the sign of the phase constant thereby producing a phase reversal of the output current and voltage. One specific form of network of the type shown in Fig. 4 is illustrated in Fig. 5, the structure of which comprises a series inductance L₁, and two equal shunt con-

densers of capacity C₂. This is a single section of a well-known form of low-pass filter. At the frequency corresponding to the resonance of inductance L₁ with capacity C₂ the phase constant has the value

 $\frac{\pi}{2}$

The values of L₁ and C₂ should therefore be chosen so that this resonance occurs at the operating frequency of the amplifier, that is at the frequency of the modulated carrier wave. Fig. 6 shows a network in which the reactances are of opposite sign from those of Fig. 5 and which, therefore, has a phase constant of minus

 $\frac{\pi}{2}$

when the reactances of the series and shunt branches are of equal magnitude.

The frequency variation of the phase constants of the networks of Figs. 5 and 6 are illustrated by the curves 7 and 8, respectively, of Fig. 7, which show their values as functions of the ratio of the variable frequency to the cut-off frequencies of the filters. For the low-pass structure the phase constant is equal to

 $\frac{\pi}{2}$

30 at a frequency equal to $.707f_c$, f_c being the cutoff frequency, and for the high-pass structure of Fig. 6 the phase constant has its desired value at a frequency equal to $1.414f_c$

a frequency equal to $1.414f_c$. As the frequency is varied through the transmission band the phase angles of the single section networks illustrated undergo changes of π radians or 180 degrees. If several sections are connected in tandem the phase constant at any frequency is increased in proportion to the num-40 ber of sections. Since it is characteristic of most single section filters, whether low-pass, highpass, or band-pass that the phase angle changes by 180 degrees between the band limits, it follows that, in the case of the band-pass filter the rate 45 of change of the phase shift will increase rapidly as the width of the transmission band is reduced. If very narrow band filters are used for the selective circuits 3, 4, 5 and 6 the resultant phase shifts may vary very rapidly with fre-50 quency in the neighborhod of the operating frequency with the result that the side-band frequencies corresponding to the modulating signal will not all be subject to the desired phase shift of 90 degrees. For this reason it is prefer-55 able to use networks having band widths as great as possible consistent with an adequate degree of suppression of the harmonics of the carrier wave. In practice band widths of the order of 60 kilocycles or greater have been found adequate for 60 dealing with speech modulated waves. At carrier frequencies of 500 kilocycles and upwards this permits a substantial suppression of harmonics

A form of band-pass network which has been found to be useful in practice is illustrated in Fig. 8 which corresponds to Fig. 5 with the addition of anti-resonant circuits LC in parallel with the shunt branches. If these anti-resonant circuits are tuned to the carrier frequency their effective reactance becomes infinite and the overall reactance relationship of the series and shunt branches at this frequency remains unchanged. The tuned circuits may be added at each end as illustrated or a single tuned circuit may be added at one end alone if desired. The network of

Fig. 6 may also be modified in the same way without changing the over-all reactance relationship. In networks of this type the operating frequency lies substantially midway between the cut-off frequencies. Other well-known forms of broad-band filter networks may also be used, the elements being appropriately proportioned so that the phase constant is equal to

 $\frac{\pi}{2}$

at the operating frequency. If desired also several sections may be connected in tandem thereby giving phase constants of any desired multiple of

 $\frac{\ddot{2}}{2}$

The complete amplifier schematic of Fig. 9 illustrates one system of phase shifting and selective networks and also circuit arrangements for providing the required input voltage characteristics. Network 5 in the output of amplifier is of the band-pass type shown in Fig. 8 and network 4 in the input of amplifier 2 is of the same type, the phase constants of the two networks being of the same sign and magnitude so that the over-all phase changes in the two paths are equal. At their input ends the two paths are connected to adjustable taps on potentiometer 9 in the output circuit of the modulated wave source S. Biasing potentials are supplied to the amplifier grids from sources B1 and B2 through resistors 10 and 11, blocking condensers 12 and 13 being provided to isolate the grids from other parts of the circuit. Plate current is supplied to the amplifier tubes through choke coils 16 and 17 with which are associated blocking condensers 14 and 15. In this circuit amplifier 2 is connected directly to the load and the phase difference in the two output paths is compensated by the network 4 in the input of amplifier 2. Since no phase shift is required in the input of amplifier 1, network 3 of Fig. 1 is omitted and a direct connection used.

The biasing voltage on the grid of amplifier I is adjusted to the plate current cut-off point and the input voltage from the wave source is adjusted so that at the normal carrier amplitude the grid is just becoming noticeably conductive. As the input voltage increases above this value the increased grid current flowing through resistor 19 produces an increased negative bias on the grid and a corresponding diminution in the amplification with a resultant tendency toward reduction of the output voltage. At the same time the impedance presented by the input circuit diminishes due to the increased conductivity of the grid and the amplitude of the oscillations reaching the grid is likewise diminished. Both effects tend in the same direction and by proper adjustment may be made to maintain the required output voltage characteristic. The proper output current characteristic of amplifier 2 is obtained by the coordinated adjustment of the voltage of biasing source B2 and of the amplitude of the input voltage by means of potentiometer 9 in the manner previously described.

In the modified form of the invention shown in Fig. 10 the phase shifting input network is inserted in front of amplifier ! instead of in the input of amplifier 2. Equalization of the total phase constants of the two paths is obtained by making the phase constants of the two networks equal in magnitude and opposite in sign. The output network 5 is of the same type as in Fig. 10

while the input network 3 corresponds to the type shown in Fig. 6 with the addition of an antiresonant circuit at one end. The biasing voltage is fed to the grid through the adjacent coil of the network instead of through the terminating resistor 10. The grid bias is thus substantially unaffected by any flow of grid current.

The output load is shown as an antenna 19 coupled to the amplifier through a tuning net10 work 18, the combination being proportioned and tuned so that it presents to the amplifier a pure resistive impedance of the desired magnitude at the operating frequency. Network 18, may be a narrow band wave filter or other well-known form of coupling circuit and may be as sharply tuned as desired.

Control of the output voltage depends wholly upon the change of the tube input impedance when the grid becomes conductive. The theory of the operation of this method of control is as follows:

The input circuit may be regarded as being constituted by a net work of the type shown in Fig. 4 terminated at its output end by a resistance R₂ which represents the resistor 10 in parallel with the grid to cathode space path, and supplied from a wave source of voltage E and effective internal resistance R₁.

At the operating frequency, for which the netse work has a phase constant equal to

$$\frac{\pi}{2}$$

the output voltage of the network denoted by E₂, may be shown to be given by

$$\frac{E}{E_2} = \left(\frac{R_1}{X}\right) + \left(\frac{X}{R_2}\right) \tag{13}$$

When no grid current flows the ratio of the two voltages is determined by the substantially fixed resistances R₁ and R₂ and the fixed resistance X. When a substantial grid current flows resistance R₂ is reduced to a new lower value and the ratio of the voltages is changed in such manner that the output voltage increases more slowly with increase of the source voltage. By proper selection of the value of resistor 10 the input voltage may be controlled to give the required output characteristic shown in Fig. 3.

The voltage at the network input terminals, which is also the voltage impressed on the grid of amplifier 2 is given by

$$\frac{E_1}{E} = \frac{X^2}{R_1 R_2 + X^2} \tag{14}$$

where E₁ denotes the input voltage. The reduction of R₂ due to the flow of grid current when the source voltage rises above the normal carrier value, results in a corresponding increase in the voltage impressed upon amplifier 2. This voltage increase assists in bringing amplifier 2 into action more rapidly and in many cases may be sufficient to permit the two amplifier paths to be operated at the same signal input voltage.

Another modification of the invention is shown in Fig. 11 in which the control of the input voltages is effected by regeneration. The two paths are coupled to the wave source S through a tuned transformer 23 and a feedback circuit is connected from the mid-point of the secondary of this transformer through a phase controlling network 21 to transformer 20 in the plate circuit of amplifier tube 2. The feedback circuit is adjusted so that the voltage fed back is in phase opposition to the input voltage of amplifier 1 and in

phase coincidence with that of amplifier 2 at the operating frequency. Since no current flows in the plate circuit of amplifier 2 so long as the input voltage is less than the normal carrier value, the feedback becomes effective only when the input voltage exceeds this value and then it operates to reduce the grid voltage on tube 1 and to increase that on tube 2. By proper adjustment of the amount of feedback the desired output voltage characteristics can be obtained. The primary winding of transformer 20 should preferably have a rather low inductance so that its presence in the plate circuit will not seriously affect the phase of the output currents.

In Fig. 11 the two networks 3 and 5 are both 15 of types giving positive phase shifts of

 $\frac{\pi}{2}$

the total phase change in the path being equal to π plus the phase change in the vacuum tube. To compensate this phase change the input voltage to the second path is reversed in phase by its connection to the secondary of tuned transformer 23 in the manner indicated in the drawing.

In the operation of the amplifiers of the invention it may be found that some slight deviation from the ideal linear amplification characteristic occurs at voltages near the normal carrier value, this being due to an improper fit of the characteristics of the two tubes and to lack of linearity of their individual characteristics. Such irregularities may be substantially eliminated by the use of degenerative feedback in the manner described in an article by H. S. Black, entitled Stabilized feedback amplifiers, Bell System Technical Journal, vol. XIII, January 1934. One circuit arrangement for applying the stabilizing feedback is illustrated schematically in Fig. 12. In this figure, the amplifier 24 which may correspond to any of the modifications shown in the previous figures derives its input voltage from a preceding amplifier 25 coupled to a suitable source through tuned coupling transformer 26. Coupling between amplifiers 24 and 25 is effected 45 through tuned circuit 21 and condenser 28. The feedback is applied through conductor 29, connected to the high voltage side of the amplifier output, resistance 32, and a phase controlling filter consisting of inductance 30 and condenser 31 to the input of amplifier 25. Since two amplification stages are used in this system a phase reversal of the feedback is necessary to bring it into the proper degenerative relation. This is effected by making the reactance of the coil 30 large relatively to that of condenser 31 at the operating frequency. The amount of the feedback may be controlled by varying the inductance of the coil.

The degenerative feedback also makes it possible to simplify the input circuit arrangements, since any lack of linearity due to improper relationship of the input voltage magnitudes in the two paths will be suppressed thereby. It becomes possible therefore to apply the same input voltage to each path as shown in the figure.

In Fig. 1—a, which represents a modification of the circuit of Fig. 1 to the right of line XX', there is shown an alternative connection of the load impedance whereby it is supplied with current from the two amplifiers in series instead of in parallel. The same operating characteristics are obtained and the same requirements on the input voltages apply as in the circuit of Fig. 1. Due to the series relationship of the load, how-75

ever, it is necessary that the coupling networks 3, 4, 5 and 6 have their phase characteristics interchanged. Amplifier 2, which is intermittently excited, must be coupled to the load through a network having a phase constant equal to

 $\frac{\pi}{2}$

or an odd multiple thereof while the output cir-10 cuit of amplifier I should have a phase constant of zero or an even multiple of

 $\frac{\pi}{2}$

15 It is obvious that in this modification amplifier 2 cannot be directly coupled to the load, since then the load would receive current only during the intervals that this amplifier is excited. By the modification of the phase constants of the coupling networks as indicated above the load can receive power at all instants.

What is claimed is:

1. An amplifying system comprising a source of modulated carrier waves, a load circuit, two 25 parallel transmission paths connecting said source and said load, an amplifying device included in each of said paths, phase controlling means between said devices and said source for producing a phase quadrature relationship be-30 tween the waves impressed upon the input terminals of said devices from said source, and phase controlling means between the output terminals of said devices and said load for producing a compensating relative phase shift of the 35 amplified oscillations whereby they combine in additive phase in said load circuit and the impedance presented to the output of one of said amplifying devices varies inversely with the effective impedance of the load circuit.

2. An amplifying system for supplying amplitude modulated carrier waves comprising a wave source, a load circuit, two parallel transmission paths connecting said source and said load, an amplifying device included in each of said paths, 45 phase shifting means between at least one of said devices and said source for producing a phase quadrature relationship between the waves impressed upon the input terminals of said devices from said source, phase shifting means between 50 at least one of said devices and said load for producing a compensating relative phase shift of the amplified oscillations whereby they combine in like phase in said load and whereby the impedance presented to the output of the last-55 mentioned one of said devices varies inversely with the terminating impedance of the last-mentioned phase shifting means, and biasing means for rendering one of said amplifying devices inoperative for input amplitudes less than a pre-60 assigned value.

3. An amplifying system for supplying amplitude modulated carrier waves comprising two amplifying devices, a wave source, a load circuit, separate input paths coupling the input terminals of said devices to said source, separate output paths coupling the output terminals of said devices to said load circuit, a reactance network in one of said output paths producing a phase shift of the amplified waves equal to 90 degrees at the frequency of said source, the phase shift in the second of said paths differing therefrom by 90 degrees whereby the impedance across the output terminals of that one of said amplifying devices connected to said one of said output paths varies inversely with the effective terminating

impedance of said reactance network, and phase controlling means in said input paths producing compensating phase displacements of the input waves whereby the output waves from the two amplifying devices combine in like phase in said load impedance.

4. An amplifying system comprising two amplifying devices, a pair of common input terminals, a pair of common output terminals, separate input paths connecting said devices to said common input terminals, separate output paths coupling said devices to said common output terminals, a reactance network in one of said output paths having a phase constant equal to an odd integral multiple of

 $\frac{\pi}{2}$

radians at a preassigned operating frequency, coupling means in the other output path having a phase constant at said frequency equal to an even integral multiple of

 $\frac{\pi}{2}$

radians including zero, and coupling means in said input paths having phase constants of such magnitudes at the prearranged frequency as to equalize the over-all phase constants of the two paths between said common input terminals and said common output terminals.

5. An amplifying system comprising two amplifying devices, a pair of common input terminals, a pair of common output terminals, separate input paths coupling said devices to said input terminals, separate output paths coupling said devices to said output terminals, a reactance network in one of said output paths having a phase constant equal to an odd integral multiple of

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65

radians at a preassigned operating frequency, coupling means in the other of said output paths having a phase constant at said frequency equal to an even integral multiple, including zero, of

 $\frac{\pi}{2}$

radians, coupling means in said input paths having phase constants of such magnitudes at said preassigned frequency as to equalize the overall phase constants of the two paths between said common pairs of terminals, and biasing means for rendering one of said amplifying devices inoperative at input amplitudes less than a preassigned value

6. An amplifying system comprising two amplifying devices, a pair of common input terminals, a pair of common output terminals, separate input paths coupling said devices to said input terminals, separate output paths coupling said devices to said output terminals, a reactance network in one of said output paths having a phase constant equal to

 $\frac{\pi}{2}$

radians at a preassigned carrier frequency, the other of said output paths having a phase constant equal to zero at said frequency, coupling means in said input paths having phase constants of such magnitudes as to equalize the overall phase constants of the two paths between said common pairs of terminals, and biasing means 75

for rendering one of said amplifying devices inoperative at input amplitudes less than a pre-

assigned value.

7. An amplifying system for carrier waves of variable amplitude comprising two space discharge amplifiers, a load impedance, a wave source, a reactance network coupling one of said amplifiers to said load, said network having a phase constant substantially equal to

 $\frac{\pi}{2}$

radians at a preassigned operating frequency, a separate coupling having substantially zero phase 15 constant between the second of said amplifiers and said load, input circuits coupling said amplifiers and said source, means in said input circuits for producing a quadrature phase relationship of the waves impressed upon said amplifiers, 20 and biasing means for rendering said second amplifier inoperative at input amplitudes less than a preassigned value.

8. An amplifying system for carrier waves of variable amplitude comprising a space discharge 25 amplifier, a load impedance, a source of carrier waves, an impedance inverting reactance network coupling said amplifier and said load, an input circuit coupling said amplifier and said source, an additional transmission path between 30 said source and said load, said additional path and the path constituted by said amplifier and its associated coupling circuits having equal phase constants, and biasing means for blocking said additional path to waves of less than a preassigned amplitude.

9. An amplifying system for supplying amplitude modulated carrier waves comprising a space discharge amplifier, a load impedance, a source of carrier waves of preassigned frequency and preassigned normal amplitude, a reactance net-

work having a phase constant of

45 radians at said preassigned frequency coupling said amplifier and said load, an input circuit coupling said amplifier and said source, an additional transmission path between said source and said load, said additional path and the path constituted by said amplifier and its associated coupling circuits having equal phase constants, and biasing means for blocking said additional path to waves of less than the said normal carrier amplitude.

10. In an amplifying system a source of amplitude modulated carrier waves, a pair of amplifying paths, a load circuit, an impedance inverting network connected between the output 60 of a first of said amplifying paths and said load circuit, direct connections from the output of the second of said amplifying paths to said load circuit, means for so exciting said amplifying paths from said wave source that the outputs 65 therefrom are in phase in said load circuit, means for blocking said second amplifying path to waves of less than a predetermined normal amplitude, and means for so controlling the operation of said first amplifying path that its out-70 put voltage is substantially constant for input

11. An amplifying system for carrier waves of variable amplitude comprising a space discharge amplifier, a load impedance, a source of carrier 75 waves of preassigned frequency and preassigned

voltages above said normal value.

normal voltage, a reactance network having a phase constant of

 $\frac{\pi}{2}$

radians at said preassigned frequency coupling said amplifier and said load, an input circuit coupling said amplifier and said source, an additional transmission path between said source and said load, a second space discharge amplifier included 10 in said path and coupled to said load with zero phase shift, the two paths between said source and said load having substantially equal phase constants, biasing means for blocking said second amplifier to waves of less than said normal 15 voltage, and means for automatically controlling the amplitude of the waves impressed on said first amplifier in response to excesses of said wave source voltage above said normal value whereby the output voltage of said first amplifier is 20 maintained constant.

12. An amplifying system in accordance with claim 11 in which the first-mentioned space discharge amplifier is biased substantially to the

plate current cut-off point.

13. An amplifying system in accordance with claim 11 in which the said reactance network is proportioned with respect to the value of said load impedance to provide an equal division of the total output power between the said two amplifiers at a voltage of said wave source equal to twice the said normal value.

14. An amplifying system in accordance with claim 11 in which the said amplifiers are gridcontrolled vacuum tubes, the grid of the firstmentioned amplifier being negatively biased to substantially the plate current cut-off point, and the said normal wave source voltage being just sufficient to produce a positive grid current in said first-mentioned amplifier.

15. An amplifying system in accordance with claim 11 in which the said automatic controlling means comprises a coupling for producing from the output circuit of said second amplifier a feedback voltage and supplying said feedback voltage to the input circuit of said first amplifier in opposition to the input voltage supplied thereto from said source.

16. An amplifying system in accordance with claim 8 and in combination therewith degenerative feedback coupling means between said load and said source.

17. In an amplifying system comprising a source of amplitude modulated carrier waves, a load circuit, and a pair of amplifying devices included in separate transmission paths between said source and said load, an impedance inverting network coupling the output terminals of a first of said devices to said load, direct connections from the output of the second of said devices to said load, coupling circuits between said devices and said source having phase shift characteristics such that the output currents from said devices are in phase in said load circuit, means for blocking said second device to waves from said source of less than a predetermined normal amplitude, and means operative when the voltage from said source exceeds said normal amplitude to control the transmission between said source and said first device whereby its 70 output voltage is maintained constant.

18. In an amplifying system comprising a source of amplitude modulated waves of preassigned normal carrier voltage, an amplifying device, and a load circuit, an impedance inverting 75

network coupling the output terminals of said device and said load circuit, an input circuit coupling said device and said source, means responsive to excesses of the voltage of said source above said normal value for reducing the effective impedance at the output terminals of said device substantially in proportion to the magnitude of the excess voltage, and means simultaneously operative to control the transmission efficiency of said input circuit whereby the output voltage of said device is maintained constant.

19. In an amplifying system comprising a source of amplitude modulated waves of preassigned normal carrier voltage, an amplifying device, an input circuit coupling said device to said source, and an output circuit connected to the output terminals of said device, means responsive to voltages of said source above said normal value for reducing the effective impedance of said output circuit and means simultaneously operative to control the transmission efficiency of said input circuit whereby the output voltage of said device is maintained constant.

20. In an amplifying system, a source of elec-25 trical oscillations of varying amplitude, a load circuit, a first amplifier, circuits for connecting the input of said first amplifier to said source and the output thereof to said load circuit so that for oscillations from said source of amplitude 30 below a predetermined level the output voltage of said amplifier is substantially linearly proportional to the amplitude of said oscillations and for oscillations from said source of a range of amplitudes above said predetermined level 55 the output voltage of said amplifier is substantially constant, a second amplifier, and circuits for connecting the input of said second amplifier to said source and the output thereof to said load so that the output current from said second am-40 plifier is substantially zero for oscillations below said predetermined level and substantially proportional to the amplitude of said oscillations for oscillations of said range of amplitudes above said predetermined level and so that the currents 45 supplied by said amplifiers to said load are in phase.

21. A system according to claim 20 in which at least one of said connecting circuits includes at least one network of reactive elements so artanged and proportioned as to have a phase constant of some odd integral multiple of

$\frac{\pi}{2}$ radians

55 22. In an amplifying system, a source of electrical oscillations of varying amplitude, a load circuit, two amplifiers interconnecting said source and said load, means for so connecting and operating one of said amplifiers that for oscillations from said source below a predetermined amplitude the current supplied thereby to said load circuit is substantially linearly proportional to the amplitude of said oscillations and for a range 65 of amplitudes above said predetermined amplitude the current supplied thereby to said load circuit is substantially constant, and means for so connecting and operating the other of said amplifiers that for oscillations from said source 70 of amplitude below said predetermined amplitude substantially no current is supplied to said load circuit and for a range of amplitudes above said predetermined level the current supplied thereby to said load is substantially linearly pro-75 portional to the amplitude of the oscillations

from said source and in phase with the load current supplied by said first amplifier.

23. A system in accordance with claim 22 in which said means include means for causing the load currents supplied to said load by said 5 amplifiers to be equal for oscillations from said source of amplitude substantially twice said predetermined amplitude.

24. In an amplifying system, a source of amplitude modulated carrier oscillations, a load 10 circuit, a first vacuum tube amplifier, an input circuit for said first amplifier connected to said source, an output circuit for said first amplifier connected to said load circuit, said output circuit comprising reactance elements of such 15 character and so arranged and proportioned that the impedance into which said amplifier works is inversely proportional to the impedance of said load circuit and of a value such that said amplifier operates at a high efficiency at the 20 carrier amplitude of oscillations from said source, a second vacuum tube amplifier, an input circuit for said second amplifier connected to said source, an output circuit for said second amplifier connected to said load circuit, said output 25 circuit being of such a character that the impedance into which said second amplifier works is directly proportional to the load impedance, and means in said input circuits for said first and second amplifiers for causing said second 30 amplifier to be operative only for oscillations from said source greater than the carrier amplitude and for causing the currents supplied by said amplifiers to said load for higher amplitudes to be in phase and substantially equal for ampli- 35 tudes of oscillations from said source twice the carrier amplitude.

25. A system in accordance with claim 24 in which said means comprises a phase shifting network.

26. A system in accordance with claim 24 in which said means comprises means for biasing the input of said second amplifier so that no plate current flows for input oscillations below the carrier amplitude.

27. A system in accordance with claim 24 in which said means comprises a phase shifting network in the input circuit for said first amplifier.

28. A system according to claim 24 in which 50 said means comprises a phase shifting network in the input circuit of said first amplifier, an input bias for said first amplifier and means for causing the input bias of said first amplifier to increase for input oscillations above the carrier 55 amplified.

29. In an amplifying system, a source of amplitude modulated carrier waves, a load circuit, a continuously operating amplifier, means for connecting the input of said amplifier to said 60 source so that for voltages from said source below the carrier amplitude the voltage output of said amplifier is substantially proportional to the input voltage and for higher voltages from said source the output voltage of said amplifier is 65 substantially constant, a circuit for connecting the output of said amplifier to said load comprising reactance elements so arranged and proportioned that the output current thereof is independent of the terminating impedance, a sec- 70 ond amplifier, means for connecting said source to said second amplifier so that for voltages from said source below the carrier amplitude no output is produced in said amplifier and that for voltages from said source above the carrier am- 75

plitude the output voltage of said amplifier varies substantially proportionately with the input voltage, and means for connecting the output of said second amplifier to said load so that the current supplied thereto from the second amplifier is substantially directly proportional to the output voltage of said second amplifier, said means for connecting the inputs of said continuously operating amplifier and said second amplifier to said source having such phase constants that the load currents supplied by said amplifier to said load are in phase.

30. In an amplifying system, a source of amplitude modulated carrier oscillations, a first amplifier having a grid, a cathode and anode, a load circuit, a circuit comprising reactance elements so arranged and proportioned that the phase constant of said circuit is an odd integral

multiple of

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 $\frac{\pi}{2}$

radians and having input terminals connected to the anode-cathode circuit of said amplifier 25 and output terminals connected to said load circuit, a second amplifier having a grid, a cathode and an anode, a circuit connecting the anodecathode circuit of said second amplifier to said load circuit and having a phase constant of an 30 even integral multiple of

 $\frac{\pi}{2}$

radians including zero, and means connecting the grid-cathode circuits of said amplifiers to said source so that the anode-cathode output voltage of said first amplifier is directly proportional to the voltage of the oscillations from said source for amplitudes below the carrier level and substantially constant for voltages from said source above the carrier level and so that said second amplifier produces substantially no output for voltages from said source below the carrier level and that for voltages from said source above the carrier level the output of said second amplifier is substantially linearly proportional to said voltages and the current supplied to said load circuit by said amplifiers are in phase.

31. In combination, a load circuit and an amplifying system for supplying modulated waves thereto comprising a vacuum tube having an input circuit and an output circuit, a quarter wave network connecting said output circuit to said load circuit, a second vacuum tube, and means for so connecting said second vacuum tube to said load circuit that for outputs above a predetermined value only, current will be supplied thereby to said load circuit in phase with the current supplied by said first vacuum tube.

32. An amplifying system for supplying signal modulated carrier waves comprising two space discharge devices each having an input circuit and an output circuit, an output network comprising a substantially non-dissipative quarter 65 wave-length passive transducer and a work circuit coupled to said transducer at one end thereof, circuits coupling the output circuits of said discharge devices to the opposite ends of said network respectively, and wave source means in 70 the input circuits of said devices impressing oscillatory voltages thereon, the voltage in said input circuits being so related as to produce by their conjoint action high frequency oscillations varying in amplitude in accordance with the sig-75 nal at the output terminals of one of said devices and at the terminals of said work circuit, and throughout at least a portion of the signal cycle, a substantially constant high frequency voltage at the terminals of the other of said devices.

33. An amplifying system for supplying high frequency signal modulated carrier waves comprising two vacuum tubes each having an anode, a cathode and a control grid, an output network comprising a substantially non-dissipative quarter 10 wave-length passive transducer and a work circuit constituting a load impedance coupled to said transducer at one end thereof, circuits coupling the anode space paths of said tubes to the opposite ends of said network respectively, input cir- 15 cuits connected respectively to the grids of said tubes, and wave source means included in said input circuit for impressing oscillatory voltages on said grids to produce in said work circuit high frequency oscillations modulated in amplitude in 20 accordance with a signal, the voltages impressed on said grids being so related as to produce by their conjoint action a substantially constant high frequency voltage across the anode space path of one of said tubes in the presence of 25 varying currents in the anode space path of the other of said tubes.

34. An amplifying system for supplying high frequency waves modulated in amplitude in accordance with a signal comprising two space dis- 30 charge devices, each having an anode, a cathode and a control grid, an output network comprising a substantially non-dissipative quarter wavelength passive transducer and a work circuit constituting a load impedance coupled to said transducer at one end thereof, circuits coupling the anode space paths of said discharge devices to the opposite ends of said network respectively, input circuits connected respectively to the grids of said discharge devices, and wave source means 40 included in said input circuits impressing oscillatory voltages on said grids whereby high frequency current pulsations of amplitude varying with the signals are produced in the anode space paths of said discharge devices, the voltages im- 45 pressed on said grids producing by their conjoint action, voltages of signal modulated amplitudes at the output terminals of one of said devices and at the terminals of said work circuit and throughout at least a portion of the signal 50 cycle, a substantially constant output voltage at the output terminals of the other of said dis-

charge devices. 35. In an amplifying system for supplying high frequency waves modulated in amplitude in ac- 55 cordance with a signal, a space discharge device having input and output circuits, means including a high frequency wave source in the input circuit of said device for producing in the output circuit thereof carrier frequency current 60 impulses varying in amplitude in accordance with the signal, a second space discharge device having input and output circuits, means in the input circuit of said second device for impressing oscillatory voltages thereon and producing 65 in the output circuit thereof high frequency current impulses in phase quadrature with the current impulses of said first device, a work circuit, and circuits coupling the output circuits of said devices to said work circuit, one of said coupling 70 circuits including a quarter wave-length reactive network whereby the output current of said devices are brought into phase in said work circuit, the voltages impressed on said space discharge devices being so related in their magnitudes and 75

variations as to produce by their conjoint action voltages of signal modulated amplitudes at the output of one of said devices and at the terminals of said work circuit and, at least during a portion of the signal cycle, a substantially constant voltage at the output of the other of said devices.

36. An amplifying system comprising in combination a load circuit, a space discharge device 10 having an input circuit and an output circuit, a quarter wave-length network coupling said output circuit to said load circuit, a second space discharge device having an input circuit and an output circuit, a circuit coupling the output 15 circuit of said second device to said load circuit, and means impressing oscillatory voltages on the input circuits of said devices whereby high frequency currents of signal dependent amplitudes are produced in the output circuits thereof, the 20 voltages impressed on said input circuits being so related that the currents from the said output circuits combine in phase with each other in said load circuit and that the high frequency current in the output of one of said devices is 25 zero in the absence of signal variations.

37. An amplifying system in accordance with claim 36 in which the voltages in said input circuits are so related as to produce a voltage of signal modulated amplitude at the output ter-30 minals of one of said devices, and, during at least a portion of the signal cycle, a voltage of substantially constant amplitude at the output terminals of the other of said devices.

38. A system for supplying amplitude modu-35 lated waves comprising a plurality of vacuum tubes, a load circuit, means causing one of said tubes to operate with a variable voltage at its output terminals, and means cooperating with said one tube for causing another of said tubes 40 to operate during at least a portion of the modulation cycle at a substantially constant output voltage while supplying varying current energy.

39. A system for supplying amplitude modulated waves comprising a plurality of vacuum 45 tubes, a load circuit, means causing one of said tubes to operate with an oscillating voltage of variable amplitude at its output terminals, and means including a quarter wave-length network cooperating with said one tube for causing an-50 other of said tubes to operate during at least a portion of the modulation cycle at a substantially constant oscillatory output voltage while supplying varying current oscillatory energy.

40. The method of operating a multiple tube 55 amplifier at high efficiency for supplying amplitude modulated oscillations to a load circuit which comprises supplying the oscillations from the plurality of vacuum tubes, operating one of said tubes to supply varying energy at a sub-60 stantially constant output voltage during at least a portion of the modulation cycle, and operating a second tube at a varying output volt-

41. The method of operating at high efficiency 65 an amplifier comprising a plurality of vacuum tubes for supplying signal modulated oscillations to a load, which comprises operating one vacuum tube at a constant oscillatory output voltage and varying oscillatory output current during at least 70 a portion of the modulation cycle, converting said variable current-constant voltage oscillations into constant current-variable voltage oscillations, supplying such converted oscillations to the load, operating a second vacuum tube at 75 an oscillatory output voltage of varying amplitude, and combining the outputs of said tubes. 42. In an amplifying system for supplying amplitude modulated carrier waves, an output network comprising a quarter wave-length transducer and a work circuit connected at one end 5 of said transducer, space discharge means supplying constant amplitude high frequency oscillations to one end of said output network during at least a portion of the modulation cycle, and other space discharge means supplying to 10 the other end of said network high frequency oscillations of signal dependent ampitudes and in quadrature phase relation to the oscillations supplied to said network from the first discharge means.

43. In an amplifying system, a work circuit, space discharge means coupled to said work circuit and supplying thereto high frequency oscillations of constant amplitude during at least a portion of the cycle of operation, a second 20 space discharge device separately coupled to said work circuit and supplying thereto high frequency oscillations of signal dependent amplitudes, a quarter wave-length network included in the coupling between one of said devices and 25 said work circuit and means controlling the relative phases of the oscillations from said devices whereby they combine in phase in said work circuit.

44. In an amplifier for supplying to a load cir- 30 cuit high frequency oscillatory energy of amplitudes varying in accordance with a signal, a vacuum tube having an output circuit, a passive transducer having two sets of terminals and of such characteristics that the impedance measured at one set of terminals is the reciprocal of the impedance of a circuit connected to the other set of terminals, a circuit for supplying power from said output circuit to the load circuit and including connections from said output circuit to 40 one of said sets of terminals, a second vacuum tube having an output circuit connected to the other of said sets of terminals, and means for so exciting said second vacuum tube that the impedance of its output circuit varies in accordance 45 with the signal at least throughout a portion of the signal cycle, and the current in said output circuit is in phase quadrature with the current in the output circuit of the first vacuum tube.

45. An amplifying system for supplying high 50 frequency waves modulated in amplitude in accordance with a signal, comprising two space discharge devices each having an anode, a cathode and a control grid, a substantially non-dissipative quarter wave-length passive transducer hav- 55 ing one terminal connected to the cathode-anode circuit of one of said devices, a work circuit constituting a load impedance connected to the other terminal of said transducer in parallel with the anode-cathode circuit of the other of said 60 devices, input circuits connected respectively to the grids of said devices and wave source means included in said input circuit impressing oscillatory voltages on said grids whereby high frequency current pulsations of amplitudes varying 65 with the signal are produced in the anode space paths of said devices, the voltages impressed on said grids producing by their conjoint action voltages of signal modulated amplitudes at the terminals of said work circuit and during at least 70 a portion of the signal cycle, a substantially constant output voltage across the anode space path of said one of said devices.

46. An amplifying system for supplying high frequency waves modulated in amplitude in ac- 75

cordance with a signal comprising two space discharge devices each having an anode, a cathode and a control grid, a work circuit, a substantially non-dissipative quarter wave-length passive 5 transducer having one set of terminals connected to the anode-cathode circuit of one of said discharge devices and a second set of terminals connected to said work circuit and the anodecathode circuit of the other of said devices in 10 parallel, input circuits connected respectively to the grids of said discharge devices, and wave source means included in said input circuits for impressing oscillatory voltages on said grids whereby high frequency current pulsations of 15 amplitudes varying with the signal are produced in the anode space paths of said discharge devices, the voltages impressed on said grids producing by their conjoint action a substantially constant voltage across the anode space path of 20 said one of said discharge devices in the presence of varying current in the anode space path of said other of said discharge devices.

47. An amplifying system comprising two space discharge devices having output circuits and input circuits, a load circuit, circuits respectively 5 coupling the output circuits of said amplifiers to said load circuit, impedance inverting network means included in one of said coupling circuits, wave source means impressing oscillations on the input circuits of said space discharge devices to 10 produce in the output circuits thereof and in said load circuit high frequency currents varying in amplitude in accordance with a signal, and means controlling the relative amplitudes of the impressed oscillations whereby the output 15 voltage of one of said discharge devices is maintained substantially constant in the presence of currents of varying amplitude in the output circuit of the other of said devices.

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