Aug. 7, 1945. T. H. PRICE 2,381,181

RADIO TRANSMITTER

Filed April 5, 1943 4 Sheets-Sheet 1

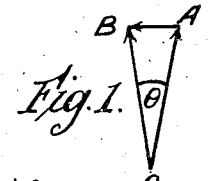



Fig. 1.

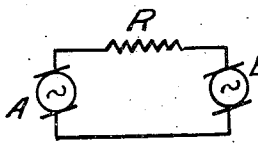



Fig. 2.

Fig. 3.

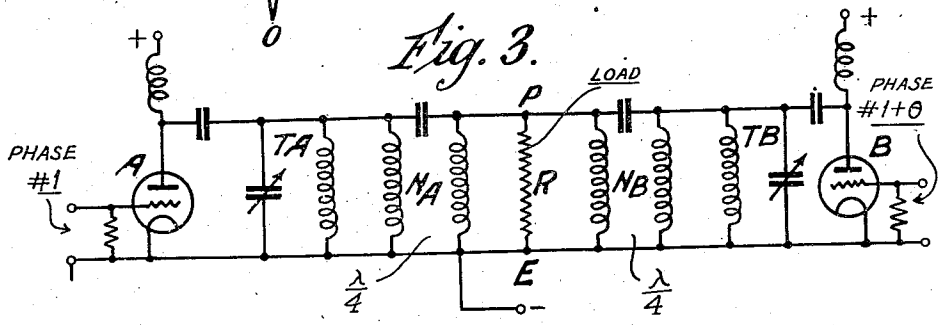



Fig. 4.

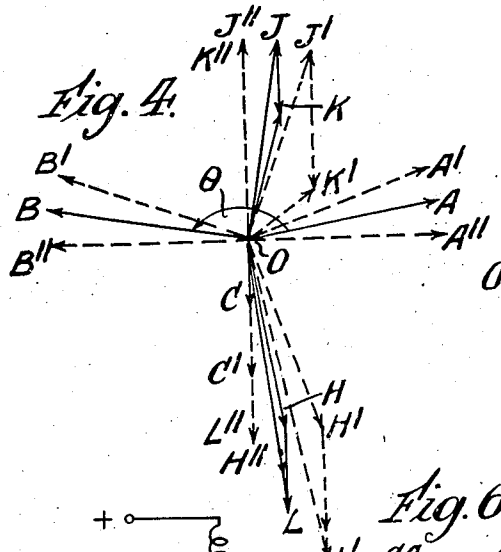



Fig. 5.

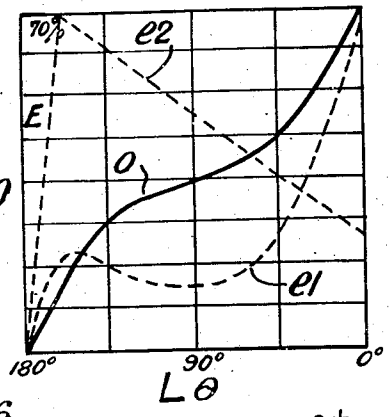



Fig. 6.

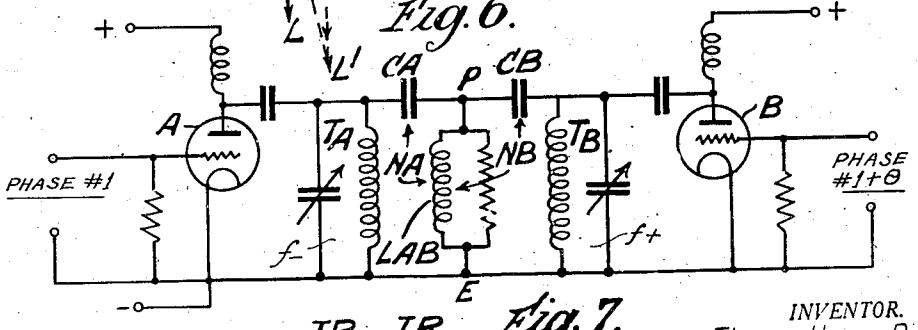
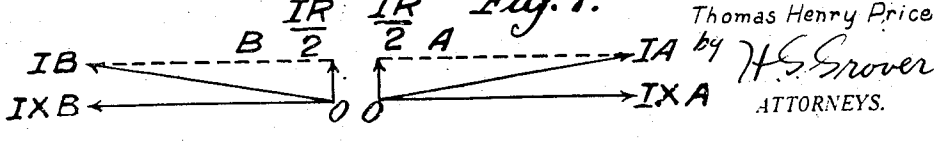



Fig. 7.

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RADIO TRANSMITTER

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Fig. 8.

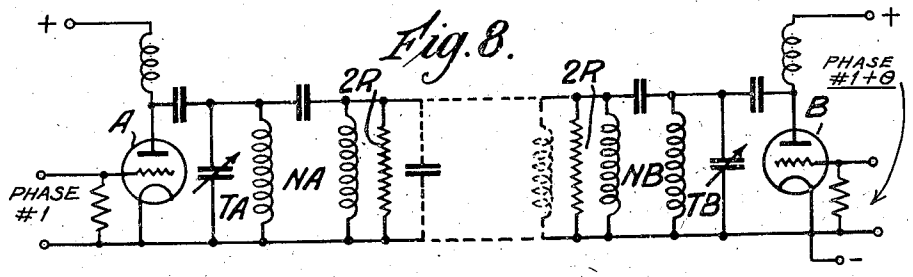



Fig. 9.




Fig. 10.

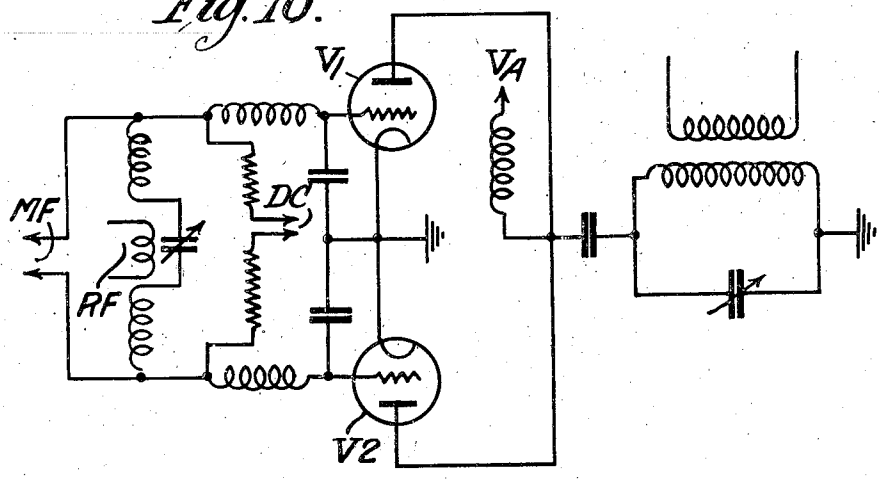
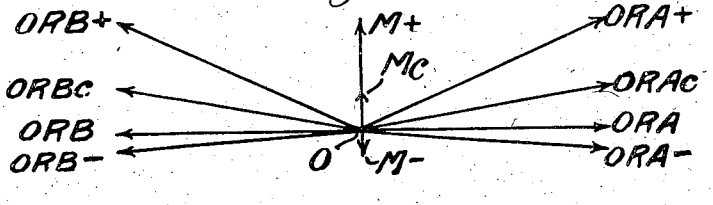



Fig. 12.

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RADIO TRANSMITTER

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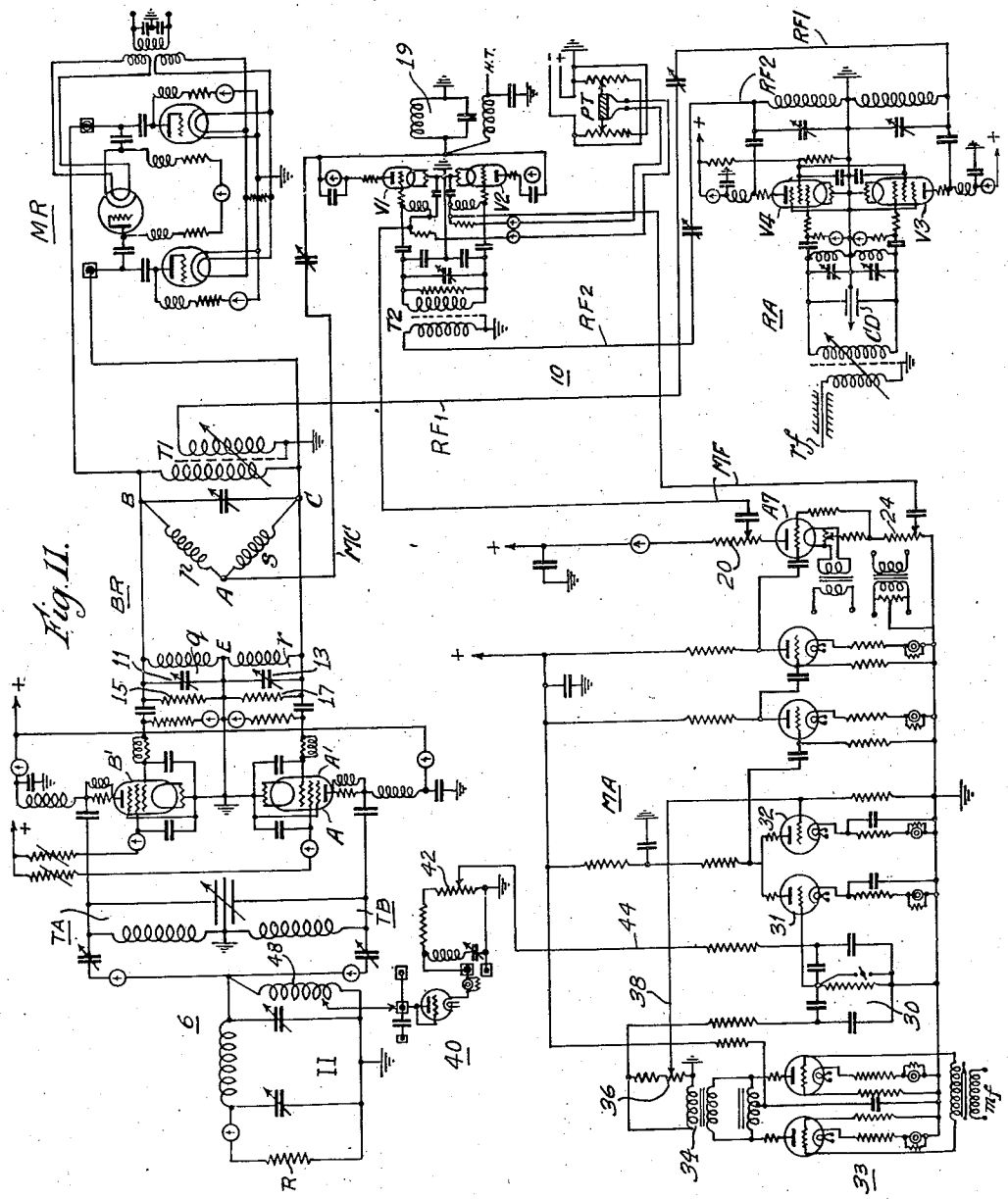


Fig. II.

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Fig. 13.

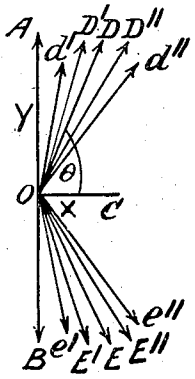
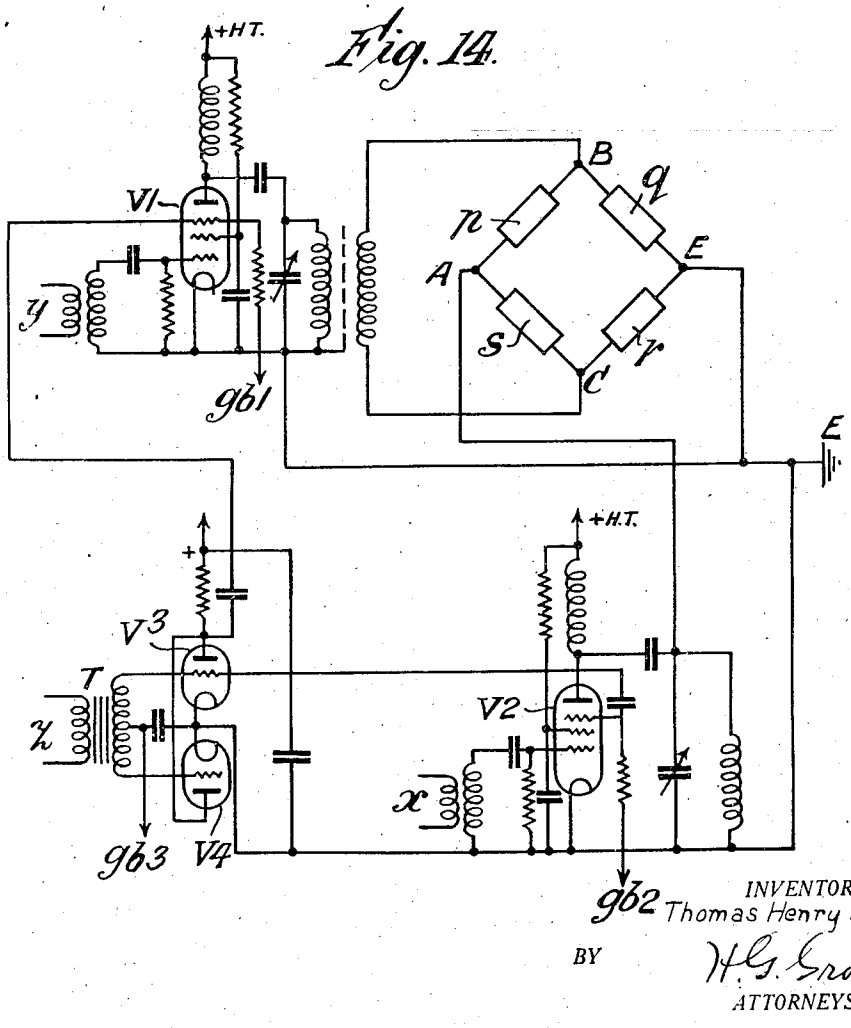



Fig. 14.

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Patented Aug. 7, 1945

2,381,181

# UNITED STATES PATENT OFFICE

2,381,181

## RADIO TRANSMITTER

Thomas Henry Price, Chelmsford, England, assignor to Radio Corporation of America, a corporation of Delaware

Application April 5, 1943, Serial No. 481,822
In Great Britain January 16, 1942

16 Claims. (Cl. 250—17)

The present invention relates to radio transmitters, and aims at improving such transmitters towards the production of a simple transmitter capable of passing to a power-absorbing device, such as an aerial, a completely modulated wave the transmitter being desired to have the maximum possible efficiency of conversion of supplied power to output power, and being desired also to satisfy other various requirements of a high-performance transmitter.

Such an aim is, of course, that which many workers in the field of radio-engineering have had before them, and many types of radio-transmitter have been produced achieving, to a greater or less degree, this aim, or at any rate, satisfying some of the demands to be made upon a perfect radio-transmitter.

Fig. 1 is a vector diagram used to explain the nature of my invention. Fig. 2 is a basic diagram of two generators supplying voltages to a common load.

Fig. 3 is a simplified diagram of two generators arranged in accordance with my invention to supply power to a common load.

Fig. 4 is a vector diagram illustrating the operation of the generators of Fig. 3.

Fig. 5 shows by curves the relation between current in the load and the phase relation of the currents flowing toward the load; the ratio of total output power to total input power, etc.

Fig. 6 illustrates by simplified diagram a circuit arranged in accordance with my invention.

Fig. 7 is a vector diagram illustrating the operation of the system.

Fig. 8 is a modification of the arrangements of Fig. 6.

Fig. 9 is a vector diagram illustrating operating characteristics for 100% modulation on the negative half cycles of the modulation.

Fig. 10 is a modification arranged for operation as illustrated by the vector diagram of Fig. 9.

Fig. 11 illustrates somewhat completely a phase swing transmitter arranged in accordance with my invention.

Fig. 12 is a vector diagram illustrating the operation of the arrangement of Fig 11; and

Fig. 13 is a vector diagram used to illustrate the operation of the arrangement of Fig. 1.

Fig. 14 is a modification of the arrangement of Fig. 11.

In the present state of the art of radio-transmission, thermionic valves have to be used and the best efficiency of conversion that can be obtained is that in which the valves are operated in their class "C" mode of operation.

In carrying out the present invention, two valves (or two groups of valves) each valve (or all the valves of each group) operating in its (or their) class "C" mode of operation, are employed and a power-absorbing device, for example an aerial, is suitably coupled between the two valves (or groups of valves). Oscillatory energy is applied to the two valves (or groups of valves) and the relative phase positions of the oscillations applied to the two valves (or groups of valves) are so operated upon during modulation as to produce a variation of the output power, such variation according with the modulation signals applied to the radio-transmitter. Such a transmitter may be, and hereinafter will be, referred to as a phase-swing transmitter.

Hereinafter, for greater simplicity of language, the phrase "two valves, or groups of valves," will be avoided and the phrase "two valve-stages" will generally be used alone to indicate the same idea. This is to be understood not to imply a limitation.

According to the invention, each of two valve-stages of a phase-swing transmitter is coupled to a power-absorbing device, for example an aerial, by a quarter-wave impedance-inverting network or device.

In considering the matter of connecting or coupling two valves to a power-absorbing device so as to produce a phase-swing transmitter, it is found, that it is not possible merely to connect or couple the two valves to the power-absorbing device, and, at the same time, to retain high efficiency and to provide for the necessary linearity of output with respect to modulation signal at the input, because power input to the power absorbing device cannot be in phase with the voltages generated by both of the two valves.

The vector diagram constituting Figure 1 of the drawings illustrates how this comes about. The two vectors OA and OB represent the voltages generated by two alternating current generators (A and B, respectively of Figure 2) having a phase difference represented by $\theta$. The resultant voltage is represented by A, B and this is the effective voltage across a resistive load R connected from one generator to the other. Current in such a load would be in phase with AB and, therefore, out of phase with both OA and OB. In practice, owing to the dependency of the resistance-current upon $\theta$, small values of $\theta$ would have to be employed in order to obtain linearity of resistance-current for different values of $\theta$. This in its turn would cause a still greater phase difference between the power com-

ponent of current and the voltage vectors. This effect could no doubt be compensated for by the use of suitable shunt reactances across the generators, but this solution would introduce the difficulty that as $\theta$ is made to disappear a considerable component of reactance-current would be drawn from the generators. In the case of valve generators this would be particularly objectionable in that the efficiency would be low. Therefore, a circuit is required with which the power component of current is held closely in-phase with the generator voltages throughout the cycle of changes of relatively phase of these voltages.

An embodiment of the invention is illustrated in Figure 3 of the drawings. In this figure, A and B are two similar valves to the grids of which in operation high or radio-frequency oscillatory energy is applied in antiphase. The anode circuit of each of these valves includes a tuned circuit, respectively TA and TB, these two circuits being similar, the one to the other. Each tuned circuit is coupled to the power-absorbing device, represented by the resistor R, by means of a quarter-wave impedance invertor, illustrated, respectively, as networks NA and NB. The two valves A and B operate in their class C mode and deliver closely similar voltages which, since they are in antiphase, and since the two impedance invertors are similar, produce zero voltage across points P and E when both tuned circuits are tuned to resistive condition for the valves.

The vector diagram constituting Figure 4 of the drawings illustrates the operation of the arrangement of Figure 3. In this figure OA and OB respectively represent the currents, dephased by the angle $\theta$, assumed to be 160°, flowing towards point P of Figure 3 from networks NA and NB in the initial, or unmodulated condition. (A discussion on the choice of $\theta$=160° is given below.) Since one of the properties of a quarter-wave network is to cause the current flowing out of one end to be out of phase to the extent of 90° with the voltage applied at the other end, the current having a constant amplitude for a constant applied voltage, the voltages across the inputs of the networks, represented respectively by vectors OJ and OH, are 90° behind the currents represented respectively by vectors OA and OB. (Here it may be observed, that whether OJ and OH shall lag behind or lead OA and OB, will depend upon the kind of quarter-wave network employed.) The voltage, in this condition, across, and the current in, resistor R, are represented by vector OC. The two triangles OKJ and OHL then respectively represent the voltages around networks NA and NB, respectively, OK and OL being the voltages across the series limbs of the respective networks. (For convenience unity ratios of voltages and corresponding currents is assumed.)

Vectors OA' and OB' respectively represent the currents flowing towards point P of Figure 3 from networks NA and NB at the instant of peak positive modulation for 100% modulation, this condition being represented by the vector OC' which represents the voltage across and current in resistor R, and which is twice the length of vector OC. For this instant the voltages across the inputs of the networks are OJ' and OH' and are shifted in phase in opposite senses from their initial positions towards each other under the influence of modulation signals applied to the grids of valves A and B. The voltage across the series limbs are now OK' and OL'.

Vectors OA'' and OB'' respectively represent the currents flowing towards point P of Figure 3 from networks NA and NB at the instant of peak negative modulation. In this condition the voltage across and current in resistor R is zero. For this instant the voltages across the inputs of the networks are OJ'' and OH'' and are shifted in phase in opposite senses from their initial positions away from each other, under the influence of modulation signals applied to the grids of valves A and B. The voltage OJ'' across the input of network NA is in phase with the voltage OK'' across the series arm of network NA, and the voltage OH'' across the input of the network NB is in phase with the voltage OL'' across the series arm of network NB. Voltages OJ'' and OH'' are in anti-phase with each other as are, consequently OK'' and OL''.

Figure 5 of the drawings shows three curves $o$, $e_1$ and $e_2$, $o$ and $e_1$ relating to the arrangement of Figure 3. All three of the curves assume negligible loss in the circuits.

Curve $o$ connects the angle $\theta$, which is the phase difference between the vectors of the oscillatory voltages at the anodes of the two valves, with output current O flowing in the power absorbing device R. The angle $\theta$ is assumed for the preparation of the curves to be varied from 180° to 0°. Curve O shows that there are linear relationships between $\theta$ and O, for high, and for low, values of $\theta$ and a non-linear relationship about the middle of the curve. Also it shows that O is a maximum when $\theta$ is zero and is zero when $\theta$ is 180°. Since systems of modulation used in the faithful transmission of speech and music require, that in the unmodulated condition there should be a given level of carrier power, that this should be reduced to zero for the trough of the modulating signal, and that it should be increased to four times that corresponding to the unmodulated condition for the peak of the modulating signal, and since such systems also require that the change in the level of output current should correspond substantially linearly with the change in modulating signal, the usable portion of the curve O is determined, and an angle $\theta$, for the unmodulated condition, of approximately 160° is chosen.

Curve $e_1$ connects the ratio of total output power to total anode input power with the angle $\theta$ and shows the % efficiencies E of the system of Figure 3 at different values of $\theta$. This curve has two peaks, one at an angle $\theta$ near to anti-phase, and one at a zero angle $\theta$. The variation in efficiency is due to the departure from a unity power factor for changes in the angle $\theta$.

It is clearly desirable that the greatest efficiency should be found in the region of the usable portion of the curve O, and that the efficiency should vary as little as possible during the cycle of change in phase, that is along the lower linear part of curve $o$.

Figure 6 illustrates a circuit arrangement wherein the desired conditions are met. In this figure, the tuned circuits are represented at TA and TB as before, and the two impedance inverting networks comprising the two capacitors CA and CB and the inductor LAB. In this circuit the shunt inductors of the two impedance inverting networks of Figure 3 have been replaced by an equivalent single inductor LAB. (It is again to be noted that, for the purpose of description, series capacitors and shunt inductors are assumed; they might be series inductors and shunt capacitors. The last statement general-

ised, then becomes the shunt reactors of the two impedance inverting networks have been replaced by an equivalent single reactor.) For a reason that will appear, the two tuned circuits TA and TB are slightly mistuned differentially from their parallel resonant condition, so as to provide an effective resistive or unitary power factor for the valves.

Considering the two vectors OA and OB of Figure 4, which represent the currents flowing towards the point P of Figure 3 from the two networks NA and NB, it is apparent that these two vectors represent resultants which can be resolved into their two components in quadrature as represented in Figure 7 where the resultants are named IA and IB and the components of either are represented by IX and

$$\frac{IR}{2}$$

with the appropriate subscript. Vector

$$\frac{IR}{2}$$

represents the current flowing in resistance R (Figure 3) due to the appropriate valve A or B, and IX represents the reactive-current necessarily to be supplied with

$$\frac{IR}{2}$$

to provide the resultant. This resolution shows that to satisfy the condition that the sum of the currents flowing to point P be zero, each of the valve amplifiers with its associated network and its portion of the terminating resistance R, is to be regarded as having an additional reactance for its termination. The reactance for each half of the system is represented by the other half. Whereas the terminating reactance is positive for one half, it is negative for the other. This effective condition for each side of the system is illustrated in Figure 8 where the reactive load of the one valve generator and its network on the other valve generator and its network is represented in dotted line.

This is merely the theoretical dissection of the practical circuit of Figure 6, in which reactances of similar kind in parallel have been replaced by equivalent single reactances. The resistive termination of each network is represented as 2R, which is, of course, equivalent when paralleled to R.

Now reverting to Figure 6 it is clear that if the anode tuned circuits are in parallel resonance and their associated impedance inverting networks are correctly designed, and are terminated with a combination of resistance and reactance, both valve generators will have a reactive load condition, and an effective mistuned state will exist for both, but whereas the one valve has a positive reactive load, the other has a negative reactive load. If symmetry of electric constants for the two sides of the circuit has been provided, an effective resistive or unity power factor condition for the valves can be restored by increasing the reactance value of a limb of one anode tuned circuit and reducing the reactance value of the corresponding limb of the other anode tuned circuit. This operation might be called differential retuning of the tank circuits, and may quite conveniently be carried out on either the capacity or inductance limbs. Ganging the two chosen variables to produce the required differential effect is one easy way of achieving the object, provided that the action of controlling produces equal and opposite changes in the reactors concerned.

If care has been taken in the choice of components and constants of the circuits, power will now appear in the load resistor R of the frequency of that applied to the grids of the two valves at an efficiency ratio of output-power/total-anode power very little different from that which would obtain in the simplest possible arrangement of valves and circuits. If the valves be driven in Class "C" condition the efficiency will be very closely that generally obtained in class "C" operation, that is about 70%. The amount by which the efficiency departs from that for optimum class "C" operation will depend on the losses which occur in the circuits. These can easily be kept relatively low, particularly in cases of high-power transmitter systems. The efficiency curve for the arrangement of Figure 6 is shown as $e_2$ in Figure 5.

It has been mentioned that since one side of the phase-swing system oscillates at a different phase from that at which the other side oscillates, reaction loads are reflected into the input ends of the two networks NA and NB and that to maintain unity power-factor load for the two valves the associated anode tuned circuits are differentially mistuned. This may be effected by effecting a differential change in the values of the capacitative or inductive limbs, preferably, however, in the values of those reactors which are opposite in type to the type used as the series elements of the networks NA or NB.

The exact amount of mistuning is that which will give substantially class "C" anode efficiency for the two valves of the system when the grid exciting voltages are dephased by the angle $\theta$, chosen as the operating angle and being about 160°.

This mistuning operation for unity power factor for the unmodulated condition produces, for any different angle between relative phases of the driving voltages, such as occurs throughout the modulation cycle, a slightly reactive load condition for each valve, and this causes the phase of the actual anode voltage to lag slightly on the phase of the exciting voltage applied to the grid, thus necessitating an increased degree of relative change in phase on the grids than would otherwise be required.

This state of reactive loads on the anodes of the valves during modulation requires, in order that 100% modulation for the negative half of modulating may be possible, that the relative phases of excitations of the grids of the valves be increased to somewhat more than 180° into the other quadrants of the vector diagram, Figure 9.

In this Figure 9 OA and OB represent the grid vectors for the two valves of the system in the carrier power setting, OA' and OB' representing these vectors at the positive peak of modulation, and OA'' and OB'' representing the same vectors at the negative peak of modulation.

To obtain this mode of operation of the grid exciting vectors, a valve and circuit assembly, usually called a carrier suppressor system, is used to energise the grids of valves A and B. A typical arrangement of such a system is given in Figure 10 of the drawings.

In Figure 10 valves V1 and V2 receive on their grids antiphase radio-frequency signals from a source not shown but represented at transformer RF, of constant amplitude. These radio-frequency or carrier signals are superimposed on the fixed direct current grid bias voltage, applied as at DC, which is so chosen that at the anode voltage VA very little anode current occurs for either valve and the output power from V1 and V2 is zero until a modulation signal is impressed at MF. That is, the valves work effectively in a class "B" condition. They could be set up to work in the class "A" condition also, where both valves would be permitted to pass approximately equal anode currents in the absence of modulation. This however would be an unnecessary dissipation of power when the stage is unmodulated.

The modulation signal MF is applied to the grids of valves V1 and V2 in antiphase so that that valve to the grid of which a positive half-cycle of modulation potential is being applied will act as a grid modulated amplifier, whilst the valve, to the grid of which the negative-half cycle of modulation potential is being simultaneously applied will be held in the cut off condition in spite of the signal superimposed on its anode due to the valve which, at the time, is acting as a grid-modulated amplifier. The roles of the two valves, of course, alternate. The circuit of Figure 10 will be recognised as that of a "carrier suppressor system" but for the purpose for which it is required in the present invention, namely to remove the vectors OA and OB from exact phase-opposition in the unmodulated condition some carrier is required. This requirement may be very simply met by arranging that the DC bias voltages for the grids of the valves V1 and V2 be obtained from two potentiometers ganged so that operation of the controlling means reduces the bias applied to the grid of one valve as much as the bias applied to the grid of the other is increased.

The differential biasing upsets the balance of the carrier suppressor system, and if the initial working point on the characteristic curves for these valves is suitably chosen almost perfectly linear change of high frequency output voltage against change of low frequency grid voltage is obtained from the two valves.

The output from the carrier suppressor system which in the conditions of operation above described consists of modulated radio-frequency energy with the carrier not entirely suppressed in the unmodulated condition, is applied in push-push fashion to the grids of valves A and B of the power stage, together with unmodulated radio-frequency energy the unmodulated energy being applied in push-pull fashion and being in phase-quadrature with the modulated energy.

Figure 11 illustrates a complete transmitter. In this figure, 6 represents the power stage and 10 represents the carrier suppressor (as it may continue to be called), MA represents a modulation frequency amplifier of conventional design, having its last stage constituted as a phase-splitter including potentiometers 20 and 24. In the amplifier shown feedback as disclosed in Fig. 4 of U. S. application Serial #429,144, filed Feb. 2, 1942, may be used. Modulating energy is applied at *mf* and an output, MF, in phase opposition appears in the leads MF.

RA represents a radio-frequency amplifier of suitable design to which radio-frequency energy is applied as at *rf*. Unmodulated radio-frequency carrier energy of smoothly controllable relative phase appears in leads RF1 and RF2.

Amplified modulation frequency is applied, over leads MF, in phase opposition to the grids of valves V1 and V2 in the carrier suppressor system 10. Radio-frequency energy is applied over lead RF2 and by way of transformer T2, in push-pull, to the grids of valves V1 and V2. Direct current bias potentials are applied to these same grids by and are adjustable, differentially, by means of potentiometer PT. Modulated radio frequency carrier appears in lead MC.

Modulated carrier from carrier suppressor system 10 is applied, over lead MC, between points A and E of a bridge BR, having arms *p*, *q*, *r*, *s*. Voltages due to this energy appear as components of the resultant voltages applied to the grids of valves A' and B' in power stage 6, and are represented by the vector $OM_c$ in Figure 12 These voltages appear in phase on the two grids.

Unmodulated radio-frequency carrier is applied over lead RF1, and transformer T1, between points B and C of bridge BR, and appears as voltages in phase opposition, as further components of the resultant voltages applied to the grids of valves A' and B'. They are represented, in Figure 12 by the vectors ORA and ORB. Voltages OM and ORA, and OM and ORB are in phase quadrature. Precise quadrature setting is obtained by adjusting the phase adjusting differential condenser CD in the grid circuits of valves V3, and V4.

The resultant voltages, in the unmodulated carrier condition, applied to the grids of valves A' and B' are represented in Figure 12 by the vectors $ORA_c$ and $ORB_c$.

During one-half-cycle of modulation the energy supplied to the bridge BR over lead MC rises to a given amplitude and during the other half-cycle of modulation this energy rises to approximately the same amplitude but in reversed phase. The vectors representing the components of voltages applied to the grids of valves A' and B', and due to these two conditions of modulation, are named OM+ and OM— in Figure 12 and give rise to resultants CRA+ and ORB+ or ORA— and ORB— respectively.

The grid tuned-circuits of valves A and B, comprising condensers 11 and 13 with their shunt damping resistors 15 and 17, and the other components leading back to the anode tuned circuit 19 of carrier reducer or suppressor modulator 10 are so chosen as to constitute impedance invertors so that undesired influences, due to varying grid current in the valves of the power stage 6, upon the phases of the applied voltages may be minimised or eliminated.

As explained hereinbefore in showing how the arrangement of Fig. 6 is derived from the arrangement of Fig. 3, the anodes of both the stages A' and B' are coupled by impedance invertors to the common load R fed through impedance transformer II. This will be obvious when it is kept in mind that each of the two anode load impedance invertors has its shunt impedance limb lumped with the anode tuned circuit (TA and TB) inductance of the appropriate tube. The other shunt inductance limbs of the impedance invertors are lumped together and form the inductance 48 shown across the input end of the network II.

The phase-splitting valve A1 has potentiometers 20 and 24 as anode and cathode leads, the output voltages from these being in phase op-

position and adjustable in relative value so as to compensate for any non-linear effect of the varying load of the power stages over the cycle of modulation. A varying load on the power valves causes a varying high frequency voltage at the anodes—a rising voltage for the negative half and a falling voltage for the positive half cycle of modulation. A greater degree of phase swing of the controlling voltages for the negative half than for the positive half of modulation corrects this unequal voltage condition. The potentiometers **20** and **24** enable this correction to be obtained, the effect being observed on a cathode ray oscillograph modulation check equipment.

Included in the modulating stages are a pair of valves **31** and **32** with a system of 30 simple resistance-capacity filters to permit of a considerable level of feedback. The modulation is applied at *mf*, amplified in stage **33** to appear in transformer **34** and across the potentiometer resistance **36** and in a balancing network **30**. A portion of the output of stage **33** is taken by lead **38** and also applied to the network **30** which is connected to the grid of modulation amplifier **31** wherein the modulation fed forward and fed backward are combined. Feedback energy is also derived from rectifier **40** and fed by potentiometer **42** and lead **44** to the balancing network **30**. Noise and hum are blocked out in tube **31** as described more in detail in my U. S. application Serial #429,144, filed Feb. 2, 1942. The amplifier is otherwise conventional.

Use of feedback allows of 50~ A. C. lighting of all the valves of the system with a resultant carrier noise which is well within the usual accepted figures.

Reference MR indicates a monitor which need not be described, since a similar monitor and its mode of operation have been described in British Patent No. 540,245, dated October 10, 1941, and the corresponding U. S. Patent #2,294,800, dated September 1, 1942.

Tetrodes or pentodes may be used in all stages of the phase swing transmitter thus eliminating the need for neutralising over a considerable range of wavelengths, and also providing a still better overall efficiency due to the low driving power required by tetrodes and pentodes.

A phase-swing transmitter, such as that described in connection with Figure 11, provides normal modulated carrier. It can, however, by slight adjustments, be readily made to operate as a suppressed carrier transmitter. These adjustments consist in balancing the carrier-suppressor system **10** and adjusting the anode tuned circuits of the power stage **6** and their respective networks and load to be in parallel-resonance for an anti-phase excitation of the grids of valves A′ and B′ (which anti-phase condition will arise from the balancing of the carrier suppressor system **10**).

In an increasing number of applications of thermionic vacuum tubes, it is desired to have the nearest approach to perfect linearity between one quantity, which may be called X, and another quantity, which may be proportional to a trigonometrical function of X. A special case is that in which the phase angle of an electrical quantity is desired to be varied, usually by linearly varying a modulation or modulated signal X to provide a strictly proportional variation in the value of a trigonometrical function of X.

If the change of angle be small it can often be arranged that sufficiently nearly linear relationship between the two quantities will be obtained without recourse to special devices. But if an extended angle of change with good linearity be desired, some special device may become essential. The following describes one method of achieving the purpose.

If two vector quantities Y and X, which are in quadrature relation, be added vectorially, the resultant vector may have a sine value, which may be expressed by

$$\text{sine } \theta = \frac{Y}{\sqrt{(X^2+Y^2)}}$$

and a cosine value

$$\text{cosine } \theta = \frac{X}{\sqrt{(X^2+Y^2)}}$$

(see Figure 13), where $\theta$ is the angle between the resultant vector and the quantity X. A variation in the value of these expressions can be obtained by varying the value of X alone, whilst Y is kept constant.

The expressions for the special case mentioned can be written:

$$\text{sine}\left(\tan^{-1}\frac{Y}{\sqrt{(X^2+Y^2)}}\right)$$

and

$$\text{cosine}\left(\tan^{-1}\frac{Y}{\sqrt{(X^2+Y^2)}}\right)$$

In many cases X will bear a value which, for successive instants of time, can be expressed as some fraction of Y, say KY. If curves be plotted giving the values of the above expressions against changing values of X, it will be seen that linearity of relationship is restricted to small changes of X when X is much smaller than, or nearly equal to Y.

The device here proposed effects an improvement in the relationship for a greater variation of X, and yields a result, which for many practical purposes would be regarded as sufficiently near a linear relationship of the quantities determining the curves.

In operating the device the quantity Y is modulated during the period of changing X, when non-linearity of the trigonometrical functions rapidly increases, the rate of changing of the value Y being determined by some influence, such as that of the bottom or top bend of thermionic vacuum tube characteristic curves.

The special case of the application of the idea to the cosine function, when X is smaller than Y, is illustrated in the diagram Figure 14. The case of the sine function and other trigonometrical functions will not be discussed further, as the special requirements for these will be clear to those skilled in the art of designing thermionic vacuum tube circuits for special purposes.

The quantity Y is generated by valve V1, Figure 14, and applied to two corners BC of the bridge formation, *p, q, r, s*. The quantity X, which we will assume to be KY, is generated by valve V2 and applied to corners AE of the bridge formation *p, q, r, s*. The valves V1 and V2 are shown as pentode valves and receive adequate excitation voltages on their control grids by signals applied at *x* and *y* respectively. These signal voltages, which are of the same radio or signal frequency, are arranged to be in quadrature phase relationship. Valves V3 and V4 receive antiphase audio or modulating voltages on their control grids and have their anodes strapped together,

the load resistance for these valves being a common load inserted in the common anode feed connection. These valves are so biased on their control grids that in the absence of any modulating signal the anode current is zero, or very little, depending on the correction which their action is desired to apply to the vector Y, generated by valve V1. Valve V1 is given sufficient grid bias *gb1* to accommodate the correction voltage, which will be impressed on its third grid from valves V3 and V4.

One mode of operation would be that in the absence of modulation at *z*, valve V2 would be so biased by *gb2*, on its control grid that it generated a small signal $X=KY$ (OC of Figure 13) whilst valve V1 is generating a considerable amplitude of signal Y, which appears across the bridge *p*, *q*, *r*, *s*, and may be represented by vectors OA and OB of Figure 13, the resultant vectors being OD, appearing across one arm, say *q*, of the bridge *p*, *q*, *r*, *s*, and OE appearing across another arm, say *r* of the bridge.

When modulation appears at *z* it is passed via the transformer T to the control grids of valves V3 and V4 in antiphase and from one half of the secondary winding of the transformer to the third grid of valve V2. If valves V3 and V4 are sufficiently biased on their control grids they will have no influence on the performance of the circuit. Under the influence of valve V2, OC of Figure 13 is modulated to a degree depending on the signal impressed on V2 from transformer T. The resultant vectors might then describe a change of $\theta$ ranging from OD' to OD'' and from OE' to OE''. Over this or an extended range of variation of X, i. e. OC, there would be some non-linearity of the cosine function.

If now the grid bias *gb3*, of valves V3 and V4 be reduced so as to permit them to draw anode current alternately through their common anode load, under the influence of their antiphase grid signals, this would have the effect of a progressive reduction in the value of the Y vector generated by valve V1 for reduced value of the grid bias *gb3*, the rate of reduction being adjusted by the effect the changing bias has of moving the working point along the bottom bend of the characteristics of valves V3 and V4.

The resultant vectors of Figure 13 might then range between Od' and Od'' and between Oe' and Oe'', giving a greater range of variation $\theta$, but an improved linearity of the relation $\cos\theta/X$.

The absolute amplitude of the resultant vectors might vary, but for many applications the variation would be unimportant, or could be nullified by various contrivances.

A number of variations of the circuits are possible.

In the case of the phase swing transmitter it may be shown that the expression for the output current is

$$IR=2I\cos\frac{\theta}{2}$$

where I is the current flowing out from the two networks NA and NB of the scheme, and $\theta$ is the operating angle between these two currents. The angle $\theta$ is caused to vary by linearly varying the output from the carrier suppressor stage. In other words the variation of the operating angle is produced by the variation of just such a quantity X as has been discussed in the foregoing. By adopting the idea described above the choice of operating angle for the phase swing transmitter is not so restricted. The modifications of Figure 14 for application to the phase swing transmitter consists merely in the substitution of a carrier suppressor stage for the stage of V2 of Figure 14 and the addition of valves corresponding to valves V3 and V4 of Figure 14. The modulation transformer T of Figure 14 would be dispensed with by suitably applying to the valves V3 and V4 the same anti-phase modulation signal as is applied to the carrier suppressor valves of Figure 11. The output signal from these linearity correcting valves V3 and V4 would be applied to a control grid of the valve A5 of Figure 11 which corresponds with the valve V1 of Figure 14.

This device which has been described and whose purpose is to reduce the restriction in the choice of operating angle of the phase swing transmitter is to be regarded as a refinement and would be of real value mainly in the case where the phase swing transmitter is to be used as a suppressed carrier transmitter. In such a usage it might be desired to use a bigger angle of phase swing. The device described would permit of this without objectionable effects in the matter of linearity of response of the transmitter.

Having now particularly described and ascertained the nature of my said invention and in what manner the same is to be performed, I declare that what I claim is:

1. In a phase-swing radio-frequency transmitter, a power-absorbing device, two-tube stages each having input and output electrodes, connections for applying radio-frequency oscillatory energies to the input electrodes of each stage the relative phase positions of which energies are varied during modulation, and a quarter wave impedance-inverting device or network coupling the output electrodes of each tube stage to said power-absorbing device.

2. In a phase-swing radio-frequency transmitter, two similar electron discharge tubes having grids and anodes, means for feeding to said grids radio frequency voltages in predetermined phase relationship and in which during modulation the phase positions of the voltages are caused to swing towards and away from each other about the unmodulated position, a tuned circuit coupled to the anode of each tube, the said tuned circuits being substantially similar to each other, a power-absorbing device coupled to each tuned circuit by means of a quarter-wave impedance-inverting device or network, the arrangement being such that as a result of varying the said phase positions the amplitude of the current delivered to the power-absorbing device is substantially linearly varied.

3. A transmitter as claimed in claim 1, wherein each separate quarter-wave impedance inverting network comprises a pair of conductors coupling the output electrodes of each tube stage to the power-absorbing device with reactances connected across the conductors of each pair, and a reactance connected in series with one of the said conductors of each pair at a position between two of said shunting reactances.

4. A transmitter as recited in claim 2, wherein each network includes a pair of conductors coupling the tuned circuit to the load with two reactances connected across each pair of conductors and a reactance connected in series in one conductor of each pair at a position therein between said first two reactances.

5. A transmitter as recited in claim 2, wherein the separate quarter-wave impedance inverting networks each include a reactance common

to the two networks connected across the power-absorbing device and each includes a reactance in series between the tuned circuits and the power-absorbing device.

6. A transmitter as claimed in claim 2, wherein the two tuned circuits are parallel tuned circuits and are slightly mistuned differentially from parallel resonant condition, so as to provide an effective resistive or unitary power factor for the tubes.

7. A transmitter as claimed in claim 2, wherein the two tuned circuits are parallel tuned circuits and are slightly mistuned differentially from parallel resonant condition and wherein the networks each include a reactance and the reactances are similar, and wherein common control means is connected to said similar reactances to produce equal and opposite changes in the tuning thereof.

8. A transmitter as claimed in claim 2, wherein the two quarter-wave networks each include a series reactance and the said reactances are similar and each thereof are included one in each of the tuned circuits and wherein the tuned circuits are differentially mistuned by differentially varying the reactive value of said series reactances by gang control so as to produce equal and opposite changes in the tuning thereof.

9. A transmitter as recited in claim 1, wherein the phase separation between the energies applied to the input electrodes of the tube stages is in the unmodulated condition substantially 160°.

10. A transmitter as claimed in claim 2, wherein said phase relation is about 160° in the unmodulated condition and during modulation varies around said unmodulated 160° relation, and output electrodes of each tube-stage looks into a substantially resistive load presented by the power-absorbing device and the other tube-stage and its associated tuned circuit and quarter-wave impedance-inverting device or network and whereby, to meet the varying reactive load presented by these elements during modulation the angle separating the phase relation of the energies or voltages is during the negative half-cycle of modulation caused to increase beyond 180°.

11. A transmitter as recited in claim 2, wherein said radio voltages of predetermined phase relationship fed to the grids are resultant voltages made up of two components and wherein means is provided for producing modulated radio-frequency energy (or voltage) with the carrier not entirely suppressed in the unmodulated condition, and feeding the same as one component in push-push fashion to the grids of the tubes, and other means is provided for feeding unmodulated high or radio-frequency energy as the other component in push-pull fashion to the grids of the tubes, the latter unmodulated energy being in phase quadrature with the modulated energy.

12. In a transmitter system, a power-absorbing device, a power stage including two electron discharge devices each having input and output electrodes, a tuned circuit coupled to the output electrodes of one of the devices of said power stage, a tuned circuit coupled to the output electrodes of the other of the devices of said power stage, a quarter-wave impedance-inverting network coupling each of said tuned circuits to said power-absorbing device, a carrier suppressor modulation system including two tubes each having input and output electrodes, a source of modulation current, a source of radio-frequency voltage, connections from said source of modulation current to said input electrodes for applying the modulation current in push-pull relation to the input electrodes of said two tubes, connections from said source of radio-frequency voltage to the input electrodes of said two tubes for applying radio-frequency voltage thereto in push-pull relation, a bridge circuit having one diagonal coupled between the input electrodes of said electron discharge devices and having a second diagonal coupled to the output electrodes of said tubes to feed carrier suppressed modulated radio-frequency energy in phase to the input electrodes of said two devices of the power stage, a radio-frequency amplifier stage having an input coupled to said source of radio-frequency voltage and an output coupled to said one diagonal of said bridge circuit to feed unmodulated radio-frequency energy by way of said bridge to the input electrodes of said two discharge devices of the power stage in phase opposition, the in-phase modulated energy applied to the input electrodes of said discharge devices being in phase quadrature with the phase opposed unmodulated energy supplied to the input electrodes of said devices whereby resultant radio-frequency voltages are produced on the input electrodes of the devices of the power stage which in the unmodulated condition are separated by a phase angle of substantially 160°.

13. A system as recited in claim 12, wherein the tubes are so biased and operated that the carrier suppression is incomplete, and wherein the modulating current is so applied to the input electrodes of the two tubes that the in-phase radio-frequency energy applied by the carrier suppressor system to the input electrodes of the two discharge devices increases and decreases in amplitude in accordance with alternate half-cycles of modulation energy above and below the amplitude of the incompletely suppressed carrier applied in the unmodulated condition.

14. A system as recited in claim 12, wherein the input electrodes of the tubes of the carrier suppressor modulator are so biased that the modulating currents do not operate the tubes to cut-off on the negative cycles of the modulation whereby the carrier is not completely suppressed.

15. A system as recited in claim 12, wherein a tuned circuit is coupled to the output electrodes of said two tubes and to the said second diagonal of said bridge circuit.

16. In a transmitter system, a power-absorbing device, a power stage comprising two electron discharge devices each having input and output electrodes, a coupling between the output electrodes of one of the discharge devices and said absorbing device, a coupling between the output electrodes of the other of the discharge devices and said absorbing device, a pair of tubes each having input and output electrodes, two other tubes each having input and output electrodes, a source of modulation current, a source of radio-frequency voltage, connections for applying the modulating current to the input electrodes of one of said other tubes, connections for applying the modulation current in push-pull relation to the input electrodes of said pair of tubes, a connection between the output electrodes of said pair of tubes and an input electrode of the other of said other two tubes, connections from said source of radio-frequency voltage to the input electrodes of said two tubes for applying radio-frequency voltage thereto substantially in phase quadrature relation, a

bridge circuit having two diagonals, a coupling between the input electrodes of said electron discharge devices and one diagonal of said bridge, a coupling between the output electrodes of said other of said two tubes and said one diagonal to feed radio-frequency energy in phase opposed relation to the input electrodes of said two devices of the power stage, and a coupling between the output electrodes of said one of said two tubes and the other diagonal of the bridge to feed modulated radio-frequency energy by way of said bridge to the input electrodes of said two discharge devices of the power stage co-phasially.

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