

May 18, 1948.

S. DARLINGTON

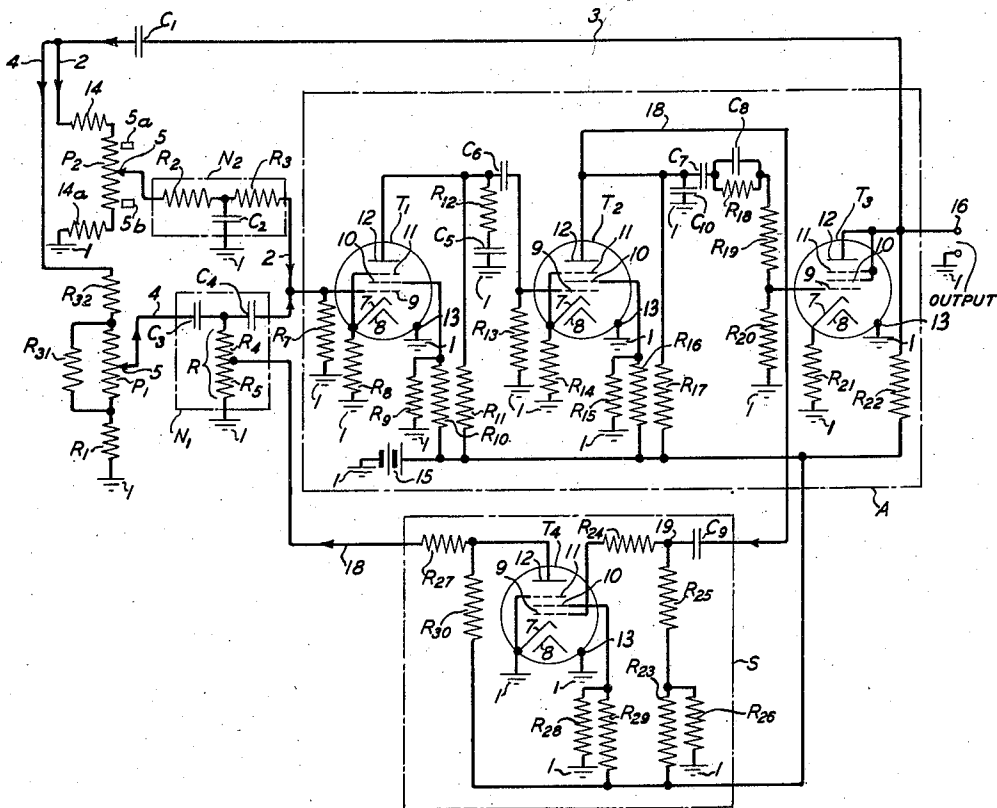
2,441,567

VARIABLE FREQUENCY OSCILLATOR

Filed Feb. 13, 1945

7 Sheets-Sheet 1

FIG. 1



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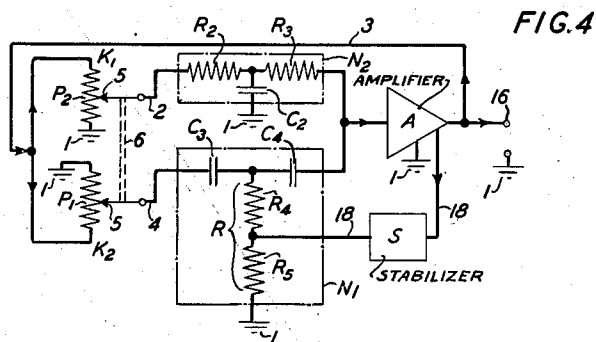
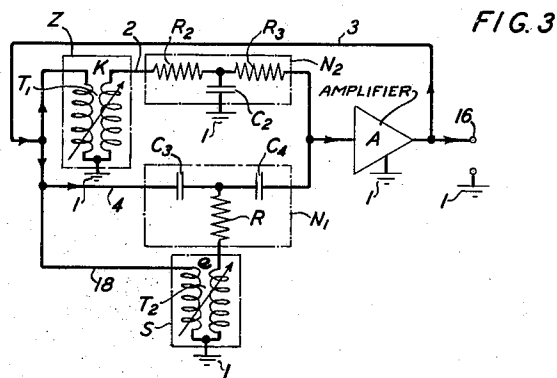
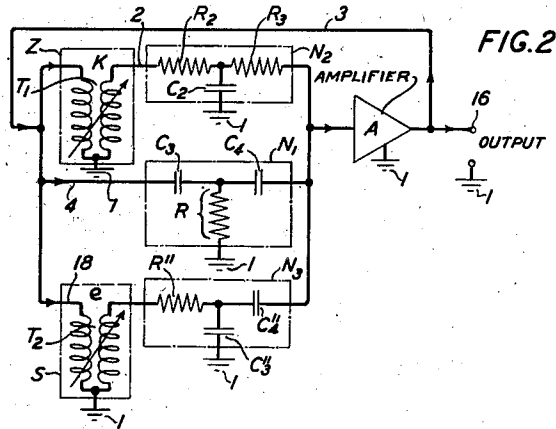
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7 Sheets-Sheet 2



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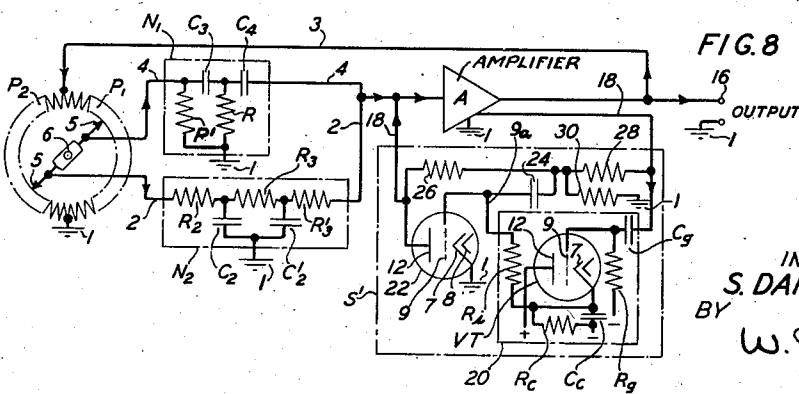
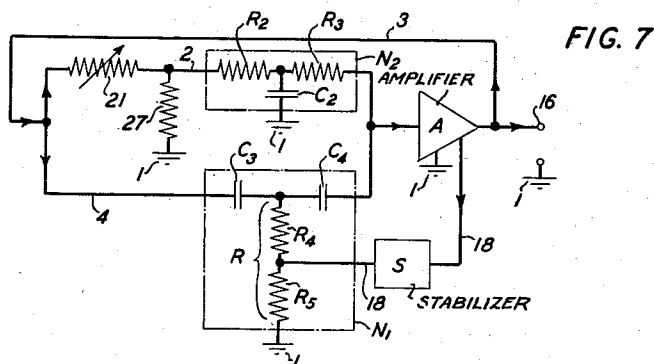
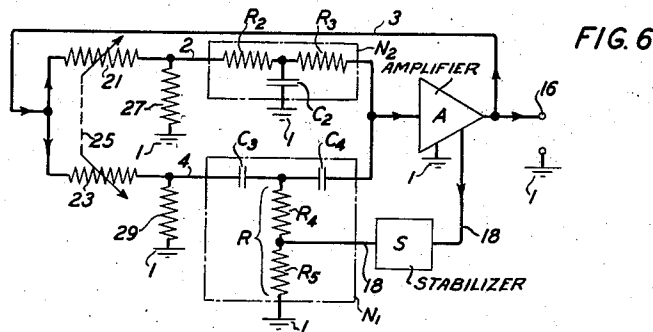
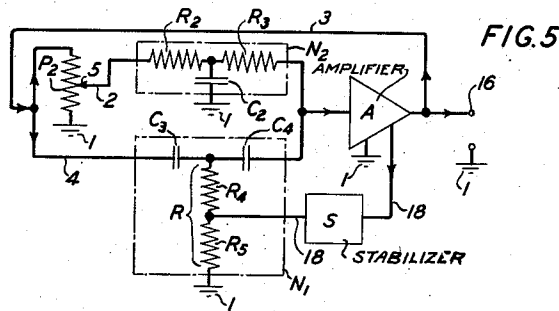
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VARIABLE FREQUENCY OSCILLATOR

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7 Sheets-Sheet 3



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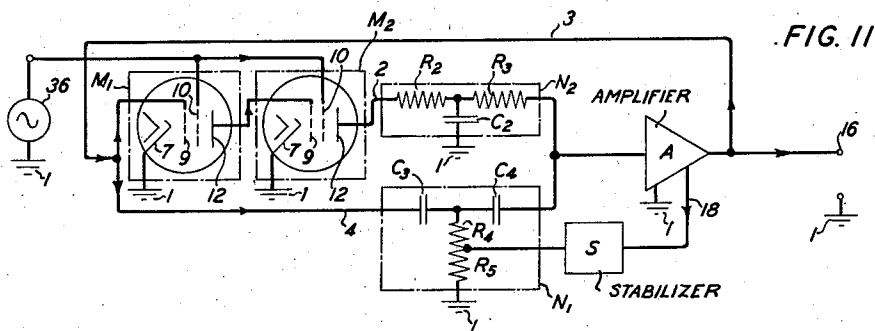
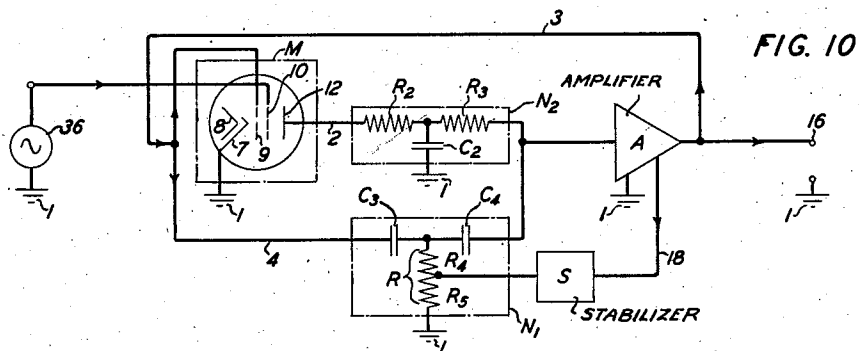
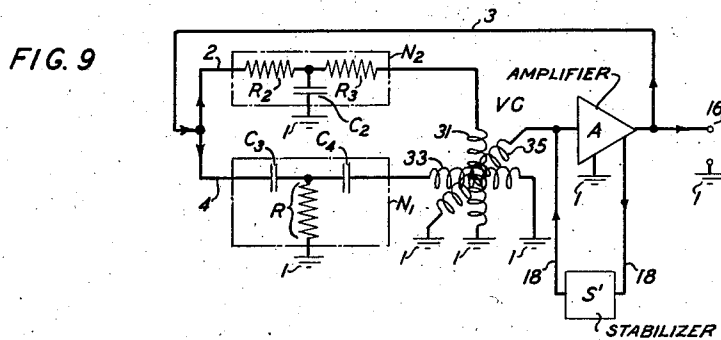
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VARIABLE FREQUENCY OSCILLATOR

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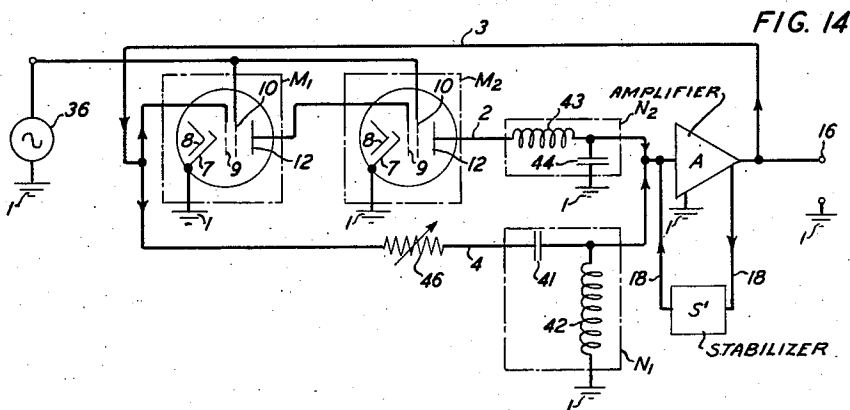
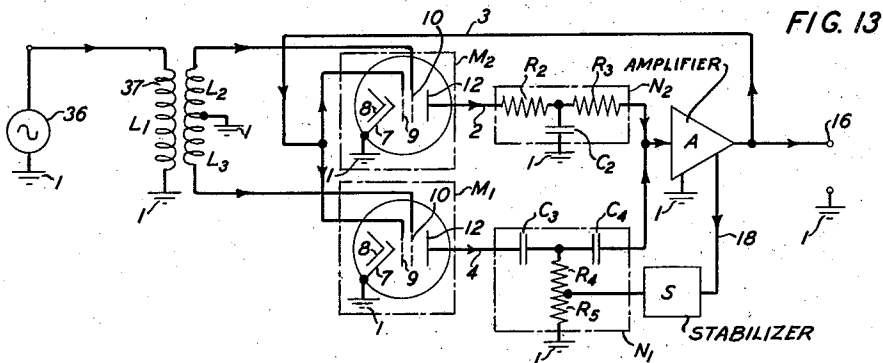
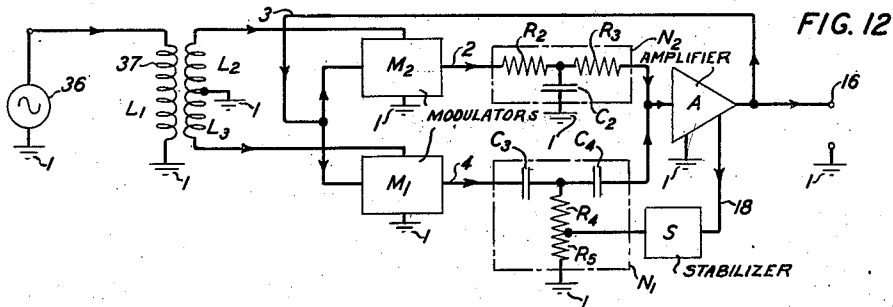
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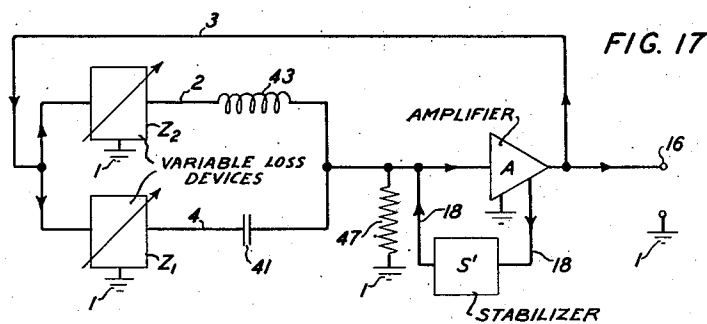
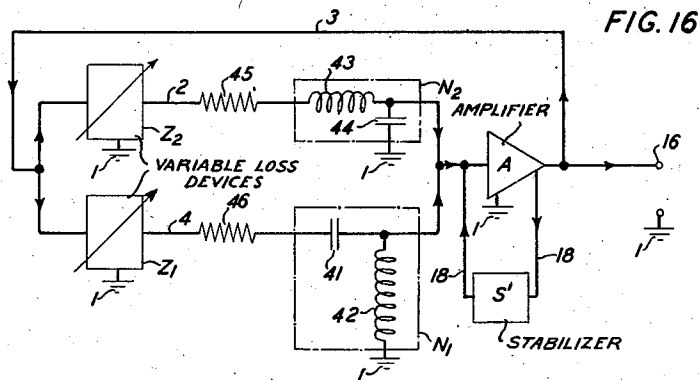
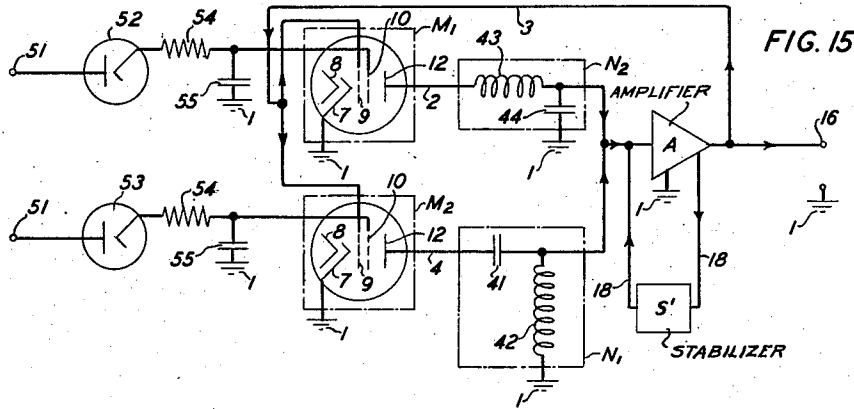
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VARIABLE FREQUENCY OSCILLATOR

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VARIABLE FREQUENCY OSCILLATOR

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7 Sheets-Sheet 7

FIG. 18

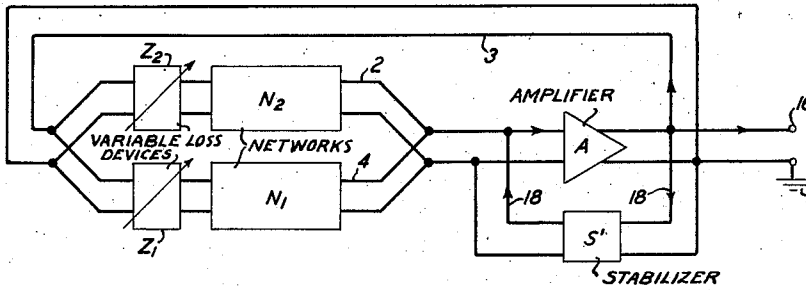


FIG. 19

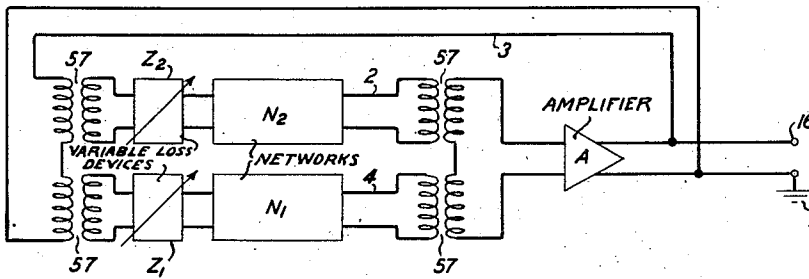
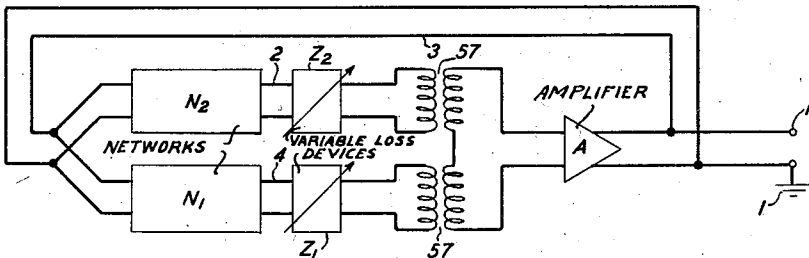


FIG. 20



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UNITED STATES PATENT OFFICE

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VARIABLE FREQUENCY OSCILLATOR

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Application February 13, 1945, Serial No. 577,622

19 Claims. (Cl. 250—36)

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This invention relates to variable frequency oscillators and particularly to means for varying or modulating the frequency of oscillations thereof, and also to means for stabilizing the amplitude level of oscillations over the operating range or band of variable frequencies.

One of the objects of this invention is to provide improvements in varying and modulating the frequency and in controlling the amplitude of oscillations in variable frequency oscillation generators.

Another object of this invention is to improve the operation of variable frequency oscillators in respect to frequency and amplitude stability.

In accordance with one feature of this invention, the frequency of oscillations may be varied or modulated with ease by means of one or more variable loss or variable gain control devices, such as variable potentiometers, resistances, inductances, capacitances, electronic vacuum tubes and similar control devices capable of introducing variable loss or attenuation into a transmission path.

In accordance with another feature of this invention, the amplitude level of oscillations may be automatically stabilized over or throughout a range of operating frequencies by means of non-linear auxiliary feedback systems which may be of the variable loss type.

In accordance with this invention, a separate variable resistance or variable potentiometer constituting a variable loss device, or a variable voltage ratio device which may provide a variable loss may be used to control the frequency of oscillations of the circuit, and may be used in place of the usual variable impedance elements such as the resistance, inductance or capacitance elements comprising the usual frequency determining network of the oscillator or amplifier tube system. Such variable loss control of the frequency may be conveniently used to provide a highly stable and continuously variable oscillator for generating variable frequencies which may be in a band below 100 cycles per second for example, or at any higher frequency. The required variable loss may be conveniently obtained by means of a suitable potentiometer such as a precision tapered potentiometer for example; or the variable loss may be obtained by electronic means or by attenuators of variable resistance. Loss control of the frequency also may be used to provide signal-modulated frequency modulation, where, by using a suitable vacuum tube or varistor modulator to introduce the frequency controlling variable loss, a simple fre-

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quency modulator may be designed for a high percentage modulation.

The variable loss principle may be applied to control not only the frequency of oscillations but also the level amplitude stabilization of oscillations, a level sensitive variable loss device replacing the conventional level sensitive resistance such as, for example, a tungsten or carbon filament lamp, a thermistor, or a gas-filled tube. A simple and effective means of providing level amplitude stabilization may be obtained with an ordinary amplifier vacuum tube using a special connection.

As applied to variable frequency loss-controlled oscillators, the invention may comprise a suitable amplifier or source of gain, a main feedback path having two parallel transmission paths each contributing to the total feedback, networks in the two feedback transmission paths, means for introducing flat loss or gain into one or both of the two feedback transmission paths, and a stabilizing element. In the two feedback transmission paths, the networks may be such that the corresponding contributions to the total feedback have the following properties: (a) The two contributions to the feedback corresponding to the two parallel transmission paths differ in phase by an amount which is substantially 180 degrees independent of frequency; (b) the amplitudes of said two contributions vary differently with frequency, the usual difference being a change in the relative levels of the contributions at a rate which may be substantially equal to 12 decibels per octave of frequency change, the above conditions applying to the relative phases and levels of the two contributions to the feedback and not to the actual phase and level of the total feedback; and (c) the two amplitudes are equal at a definite frequency which is substantially the frequency of oscillation. The means for introducing flat loss or gain into one or both of the two parallel transmission paths referred to does so in such a way that the difference between the amplitude levels of the corresponding contributions to the feedback can be changed at will by amounts which are substantially independent of frequency if levels are measured on a logarithmic or decibel basis, and in such a way as to leave the phases of the two contributions substantially unchanged, the variation in flat loss controlling the variation in frequency of oscillation of the oscillator by varying the frequency at which the amplitudes of the two contributions to the feedback are equal.

The stabilizing element referred to may be a

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level sensitive variable element which may either (a) produce an effective change in one of the networks referred to, or (b) change the feedback in a subsidiary feedback path. The stabilizing element may comprise a subsidiary or auxiliary feedback path yielding a level controlled variable loss for providing the level amplitude stabilization of the oscillations. The source of gain referred to may be provided with stabilizing elements to meet conditions of high feedback amplifier stability.

The loss-controlled variable frequency oscillator circuits particularly illustrated in the figures herein are specific embodiments of a more general class of bridge type feedback oscillator circuits, and feature the use of variable losses or variable voltage ratios for control or modulation of the frequency, as distinguished from the usual variable or adjustable bridge arms in the feedback β -circuit bridge ordinarily used heretofore for control of the frequency. The circuits disclosed in this specification also feature the use of variable losses or variable voltage ratios for level amplitude stabilization of the output voltage of the oscillator circuit. Control of the frequency by means of variable loss devices, and level amplitude stabilization by means of variable loss devices are two independent features, either of which may be used with or without the other.

While the use of variable loss or variable voltage ratio network devices to control the frequency of oscillation is disclosed in connection with particular bridge type oscillators, it will be understood that such network devices may be used for other purposes where it may be desired to introduce high loss to a frequency which can be varied at will by merely changing the loss or voltage ratio by means of variable voltage ratio device such as a variable voltage transformer, a potentiometer, or an amplitude modulator for example. Such a network may also be used as a variable frequency wave trap, or in the β -circuit of a feedback amplifier, or in current analyzers as a means for providing a variable narrow band filter.

For a clearer understanding of the nature of this invention and the additional advantages, features and objects thereof, reference is made to the following description taken in connection with the accompanying drawings, in which like reference characters represent like or similar parts and in which:

Fig. 1 is a circuit diagram illustrating a variable frequency oscillation generator of the bridge stabilized type employing a variable loss type of control device in the form of a variable resistance potentiometer for varying the frequency of oscillations, and also employing a form of variable loss type of auxiliary feedback system for automatically stabilizing the amplitude level of oscillations.

Figs. 2 and 3 are idealized or generalized circuit diagrams illustrating the principle of operation of variable loss and voltage ratio devices for controlling the frequency and the amplitude level of oscillations in variable frequency oscillators.

Figs. 4 to 7 are circuit diagrams illustrating modifications in variable potentiometers and variable resistances for controlling the frequency of oscillations.

Fig. 8 is a circuit diagram similar to the circuit of Fig. 1, but illustrating a modification in the potentiometer controls, and in the amplitude stabilizer control.

Fig. 9 is a circuit diagram illustrating a modifi-

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cation comprising a variometer for controlling the frequency of oscillations.

Figs. 10 to 15 are circuit diagrams illustrating modifications comprising signal-modulated vacuum tubes for controlling the frequency of oscillations.

Figs. 16 and 17 are circuit diagrams illustrating modifications in networks that may be used.

Figs. 18 to 20 are generalized circuit diagrams illustrating modifications in the over-all circuit system, the system in Fig. 18 corresponding to the more specific circuit diagrams of Figs. 1 to 17.

Referring to the drawing, Fig. 1 is a circuit diagram of a variable frequency bridge type oscillation generator provided with variable resistance potentiometers P1 and P2 which are utilized to adjust and control the frequency of oscillation, and also provided with an amplitude level stabilizing circuit 18 which is utilized to stabilize the amplitude of the oscillations generated. As illustrated in Fig. 1, the variable frequency oscillation generator has a bridge network which is of the resistance-capacitance type, and may be utilized for generating the relatively lower values of frequencies over or throughout a relatively wide range of low frequencies by adjustment of the variable resistance potentiometer P2.

As particularly illustrated in Fig. 1, the oscillator may comprise an amplifier or source of gain A consisting of a series of three high gain amplifying vacuum tubes T1, T2 and T3 provided with suitable interstage coupling means therebetween. A main feedback circuit 3 couples the output of the third vacuum tube T3 with the input of the first vacuum tube T1 and comprises a coupling condenser C1, and a pair of parallel-connected transmission paths 4 and 2 which include networks N1 and N2, respectively, and variable resistance potentiometers P1 and P2, respectively, both of which have a resistance component thereof connected in series circuit relation with the networks N1 and N2, respectively. The variable resistance potentiometer P1 is used only as a frequency trimmer, and its loss is thus held within narrow limits by the use of series resistances R1 and R32, and a shunting resistance R31. The main tuning resistance potentiometer P2 and the trimmer potentiometer P1 are connected with the main feedback circuit 3. The other ends of the potentiometer P1 and P2 are connected with the ground 1. The several ground connections 1, which are designated in the drawing by the conventional symbol for a ground connection, may be connected together in practice by means of any suitable wire connectors. The output oscillations may be taken from the output terminal 16 and ground terminal 1 or from points connected therewith.

In practice, it is frequently desirable to use an amplifier A in which the phase between the input and opposite circuits is reversed. Such an amplifier A may be obtained by employing an odd number of amplifying tube stages, such as the three vacuum tube stages illustrated by the direct-coupled three-stage amplifier tubes T1, T2 and T3 in Fig. 1. The tandem connected linear amplifier tubes T1, T2 and T3 of the linear amplifier A, and also the amplitude stabilizer tube T4, may be for example high gain vacuum tubes of the pentode type having an indirectly heated cathode 7, a cathode heater filament 8, an input or control grid 9, a screen grid 10, a suppressor grid 11 connected with the cathode 7, a plate or anode electrode 12, and a metal envelope or tube which may be connected to ground as illustrated

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at 13 in Fig. 1. The cathode heater filaments 8 of the tubes T1, T2, T3 and T4 may be connected in parallel and supplied with heating current in a known manner from any suitable voltage supply source (not shown).

The first tube T1 of the amplifier A may have its cathode electrode 7 and its grid electrode 9 connected to ground 1 through resistances R8 and R7 respectively, and may have its screen grid electrode 10 connected to ground 1 through a resistance R9. Suitable potentials for the screen grid electrode 10 and for the plate electrode 12 of the tube T1 may be supplied through resistances R10 and R11, respectively, from the positive (+) terminal of a suitable power supply source 15. The interstage coupling between the vacuum tubes T1 and T2 may comprise a resistance R12 with a grounded series-connected condenser C5 and a coupling condenser C6. The second vacuum tube T2 of the amplifier A may be provided with a grid resistance R13, a cathode resistance R14 and a screen grid resistance R15. Suitable potentials for the screen grid electrode 10 and for the plate electrode 12 of the second amplifier tube T2 may be supplied through resistances R16 and R17, respectively, from the positive terminal of the power supply source 15. The interstage coupling between the output of the second amplifier tube T2 and the input of the third amplifier tube T3 may include a condenser C7, a parallel-connected condenser C8 and resistance R18, a resistance R19, and a grounded condenser C10, as illustrated in Fig. 1. The third tube T3 of the linear amplifier A may be provided with a grid resistance R20 and a cathode resistance R21. Suitable potentials for the grids 10 and 11 and the plate electrode 12 of the tube T3 may be supplied through resistance R22 from the positive terminal of the power supply source 15.

The network N2 as illustrated by the block diagram labeled N2 in Fig. 1, is constructed in the form of a T-network consisting of two series connected resistances R2 and R3 and a single shunt condenser C2 which is connected at one end thereof to the ground 1. The network N1 as illustrated by the block diagram labeled N1 in Fig. 1 is constructed in the form of a T-network consisting of two series condensers C3 and C4 and a shunt resistance R which is composed of two series resistances R4 and R5, the latter being connected at one end thereof to the ground 1. The networks N1 and N2 taken together comprise a double-T or parallel-T network system and taken with the variable resistance potentiometers P1 and P2 form two parallel paths 4 and 2, respectively, in the main feedback circuit 3 of the linear amplifier A, the main feedback circuit 3 being connected between the output circuit 16 of the third amplifier tube T3 and the input of the first amplifier tube T1 of the amplifier A. The output circuits of the parallel-connected paths 4 and 2 and of the networks N1 and N2 are connected with the input of the first amplifier tube T1, and therebetween the resistance R7 is connected to ground 1 at one end thereof.

The bridge type feedback circuit networks N1 and N2 may be practically any type of network system which provides a substantially constant 180-degree phase difference and different amplitude characteristics for varying frequency values. Since there are a large number of bridge circuits which provide such transmission characteristics, the network system N1 and N2 may consist of any of a large number of possible bridge

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circuit arrangements. The particular parallel-T networks N1 and N2 when composed of resistances and capacitances only as illustrated in Fig. 1, are particularly useful at the audio and lower frequencies. For the higher frequencies, inductances instead of resistances may be used in the networks N1 and N2.

In the circuit of Fig. 1, the frequency of oscillation depends intimately upon the resistances that are used in the networks N1 and N2 and hence they are constructed as precision units where precision is desired. At low frequencies, such as, for example, frequencies below 100 cycles per second, the resistance units of the networks N1 and N2 may be made large in order to avoid the use of unduly large precision capacitance devices. Since continuously variable precision resistance units are difficult to realize in practice in large sizes, the large precision resistances of the networks N1 and N2 may be fixed or non-variable resistances, the frequency of oscillation being adjusted or varied by one of the variable potentiometers P1 or P2 which may be made to be of relatively low resistance values and of relatively small size. In the circuit of Fig. 1, the variable resistance potentiometer P2 represents a variable loss device which is used to furnish variable control over the frequency of the system. Since high precision in frequency imposes high precision on the resistances and other component parts of the networks N1 and N2, where a variable frequency is required, it is often difficult or impractical to obtain high precision in frequency control by means of adjustments in the large precision resistances and other component parts of the networks N1 and N2 themselves. For this reason, and for the same precision in frequency, the use of the potentiometer type of frequency control P2 is more easily realized in practice. The impedance of the potentiometer P2 being of relative low value is more easily constructed in adjustable form. Moreover, the potentiometer voltage ratios of the adjustable potentiometer P2 can be normally held to close limits since the voltage ratio thereof depends upon the relative resistances rather than upon the absolute values of resistance. Accordingly, changes in the absolute values of resistance of the potentiometer P2, as caused by changes in temperature, for example, do not change the voltage ratio of the potentiometer P2 except to the extent that such changes in absolute resistance may not be uniform along the potentiometer P2.

While the potentiometer P2 presents a finite impedance to the load connected to the brush 5 thereof, its effect on the frequency stability of the system may be minimized by making the impedance of the resistance potentiometer P2 reasonably small relative to the impedance value of the resistance R2, to which the brush 5 of the potentiometer P2 is connected. The effect is to add resistance to the resistance R2 by the parallel combination of the segments of the potentiometer P2 above and below the brush 5 thereof. To the extent that the effective addition of such resistance to the resistance R2 is constant, it may be compensated for by a reduction in the resistance of R2 itself. The remaining variations in the equivalent resistance of R2 as a result of the effect of the resistance of the potentiometer P2 thereon may be taken care of in part by means of the amplitude stabilizer system such as, for example, by the system comprising the auxiliary feedback circuit 13 including the stabilizer S with the tube T4 of Fig. 1, and in part

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by proper calibration of the voltage ratio provided by the resistance arms of the potentiometer P2. If desired, the taper of the resistance values of the potentiometer P2 may be used to obtain a linear variation of frequency with respect to the shaft position of the potentiometer P2. It will be understood, however, that the adverse effect of the variable resistance of the potentiometer P2 on the over-all stability of the oscillator system is not large when the impedance of the potentiometer P2 is reasonably small relative to the impedance value of resistance R2 of network N2.

As illustrated in Fig. 1, the tuning potentiometer P2 has a resistance composed of a middle portion connected with the adjustable wiper 5, and two end series resistance portions 14 and 14a. This construction, when used, insures that the potentiometer P2 cannot be set to a value corresponding to a frequency outside the range for which the circuit is designed. This is a desirable feature because the average $\mu\beta$ of the system ordinarily must or should be held within plus or minus 90 degrees of a negative real over the entire working range of frequencies. In providing the potentiometer P2 with suitable non-adjustable end resistance portions 14 and 14a as illustrated schematically in Fig. 1, the possibility of setting the potentiometer P2 outside the working range of frequencies may be prevented.

As illustrated in Fig. 1, an auxiliary feedback circuit 18 may be provided between the plate output circuit 12 of the second amplifier tube T2 and the input circuit of the first amplifier tube T1 of the amplifier A in order to stabilize the amplitude level of oscillations generated. As shown in Fig. 1, the anode 12 of the second amplifier tube T2 of the linear amplifier A is connected through a blocking condenser C3 and a resistance R24 with the input grid 9 of the pentode T4, and the output or plate electrode 12 of the tube T4 is connected through resistance R27, resistance R4, and condenser C4 with the input grid electrode 9 of the first amplifier tube T1 of the amplifier A. The input grid electrode 9 of the tube T4 is connected through resistances R23, R24 and R25 with the positive (+) terminal of the supply source 15. A resistance R26 is connected between ground 1 and a point intermediate the resistances R23 and R25. The middle grid electrode 19 of the tube T4 is connected through a resistance R28 to ground 1, and through a resistance R29 to the positive terminal of the voltage supply source 15. Potential for the plate electrode 12 of the tube T4 is supplied through a resistance R30 which is connected with the positive terminal of the power supply source 15.

In the vacuum tube stabilizer system S of Fig. 1, positive bias voltage from the supply source 15 and signal voltage oscillations from the output of the second amplifier tube T2 of the amplifier A are combined at the junction 19, and are then fed to the control grid 9 of the tube T4 through a large resistance R24. The voltage at the junction 19 is a resultant of the oscillating signal voltage applied thereto and the positive bias voltage applied thereto from the supply source 15, the latter being supplied through resistances R25 and R23. Whenever the instantaneous value of the voltage at the junction 19 is positive, the grid electrode 9 of the tube T4 draws current and produces a substantial voltage drop across the input resistance R24. Because the grid current and also the voltage drop across the resistance

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R24 increases very rapidly with increasing voltage at the grid 9 of the tube T4, the grid voltage at grid 9 of tube T4 changes very little during wide fluctuations of the voltage at the junction 19. As a result, the oscillating signal applied to the tube T4 is effectively blocked so long as the combination signal voltage and positive bias voltage at the junction 19 is positive over the entire cycle, which is so long as the amplitude of the signal voltage applied to the junction 19 is less than the positive bias voltage at the junction point 19, multiplied by the ratio of $R24$ to $R24 + R25 + R23$. When the amplitude of the signal voltage at the junction 19 begins to exceed this value of the bias voltage, the voltage of the grid 9 of the stabilizer tube T4 swings negative over a portion of each cycle. While such grid voltage is negative, the grid 9 of the tube T4 draws no appreciable current and there is no appreciable corresponding drop across the grid resistance R24. Hence, only portions of each cycle of the combination signal and bias voltage at the junction 19 are transmitted and amplified by the tube T4. The output of the tube T4 is fed into the network N1 of the oscillator through the resistance R4 and opposes the oscillating signal voltage therein thus giving the desired level amplitude stabilization at a level at which the amplitude of the signal voltage at the junction 19 is slightly greater than the bias voltage at the junction 19.

The positive grid vacuum tube T4 of Fig. 1 accordingly functions to supply pulses of alternating voltage in opposition to the signal voltage in the network N1 thereby to provide level stabilization of the amplitude of oscillations. At the lower amplitude levels of the output voltage of the amplifier A, the vacuum tube T4 due to the steady positive grid bias applied to the grid electrode 9 thereof from the supply source 15 effectively blocks the oscillations reaching it from the output of the second amplifier tube T2 over the circuit 18 and continues to do so until the amplitude of the oscillations reaches a critical level in amplitude. As the level increases further in amplitude, the tube T4 begins to transmit the oscillations to an extent that increases as the level of the amplitude of oscillations increases above the critical value mentioned. Accordingly, the effect on the circuit 18 of the tube T4 is to prevent transmission of oscillations at the low levels of amplitude and to transmit oscillations with the increasingly higher values of amplitude applied thereto.

The tube T4 is fed from the output of the second amplifier tube T2 or from the next to the last tube of the gain circuit A, rather than from the output 16 of the last amplifier tube T3 of the gain circuit A, in order to balance against the reversal through the stabilizer tube T4, the similar reversal in direction of signal that occurs in passing through the last tube T3 of the gain circuit A. A vacuum tube, such as the tubes T3 and T4, reverses the direction of the signal which it transmits. Hence the reversal through the stabilizer tube T4 is balanced by the similar reversal through the last tube T3 of the gain circuit A. The tube T4 fixes the amplitude level of oscillations at a value which is substantially independent of the frequency of oscillations, and accordingly a substantially constant oscillator output voltage level is obtained from the output 16 of the last amplifier tube T3 of the gain circuit A provided the tube T3 has a substantially constant gain over the working range of variable operating frequencies.

The amplitude stabilizer system 18 described in connection with the positive grid tube T4 of Fig. 1 has a number of advantages. It is fast acting and capable of following rapid changes of frequency due to changes in the setting of the frequency controlling tuning potentiometer P2 even though these changes of frequency setting of the potentiometer P2 call for substantial changes through the tube T4. It is a sensitive system giving a sharp rise in transmission at a definite amplitude level, and the amplifier tube T4 may be of the same type as the amplifier tubes T1, T2 and T3 of the μ circuit A of the oscillator. While considerable harmonics may be generated relative to the fundamental component transmitted by the tube T4, the amplitude of the harmonics and also of the fundamental itself that enters the β circuit over the circuit 18 is relatively small as compared with the amplitude of the oscillator signal voltage that is fed back over the main feedback circuit 3 from the output 16 of the last amplifier tube T3.

In the amplitude level stabilizer system 18 illustrated in Fig. 1, a non-linear feedback voltage e is produced which varies with the amplitude of the output voltage of the amplifier A and which operates to increase the damping of oscillations generated with increasing amplitude of output voltage. The feedback in the auxiliary feedback circuit 18 is in the same phase as that of the main feedback circuit 3 and increases with the level of the output voltage which gives satisfactory amplitude stability at low and high frequency cut-offs, is comparatively simple and operates well even at low levels of voltage applied thereto. This results from the feedback voltage from the output circuit of the oscillator being fed back through the vacuum tube T4 which has a substantial plus (+) bias on the control grid electrode 9 thereof. Where the positive or plus (+) bias and the signal are fed to the grid 9 of the tube T4 through the high resistances R23, R24 and R25, the low impedance of the grid 9 at plus (+) voltages operates to prevent any substantial signal voltage being fed back until the magnitude of the signal voltage becomes large enough to swing the grid 9 negative, above which critical level there will be a sudden increase in transmission level through the vacuum tube T4. Thus the effective stabilizing voltage e is very small in magnitude until the critical level referred to is reached and thereafter it rises very rapidly with further increasing levels of oscillating signal voltage applied to the tube T4. A substantial value of positive (+) voltage bias may be applied from the supply source 15 to the high voltage side of the grid resistances R23, R24 and R25, without getting excessive positive (+) voltages on the grid 9 itself of the tube T4 as a result of the IR voltage drop that occurs in such grid resistances when the grid 9 of the tube T4 draws current.

It will be noted that since the stabilizing tube T4 gives a phase reversal, it is fed from the main amplifier A from a different interstage point, and preferably from the first interstage back of the final output tube T3, as illustrated in Fig. 1. It will be noted that variations in the gain μ of the high gain linear amplifier A are absorbed by the non-linear stabilizer circuit 18 comprising the positive grid tube T4 without substantially any shifting of the frequency of oscillation provided the phase shift of the bridge network system N1 and N2 is relatively small. Provided that the phase shift of the bridge network N1 and N2

referred to can be kept small enough, a rather small value of gain μ may be used if the precision required is not too high.

As an illustrative example, a particular oscillator constructed in accordance with the circuit of Fig. 1 and having a variable frequency ranging from 30 to 90 cycles per second was provided with component resistances and condensers having substantially the following values. In the particular example mentioned, the values for the component resistances expressed in ohms were about: R1=65,000, R2=200,000, R3=206,000, R4=85,000, R5=12,000, R7=1,000,000, R8=2,000, R9=15,000, R10=100,000, R11=250,000, R12=25,000, R13=1,000,000, R14=2,000, R15=15,000, R16=100,000, R17=250,000, R18=10,000,000, R19=500,000, R20=500,000, R21=800, R22=10,000, R23=200,000, R24=1,000,000, R25=1,000,000, R26=30,000, R27=1,000,000, R28=15,000, R29=100,000, R30=1,000,000, R31=500, R32=720, P1=2,000, and P2=10,000 ohms total, the values of the end resistances being 750 ohms each and the value of the middle resistance covered by the wiper 5 being 8500 ohms. The values for the component condensers expressed in microfarads were about: C1=1.0, C2=.016, C3=.008, C4=.008, C5=.005, C6=2.0, C7=1.0, C8=.012, C9=.1, C10=.005; and the tubes T1, T2, T3 and T4 were standard 12SJ7 pentodes having their cathode filaments energized from a cathode heating source of about 12.6 volts and provided with a power supply source 15 of about +250 volts direct current potential.

The bridge networks N1 and N2 of Fig. 1 are represented as being composed of two three-terminal networks N1 and N2 connected in parallel. The short-circuit transfer admittances of the networks N2 and N1 may be represented by Y'12 and Y''12, respectively. The short-circuit transfer admittance Y12 is a function of frequency associated with four terminal networks in general, is frequently used in general network theory, and is defined as the ratio of the current I in a short-circuit across the output terminals of the network to the voltage E across the input terminals which produces it. In accordance with a well-known theorem of general network theory, at any frequency of infinite loss such as $\beta=0$, the transfer admittance of the two parallel-connected networks will satisfy the condition:

$$Y'12 + Y''12 = 0 \quad (1)$$

In terms of the reference notations of Fig. 1, the short-circuit transfer admittances Y'12 and Y''12 of the networks N2 and N1 respectively are:

$$Y'12 = \frac{1}{\frac{R2+R3}{R2R3} + 1 + \frac{R2+R3}{R2R3} \times C2p}$$

and

$$Y''12 = \frac{RC3C4p^2}{1 + R(C3+C4)p} \quad (2)$$

where $p = i\omega$ where ω is the frequency in radians per second.

Assuming that the values of the component resistances and capacitances of the networks N1 and N2 are made to satisfy the following restriction:

$$\frac{R2R3}{R2+R3} C2 = R(C3+C4) \quad (3)$$

and assuming that T is used to represent either of these equated quantities of Equation 3, now

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assumed to be equal, the sum of the short-circuit transfer admittances $Y'_{12} + Y''_{12}$ may be written as follows:

$$Y'_{12} + Y''_{12} = \frac{\frac{1}{R_2 + R_3} + RC_3C_4p^2}{1 + Tp} \quad (4)$$

Equating this last equation 4 to zero in order to locate frequencies of $\beta=0$ in accordance with Equation 1 and regarding $p^2 = -\omega^2$ at real frequencies, then

$$\beta=0 \text{ at } \omega^2 = \frac{1}{R(R_2 + R_3) C_3C_4} \quad (5)$$

provided Equation 3 is satisfied.

If a voltage transforming device such as, for example, the resistance potentiometer P2 of Fig. 1 or the transformer T1 of Fig. 2 is introduced at one end of the network N2 without changing the network N1, the short-circuit transfer admittance Y''_{12} of network N1 is unchanged and the short-circuit transfer admittance Y'_{12} of network N2 is modified through multiplying it by the voltage ratio K which in the case of a transformer T1 of Fig. 2, for example, may be the ratio of the secondary voltage to the primary voltage. If the short-circuit transfer admittance Y'_{12} of the network N2 as given in Equation 2 be multiplied by the voltage ratio K, the sum of $Y'_{12} + Y''_{12}$ of Equation 4 is changed to

$$Y'_{12} + Y''_{12} = \frac{\frac{K}{R_2 + R_3} + RC_3C_4p^2}{1 + Tp} \quad (6)$$

Equating Equation 6 to zero gives:

$$\beta=0 \text{ at } \omega^2 = \frac{K}{R(R_2 + R_3) C_3C_4} \quad (7)$$

In accordance with the above equations, a variable frequency of $\beta=0$ may be obtained by introducing a variable voltage ratio K from a variable voltage device P2, T1, etc., and without varying any of the resistances or capacitances of the networks N1 and N2. This accordingly meets the requirements of a variable frequency bridge oscillator provided a suitable amplitude level sensitive stabilizing device, such as the auxiliary feedback circuit 13, is introduced in order to produce the proper deviations from Equation 7 at other than the level of stable oscillations. Since Equation 7 depends on Equation 3, a level sensitive element may be used which upsets Equation 3 except at the level of stable oscillations.

Fig. 2 is a circuit diagram illustrating the principle of operation of loss-controlled variable frequency oscillators. The circuit of Fig. 2 is similar to Fig. 1 but illustrates the frequency controlling variable loss Z in the form of an idealized variable voltage ratio transformer T1 in place of the resistance potentiometer P2 of Fig. 1, the transformer T1 being employed to illustrate the principle of variable loss voltage ratio control of the frequency of oscillations. Fig. 2 also illustrates as an alternative arrangement to the stabilizer S of Fig. 1, a level controlling stabilizer S in the form of an idealized variable voltage ratio transformer T2 connected with a network N3 which may be used to illustrate the principle of operation of a level sensitive stabilizer for controlling the amplitude of the output oscillations. The feedback or β circuit in Fig. 2 employs the additional network N3 which is connected in parallel circuit relation with the parallel-connected networks N1 and N2.

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The resistance R' and the condensers C'3 and C'4 comprise the network N3 and may be of the same size as those of the network N1 but they are arranged differently as to which element is connected to ground 1 and which is connected to the input. This makes it possible to use the same type of component elements for both of the networks N1 and N3. The variable voltage ratio e of the variable transformer T2, which is connected with the network N3, is relatively small and the transformer T2 represents means for introducing variable voltage into the circuit comprising the network N3.

The frequencies for $\beta=0$ then correspond to:

$$Y'_{12} + Y''_{12} + Y'''_{12} = 0$$

where Y'''_{12} is the short-circuit transfer admittance of the additional network N3, and Y'_{12} and Y''_{12} are, as before, the short-circuit transfer admittances of the networks N2 and N1 respectively. Satisfying Equation 3 relating the component elements of network N2 to those of networks N1 and N3, the short-circuit transfer admittances of such networks are:

$$Y'_{12} = \frac{\frac{K}{R_2 + R_3}}{1 + Tp}$$

$$Y''_{12} = \frac{RC_3C_4p^2}{1 + Tp}$$

$$Y'''_{12} = \frac{eC_4'p}{1 + Tp}$$

wherein K is the voltage ratio of the transformer T1, e is the voltage ratio of the transformer T2, and T represents either of the quantities equated in Equation 3.

Adding the three values of the preceding equation, solving for the roots, and assuming that the voltage ratio e of transformer T2 is small, gives the complex frequency of $\beta=0$ as:

$$\frac{e}{2RC_3}$$

In accordance with the above equations the β circuit of Fig. 2 when restricted by Equation 3, has non-minimum phase when the voltage ratio e of the amplitude level controlling voltage ratio device T2 is negative and has minimum phase when it is positive. Hence, it is suitable for a bridge oscillator provided the voltage ratio e depends on voltage level or amplitude in such a way that it is negative at the low levels and positive at high levels.

The use of both positive and negative values of the stabilizing voltage ratio e is inconvenient when it comes to practical embodiments of the variable loss method of level sensitive amplitude stabilization of voltage. This situation may be avoided, however, by departing slightly from the restriction of Equation 3 and using a voltage ratio e which is comparatively small. If R is fixed at a value either somewhat too small or too large the voltage ratio e may always be either positive or negative respectively. In either case, however, the voltage ratio e must change with increasing level in the direction of increasing positive value of voltage ratio e. When using the practical type of stabilizer S as indicated by the tube T4 in Fig. 1, the voltage values should be chosen in such a way that the equivalent values of resistance R2, as determined by the effect of the impedance of the potentiometer P2 thereon, always correspond to departures from Equation 3 in the right direction in order to ob-

tain the desired effect which is the direction calling for positive values of feedback stabilizer voltage e in the circuit of Figs. 2 and 3.

Fig. 3 is a circuit diagram of a loss-controlled variable frequency oscillator similar to that of Fig. 2 but illustrating a modification in which the network N3 of Fig. 3 has been omitted and the variable ratio transformer T2 has been connected with the network N1 to illustrate the principle of providing level amplitude stabilization embodied in the more practical feedback stabilization circuit 18 of Fig. 1. The transition from Fig. 2 to Fig. 3 is obvious if the plate-to-ground and the grid-to-ground terminating impedances of the β -circuit networks be regarded as both zero. For, in Fig. 2 the feed through the level controlling variable loss S is through the resistance R'' to a point at which the condensers C''3 and C''4 are commonly connected, the condenser C''3 being connected to ground 1 and the condenser C''4 being connected to the grid of the amplifier A; while in Fig. 3 the feed through this variable loss S is through the resistance R to the point at which the condensers C3 and C4 are commonly connected, the condenser C3 being connected to the plate of the amplifier A and the condenser C4 being connected to the input grid of the amplifier A. The effect will be the same in the two cases of Figs. 2 and 3 if the plate-to-ground impedance is regarded as substantially zero and if the impedance to ground at the output of the level control variable loss S in Fig. 3 is also regarded as substantially zero, so that transmission through the middle network N1 is not adversely affected. In practice, the grid-to-ground impedance is relatively unimportant since the circuit is operated at or very close to balance. While the plate-to-ground impedance is not zero, it is generally very small compared to the network impedances of the β circuit. For example, in the practical example given hereinbefore in connection with Fig. 1, the plate-to-ground impedance is made less than the resistance of the potentiometer P2 which in the example given hereinbefore is 10,000 ohms, while the network impedance is of the order of 300,000 ohms, so that the plate-to-ground impedance, though not zero, is very small compared to the network impedance and the result in practice is not greatly different from that for the assumed zero plate-to-ground impedance.

The frequency stability of the steady state oscillations depends upon the amplitude level V and the real frequency ω of the steady state oscillations corresponding to the conditions of $\mu\beta=+1$ and the damping constant $\alpha=0$. One method of realizing frequency stability in an oscillator is to make the gain μ of the amplifier A large and the loss β of the feedback circuit small at the oscillation frequency. The distinguishing feature of a stabilized bridge oscillator is a β circuit such that $\beta=0$ at some real frequency $\omega=\omega_0$, when the amplitude level sensitive element S has a particular value corresponding to a particular amplitude level of $V=V_0$. The β circuit is so arranged that it is a minimum phase network when V is greater than V_0 , but is a non-minimum phase network when V is less than V_0 . The corresponding μ 70 or gain circuit is such that μ is large at frequency $\omega=\omega_0$, and such that oscillations build up when the amplitude level V is less than V_0 and damp down when V is greater than V_0 . For practical purposes, this amounts to designing the μ cir-

cuit as for a feedback amplifier in which the properties assigned to the β circuit are those assumed by the oscillator β circuit under the minimum phase condition of V greater than V_0 . Any circuit with these properties is in effect a stabilized bridge oscillator.

A large gain μ leads to a small frequency difference $\omega-\omega_0$ and also to a small amplitude level difference $V-V_0$. Accordingly, the frequency stability is related to the amplitude level and may be controlled by variation of the β circuit with amplitude level variation. The means used to obtain the required variation of β with change in the amplitude level may be any suitable level controlled stabilizing means such as the auxiliary feedback systems S which are particularly described herein in connection with bridge type oscillators although such feedback stabilizing systems S may also be used with other types of oscillators. A specific stabilizing element can usually be used at any of a multiplicity of points in a β circuit configuration which is otherwise prescribed. If any point is found such that $\beta=0$ at frequency $\omega=\omega_0$ when $V=V_0$, the β circuit can be expected to make the transition from minimum phase to non-minimum phase condition as V passes through V_0 , provided the variation is made to be in the right direction, namely that the phase is minimum when V is greater than V_0 , rather than the reverse.

The relation between μ , β and V and the control of the phase of

$$\frac{\partial \beta}{\mu \partial V}$$

is considered as follows. If stabilized oscillators are considered in general, the methods of obtaining frequency stability depend on the equation:

$$p-i\omega_0=0+1(\omega-\omega_0)=\frac{1}{\mu}-\frac{\partial \beta}{\partial V}\delta V$$

where δV is determined by $\alpha=0$, and the derivative

$$\frac{\partial \beta}{\partial V}$$

and

$$\frac{\partial \beta}{\partial p}$$

are evaluated as for $\omega=\omega_0$. The use of this equation depends on the observation that in practice β is very small at the frequency of oscillation. The value of β depends in general on the complex frequency $p=i\omega$, and on the amplitude level V of the generated signal, that is:

$$\beta=\beta(p, V)$$

where V the amplitude level is a real number. Let the β circuit be such that $\beta=0$ when the real frequency $\omega=\omega_0$ and $V=V_0$. Then for small departures from this condition, β may be in Taylor's series, and only the first order terms are 65 important:

$$\beta=\frac{\partial \beta}{\partial p}\delta p+\frac{\partial \beta}{\partial V}\delta V$$

and accordingly:

$$\delta p=\frac{\beta-\frac{\partial \beta}{\partial V}\delta V}{\frac{\partial \beta}{\partial p}}$$

Since in practice the oscillator operates at the frequency for which $\mu\beta=1$ and since this must

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be a real frequency, it is necessary that the action of the stabilizer shall make $\alpha=0$, and this determines δV . Hence $\delta p=i(\omega-\omega_0)$, and this with $\mu\beta=+1$ gives

$$i(\omega-\omega_0)=\mu\frac{\frac{1}{\delta V}\frac{\partial\beta}{\partial p}\delta V}{\frac{\partial\beta}{\partial p}}$$

as the change in frequency from the nominal ($\beta=0$) to the actual ($\mu\beta=1$) operating point. Maximum frequency stability requires that $\omega-\omega_0$ shall be zero, and the closer to zero that it can be brought, the better the oscillator that will result. The expression above may be made zero, or made small, under the following conditions:

(1) The expression may be made zero by making

$$\frac{1}{\mu}=\frac{\partial\beta}{\partial V}\delta V$$

Since V , and hence δV , is a real number, this requires that

$$\frac{\partial\beta}{\mu\partial V}$$

shall be a negative real number.

(2) The expression may be made very small by making

$$\frac{\partial\beta}{\partial p}$$

very large. This is what is done in a crystal oscillator, or by making the reactance-resistance ratio Q of the β circuit high in an inductance-capacitance type oscillator. In a variable frequency oscillator it is difficult to obtain a sufficiently high Q to make this above an adequate means of obtaining adequate frequency stability, though in a fixed frequency oscillator it may be sufficient. In the resistance-capacitance type of oscillator, the value of

$$\frac{\partial\beta}{\partial p}$$

is strictly limited in magnitude.

(3) The expression may be made small by making μ large, so that a given percentage change in

$$\frac{1}{\mu}$$

will be a small absolute change. But this alone is not adequate. Since $\mu\beta=1$, a large μ requires a small β , and if the increase in μ is accompanied by a decrease in

$$\frac{\partial\beta}{\partial p}$$

obviously not much is gained. But if β is made up of contributions from two parallel paths as occurs in bridge type oscillators, it is possible to have μ very large and β very small while

$$\frac{\partial\beta}{\partial p}$$

is maintained fairly large.

In the oscillators here considered, μ is thus made large and

$$\frac{\partial\beta}{\partial p}$$

is kept from becoming small by using a bridge circuit as noted above, thus satisfying condition (3) without violation of condition (2). As in the

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oscillator described in connection with Fig. 1, it is found that the phase of

$$\frac{\partial\beta}{\mu\partial V}$$

is then identical with the average phase of $\mu\beta$, as this must be kept within ± 90 degrees of a negative real number if the oscillator is to have a single mode of oscillation at any setting of the potentiometer P2 of Fig. 1. This accordingly approaches a close realization of condition (1) above, and is thus a means of improving the frequency stability in view of the simultaneous satisfaction of conditions (2) and (3) mentioned above. The condition (1) would not be satisfied if the usual resistance stabilizer were used. But the use of the loss-controlled type of stabilizer circuit 18 thus permits better frequency stability than that of a voltage variable resistance stabilizer of the usual type. The matter of frequency stability may be carried further by either of the following methods:

(a) Design of the circuit so that the average $\mu\beta$ over the whole working range of frequencies is very close to a negative real number. Identity of phase of

$$\frac{\partial\beta}{\mu\partial V}$$

with average $\mu\beta$, as obtained by the loss-controlled stabilizer circuit 18, then substantially satisfies condition (1) above simultaneously with condition (3) and also with condition (2) if sufficiently high Q networks are used.

(b) Variation of the phase of μ as the value of the voltage ratio K of the potentiometer P2 is changed, as by simultaneous control of a phase shifting network in the μ circuit in order to keep $\mu\beta$ equal to a negative real number, or nearly so. Accordingly, a stabilizer system may be utilized in which

$$\frac{\partial\beta}{\mu\partial V}$$

satisfies condition (1) while conditions (2) and (3) may also be independently satisfied, and this result may be obtained independently of the means for controlling the frequency. In accordance with condition (1), the phase of the correction introduced by operation of the voltage stabilizer is, when added to the phase of μ which is the amplifier part A of the oscillator, equal or nearly equal to 180 degrees, for maximum frequency stability. The circuit illustrated in Fig. 1 meets this requirement near the middle of the frequency range where average $\mu\beta$ is approximately -1 and accordingly improved performance is obtained in this range. To illustrate this properly, stops 5a and 5b are shown in Fig. 1 to confine the movement of the adjustable brush 5 of the potentiometer P2 to this region. This result may also be obtained by designing the circuit so that the phase of average $\mu\beta$ over the working range is very close to 180 degrees.

Fig. 4 is a simplified circuit diagram of the loss-controlled variable frequency oscillation generator of Fig. 1, but illustrates a modification in which the two resistance potentiometers P1 and P2 may have both wipers or brushes 5 on the same control shaft 6, as in Fig. 8, so that both of the potentiometers P1 and P2 may be operated simultaneously from a single control shaft 6 in order to control the frequency of oscillation and to balance out the effects of the finite impedances of the potentiometers P1 and P2 on the required

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amplitude stabilization thereby to reduce the amount of amplitude stabilization required by the stabilizer S. The finite impedances of the potentiometers P1 and P2 balance out as regards the real condition for $\beta=0$ at a real frequency corresponding to Equation (3) when $K2=1-K1$ and when $Y''11=Y'11$ where $Y'11$ and $Y''11$ are as before, the short-circuit driving point admittances of the networks N2 and N1 respectively. The two potentiometers P1 and P2 so arranged vary more slowly with frequency and permit a wider spread of frequencies with practical forms of the potentiometers P1 and P2.

Fig. 5 is a circuit diagram of a loss-controlled variable frequency oscillation generator similar to those of Figs. 1 and 4 but illustrating a modification in which a single resistance potentiometer P2 is used to variably control the frequency of the oscillator circuit. While in Fig. 5, the potentiometer P2 is shown connected in circuit 2 with the network N2, alternatively it may be removed and similarly connected in circuit 4 with the other network N1.

Fig. 6 is a circuit diagram of a loss-controlled variable-frequency oscillation generator similar to those of Figs. 1 and 4 but illustrating a modification thereof in which the frequency of oscillation is variably controlled by means of a pair of variable series resistances 21 and 23, instead of by the pair of potentiometers P1 and P2. The variable resistances 21 and 23 of Fig. 6 may be unitarily controlled in opposite directions electrically by a link mechanism 25, in the manner of controlling the two potentiometers P1 and P2 of Figs. 4 and 8 by means of the common control shaft 6. Resistances 27 and 29 may be disposed between the circuits 2 and 4 respectively and the ground 1.

Fig. 7 is a circuit diagram of a variable frequency oscillation generator similar to that of Fig. 6 but illustrating a modification thereof in which the frequency of oscillation is variably controlled by means of a single variable resistance 21 connected in series circuit relation with one of the networks N2, instead of by the pair of resistances 21 and 23 of Fig. 6 connected in both of the parallel circuits 2 and 4. The circuits of Figs. 6 and 7 accordingly utilize for controlling the frequency either one or two variable series resistances 21 and 23, in place of the potentiometers P1 or P2 of Figs. 1, 4 and 5. The variable control resistances 21 and 23 are relatively much smaller in resistance value than the resistances used in the networks N1 and N2.

Fig. 8 is a circuit diagram illustrating modifications in the frequency control and stabilization of a variable frequency oscillation generator of the bridge stabilized type. As illustrated in Fig. 8, the circuit comprises a μ -circuit amplifier or source of gain A, a main feedback circuit 3 interconnecting the output circuit 16 of the amplifier A with the input circuit thereof and comprising two separate parallel-connected transmission paths 4 and 2 which include therein two fixed networks N1 and N2, respectively, and two variable simultaneously-controlled potentiometer resistances P1 and P2, respectively, which are used for varying the frequency of oscillation, and a modification in an auxiliary feedback circuit 18 which may be used for automatically stabilizing the amplitude level of oscillations generated by any of the oscillations illustrated in this specification. The amplifier A may comprise any suitable source of gain such as a vacuum tube system comprising one or more linear high-gain

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vacuum tubes which function to amplify oscillations in a known manner. As an illustrative example, the amplifier represented by the block diagram A may comprise an amplifier as illustrated by the three-stage vacuum tube system, T1, T2 and T3 of Fig. 1.

The network N1 transmits the higher values of frequency with a relatively small loss and attenuates the lower values of frequency in the parallel branch 4 of the main feedback circuit 3 and 4. The network N2 transmits the lower values of frequency with a relatively small loss and attenuates the higher values of frequency in the other parallel branch 2 of the main feedback circuit 2 and 3. The two parallel-connected networks N1 and N2 taken together represent a frequency determining bridge network for the oscillation generator. The component elements of the network N1 and of the network N2 may be arranged in the form of any bridge suitable for the desired frequency of oscillation. As an illustrative example, the networks N1 and N2 may be in the form of known T-networks such as so-called double-T or parallel-T networks. As illustrated in Fig. 8, the networks N1 and N2 may consist only of resistances and condensers where it is desired to utilize the lower values of oscillation frequency, and such component resistances and condensers may be disposed in the form of T-networks of any desired form. As particularly shown in Fig. 8, the network N1 consists of two series condensers C3 and C4 and two shunt resistances R and R', forming a three-terminal network N1; and the three-terminal network N2 of Fig. 8 consists of three series resistances R2, R3 and R3' and two shunt condensers C2 and C2'. The component elements of the networks N1 and N2 may be constructed as fixed elements, as distinguished from mechanically variable or adjustable elements, and being of such fixed construction may be readily constructed in precision form.

In the case of low frequencies, capacitances and resistances only may be employed in the bridge networks N1 and N2 which may be of the type known as the double-T or parallel-T network as described for example in United States Patent 2,341,067 to R. O. Wise dated February 8, 1944. The double-T network comprises two individual T-networks one comprising series resistances and shunt capacitance, and the other comprising series capacitances and shunt resistance. The two networks are connected in parallel circuit relation with respect to each other in the main feedback circuit 3 of the amplifier A. While particular forms of the networks N1 and N2 have been illustrated in Fig. 8, it will be understood that the networks N1 and N2 may be of other forms such as for example those illustrated in Fig. 1 and in other figures of this application. The networks N1 and N2 may include resistances and condensers only, or they may be more complex and contain known inductances, amplifiers, phase shifters, etc. The component elements of the networks N1 and N2 may be chosen in accordance with known design techniques to give the same value of short circuit driving point admittance $Y11$ and must have suitable values for the short-circuit transfer admittance $Y12$.

The frequency control is provided by the variable voltage ratio, loss or gain control devices of Fig. 8 and is comprised of the potentiometers P1 and P2. Instead of being placed at the left end of the fixed networks N1 and N2 as illustrated in Fig. 8, they may be placed in a corresponding po-

sition at the right thereof, which is at the other or opposite side or end of the fixed networks N1 and N2, as illustrated in Fig. 20. The potentiometers P1 and P2 as illustrated in Fig. 8 may be constructed of two equal resistances P1 and P2 which are connected in the main feedback circuit 3 of the amplifier A, the potentiometer P1 being connected with the network N1 in the branch circuit 4 by means of the adjustable wiper 5 of the potentiometer P1, and the potentiometer P2 being connected with the other network N2 in the other parallel branch circuit 2 by means of the corresponding adjustable wiper 5 of the potentiometer P2. The two wipers 5 of the two potentiometers P1 and P2 respectively are insulated electrically with respect to each other but are interconnected mechanically so that a rotation of the potentiometer shaft 6 in one direction mechanically will introduce resistance values in opposite directions electrically into the parallel circuits 4 and 2, respectively. For example, a clockwise rotation of the potentiometer shaft 6 moves the movable wipers 5 from the position illustrated in Fig. 8 and will increase the resistance above the wiper 5 of the potentiometer P1 that is introduced into the circuit 4, and simultaneously will decrease the resistance above the wiper 5 of the potentiometer P2 that is introduced into the circuit 2, thereby simultaneously increasing the magnitude of the loss or attenuation introduced into the circuit 4 while decreasing the loss, or increasing the gain, introduced into the other parallel transmission path 2. The variable potentiometers P1 and P2 accordingly operate to introduce variable transmission loss in one of the two parallel transmission paths 2 and 4 with respect to the other for correspondingly varying the frequency of oscillation, and the frequency of oscillation may be selectively varied over a wide range of frequencies merely by adjustment of the potentiometers P1 and P2.

It will be noted that the potentiometers P1 and P2 are each made up of two resistances, namely, the resistances above and below the wipers 5 that are connected to the ground connection 1 and to the main feedback circuit 3. The absolute values of the over-all resistances of the potentiometers P1 and P2 have little effect upon the frequency control provided they are kept small relative to the values of the resistances which comprise the networks N1 and N2. The relative values of the resistances of the potentiometers P1 and P2 above and below the wipers 5 thereof are however of importance, and may easily be held to close limits for precision frequency control. From a practical standpoint, variable potentiometers such as the variable potentiometers P1 and P2 whether simultaneously or separately controlled are one of the most precise types of variable elements since their relative rather than absolute values of resistances may be used to make adjustments for control of the frequency of oscillation. The relative values of resistance above and below the wipers 5 may be kept to closer limits than absolute values of resistance and hence the potentiometer form of resistance is useful in precision frequency control as compared with series resistances of the non-potentiometer form where the absolute values of resistance become of more importance in precise frequency control.

The frequency of oscillation may be independent of changes in the absolute resistance values of the potentiometers P1 and P2 provided the relative resistances above and below the wipers 5 thereof remain correct to the extent that the ad-

mittances thereof are relatively greater than the short-circuit driving point admittances of the networks N1 and N2. For example, if the admittances of the potentiometers P1 and P2 are very large and the short-circuit driving point admittances of the networks N1 and N2 are sufficiently small in comparison, a change of 1 per cent in the absolute value of the resistances of the potentiometers P1 and P2 will have an effect on the frequency which is very small in comparison with 1 per cent, and this small effect will not appear if the absolute values of resistances of the two potentiometers P1 and P2 change by like amounts, as caused by equal temperature changes for example. If desired, one of the resistance potentiometers P1 and P2 may be omitted or not used, while the other is used to adjust the frequency of oscillations. For example, the potentiometer P1 corresponding to the network N1 may be omitted as illustrated in Fig. 5, and the voltage ratio of the other potentiometer P2 may be varied approximately as the square of the frequency of oscillation, or if a linear relation of frequency to potentiometer shaft position 6 of Fig. 8 is desired, this calls for an appropriate shaping for the resistance values and control mechanism of the potentiometer.

In order to stabilize the amplitude level of oscillations generated in the oscillator system comprising the amplifier A and the main feedback circuit 3, 2 and 4 including the networks N1 and N2 and the variable potentiometers P1 and P2, some form of non-linear level sensitive device may be utilized. It is desirable that the level sensitive device be capable of stabilizing the level of the amplitude of the oscillations generated throughout the operating range or band of frequencies in a variable frequency oscillator and for this purpose an auxiliary feedback circuit 18 from the output of the amplifier A to the input thereof may be provided as illustrated for example in Fig. 1 or in Fig. 8.

As illustrated in Fig. 8, the auxiliary feedback circuit 18 includes an automatic volume control rectifier 20 which may be of conventional form and which functions to rectify signal voltage from the output circuit 16 of the amplifier A and supply a proportional direct current control voltage to the grid electrode 9 of a variable transconductance vacuum tube 22. The tube 22 here operates as a variable loss device in accordance with variations in the amplitude of oscillations and may be adjusted to vary its transmission loss over a suitable range to stabilize the level of the amplitude of oscillations over the range of operating frequencies. A condenser 24 and a resistance 26 may be used in connection with the variable loss tube 22. Resistances 28 and 30 may be utilized to provide a fixed transmission loss of suitable value. The resistances 26, 28 and 30 may be adjusted to values that provide an infinite transmission loss for small amplitudes applied thereto from the output 16 of the amplifier A and to values that provide increasingly less loss in transmission for higher amplitudes of voltages. The voltage level stabilizer S of Fig. 8 accordingly may comprise some form of variable gain vacuum tube 22 with an automatic volume control connection 26. The tube 22 may be a non-linear type triode vacuum tube 22 which is biased below cut-off at low levels of signal voltage and overloaded at high levels of voltage. The resistance of the series resistance 26 and the capacitance of the condenser 24 interconnecting the plate 12 and the grid 9 of the triode 22 may be chosen

to be of suitable values to assist in obtaining a constant output voltage level over the band of operating frequencies.

A portion of the output of the amplifier A which is fed through the feedback circuit lead 18 reaches two parallel circuits namely, the circuit through the resistance 28, and the circuit through the automatic volume control rectifier 20 which may, as an illustrative example, be arranged as shown in Fig. 8. With no signal voltage applied to the lead 18, the negative voltage applied through the resistance R_g biases the vacuum tube VT to cut-off so that plate current does not flow, and the direct current potential of the lead wire 9a is then equal to the negative voltage applied at the bottom end of the resistance R_c . When signal voltage is applied through the lead 18 and the condenser C_g , plate current will flow during a portion of each cycle when the grid 9 of the tube VT is driven to a potential above cut-off, and the flow of this current through the resistance R_c will raise the potential of the cathode 7 and of the lead 9a above its no-signal value. The condenser C_c serves to by-pass the alternating current portion of the plate current so that only the direct current potential of the lead 9a will vary as the signal voltage is changed. The resistance R_i is used to prevent the condenser C_c from shorting out the alternating current signal voltage applied to the grid 9 of the tube 22, while the direct current potential applied to the lead 9a and at the cathode 7 of the tube VT is used as the grid bias of the tube 22. The circuit is so proportioned that the grid 9 of the tube VT is never positive, as this would short down the transmission through the path in parallel with the tube VT which is the path commencing with the resistance 28.

The signal voltage is attenuated in the resistances 28 and 30 and then divides, a part flowing through the resistance 26 to the input of the amplifier A while another part reaches the grid 9 of the tube 22 through the condenser 24. The voltage applied at the bottom of the resistance R_c is sufficient to bias the tube 22 nearly to cut-off so that the signal current flowing in the plate circuit of the tube 22 in response to the signal voltage applied to the grid 9 through the condenser 24 is equal to that reaching the output lead 18 through the resistance 26. Since the signal through the tube 22 is reversed in phase, these currents through the condenser 24 and the resistance 26 may be made to counteract each other so that no signal is returned to the input of the amplifier A. By somewhat over-biasing the tube VT, this condition may be maintained until the output voltage of the amplifier A reaches a predetermined magnitude. As the output voltage of the amplifier A further increases, the transmission through the resistor 26 will remain constant at a constant transmission loss, but the action of the tube VT raises the grid potential at 9a continuously and accordingly the loss through the tube 22 continually decreases. The signal from the plate circuit of the tube 22 thus more than balances that from the resistance 26, and the signal returned to the input of the amplifier A increases more rapidly than changes in the output voltage of amplifier A. The phase of the returned signal is opposite to that of the output of the amplifier A because of the reversal or turnover in the tube 22, and hence is correct to stabilize the output of the amplifier A. With increasing output of the amplifier A, the negative direct current voltage of the grid 9 of the tube

22 gradually decreases, and the following characteristics gradually increase with increase of output of the amplifier A, namely, the signal through the plate 12 of the tube 22 which is of negative phase, and the signal through the resistor 26 which is of opposite or positive phase. The difference of the last two characteristics results in the signal returned to the input of the amplifier A which gradually increases with increase in the output of the amplifier A.

The use of the resistor 26 and the delayed automatic volume control 20 as obtained by overbiasing the tube VT insures that this stabilizer path 18 does not give a positive feedback which may or might permit undesired oscillations to build up and interfere with the desired oscillation controlled by the main feedback path 2, 3 and 4. The use of the automatic volume control rectifier 20 in Fig. 8 requires a condenser C_c or equivalent in order to get the rectifier characteristic. A change too quick in frequency will upset the stabilization until the transient in the R_c - C_c circuit has died down. The circuits 18 of Figs. 1, 2 and 3 do not require a rectifier and hence are more suitable where very high speed variations in frequency occur.

When the networks N_1 and N_2 are built to high precision and the gain of the amplifier μ circuit A is very high, the auxiliary feedback stabilizing system 18 of Fig. 8 is readily made to feed back only at high amplitude levels. To reduce the introduction of harmonics in undesirable quantities, the sensitive automatic volume control rectifier 20 and the high loss device in the form of the series resistance 28 and the shunt resistance 30 in front of the tube 22 may be used. The amplitude stabilizer 18 may be made more or less slow-acting to a degree that is dependent upon the constancy of the level of oscillations, the rate at which the frequency of oscillation is to be changed and the precision to which the network system N_1 and N_2 is built. If the network system N_1 and N_2 were in perfect adjustment, the feedback through the variable stabilizer circuit 18 at equilibrium is independent of frequency change to the extent that it is made independent of frequency change. In practice, the stabilizer will reach equilibrium corresponding to the actual conditions. In general, only a small contribution from the amplitude stabilizer circuit 18 will be needed to establish any real frequency of infinite loss, if the potentiometer resistances P_1 and P_2 are made small relative to those of the networks N_1 and N_2 or stated differently if the transfer admittance of the potentiometers P_1 and P_2 is greater than the driving point admittances of the networks N_1 and N_2 .

An important feature of this invention is that the frequency of oscillation may be varied, not by mechanically changing or varying in the conventional manner the values of resistance and capacitance of the networks N_1 and N_2 , but rather by varying the resistance and voltage ratios of the potentiometers P_1 and P_2 . Thus the resistances and capacitances or other component elements of the networks N_1 and N_2 per se may be fixed elements in so far as variation in the variable frequency of oscillation is concerned, and the frequency of oscillation may be varied merely by varying the resistances of the potentiometers P_1 and P_2 which are relatively small resistances compared with those of the networks N_1 and N_2 . The ease with which the frequency of oscillations may be controlled by the variable potentiometers P_1 and P_2 or

other variable loss control devices is an important feature of this invention. The frequency of oscillation is varied by supplying similar phased inputs to the two fixed networks N1 and N2, adding or subtracting the outputs thereof, and setting the phase of the resultant output thereof by the relative magnitudes of the inputs to the networks N1 and N2. Thus, by using the two potentiometers P1 and P2 to control the two inputs to the respective networks N1 and N2 in opposite directions, a change in phase of the resultant outputs at a given frequency will occur with a change in rotation of the wipers 5 of the potentiometers P1 and P2. Such change in phase will be little or large in the output depending upon the actual phase differences at the output of the networks N1 and N2, and the frequency of oscillation will change until the resultant phase is that necessary for oscillation.

In general, high frequency stability may be obtained with bridge type of oscillators which use a so-called μ and β loop circuit, the μ circuit comprising an amplifier or source of gain A having over-all input and output circuits which are interconnected by means of a suitable network system to form the main feedback β circuit 3. The β circuit also includes some form of non-linear voltage variable stabilizing element which might be a conventional thermistor, or tungsten or carbon lamp, but which preferably is an auxiliary feedback circuit 13 which automatically stabilizes the amplitude level of oscillations and is so arranged by proper adjustment of the amplitude level stabilizer S that it gives infinite loss or attenuation in the β circuit at or very near to the oscillation frequency. At a point of infinite loss $\beta=0$. It is known that the natural frequency modes of a simple feedback amplifier are at frequencies of $\mu\beta=+1$, which transposed gives $1-\mu\beta=0$. There will be such a frequency close to any frequency of $\beta=0$. If the amplifier gain is large at that frequency, then a small value of β times the large value of μ is equal to unity. Thus with a high gain μ -circuit A, the feedback bridge type oscillator will oscillate at a frequency close to that of $\beta=0$ with the proper adjustment of the stabilizer circuit 13 which automatically adjusts itself to the proper value by the amplitude level of the oscillations. The bridge type oscillator may be any such oscillator embodying the principle of using a point where $\beta=0$ on proper adjustment of the amplitude level stabilizers.

An important characteristic of the β -circuit is the location of the infinite loss positions. It is convenient to assume zero termination at both ends of the network or β -circuit. This is because the short-circuit admittance Y12 of parallel-connected networks N1 and N2 is the sum of the corresponding short-circuit transfer admittances of the separate networks N1 and N2, as is known. In accordance with conventional usage, Y11 is the short-circuit driving point admittance of the associated network with its output terminals shorted, Y12 is the short-circuit transfer admittance of the associated network with its output terminals shorted, and Y is a function of p where p equals $i\omega$ and of $p_0=i\omega_0$ where ω_0 is an assigned fixed frequency in radians per second. The driving point admittances Y11 on short circuit are the same for networks N2 and N1 in the circuit of Fig. 8 while their corresponding transfer admittances Y12 are of the form

$$\frac{p_0 Y}{p}$$

and

$$\frac{p Y}{p_0}$$

- 5 respectively, where Y is the same function of p for both, and where $p=i\omega$ in accordance with convention. The admittances Y_{p_0} of the two equal potentiometers from the wipers 5 to ground 1 are approximately the same for both of the
10 potentiometers P1 and P2 when the wiper positions are as per K and $1-K$, respectively, where K represents the ratio of resistances. Under these conditions, the transfer impedance of the combined networks N1 and N2 and the associated
15 potentiometers P1 and P2 is:

$$\left[K \frac{p_0}{p} + (1-K) \frac{p}{p_0} \right] Y'$$

where

$$20 \quad Y' = \frac{Y}{1 + Y11/Y_{p_0}}$$

From these expressions it follows that when the variable loss amplitude stabilizing system 13
25 is set for infinite loss, there will be a real frequency ω of infinite loss for the complete β network at

$$\omega = p_0 \sqrt{\frac{K}{1-K}}$$

- 30 As the gain μ of the amplifier A approaches infinity or is high, the singing frequency will approach this frequency ω . When the variable loss network system is set a little off infinite
35 loss setting, the natural mode becomes complex and of the form $\alpha \pm i\omega$, which corresponds to a mode of oscillation of the form $e^{-\alpha t} \sin(\omega t + \theta)$. The damping value of α will be positive which represents a decreasing oscillation level when
40 the variable loss gives transmission in phase, and negative representing an increasing oscillation level when the variable loss gives transmission with reversal in phase. Thus in order to
45 stabilize the level of oscillations, the automatic volume control of the variable loss amplitude stabilizer S should ideally be such that low amplitude levels give transmission with reversal in phase leading to increasing levels, while high
50 amplitude levels give direct transmission leading to decreasing amplitude levels. The amplitude stabilizer S of the auxiliary feedback circuit 13 of Fig. 8 conforms with the latter condition.

Fig. 9 is a circuit diagram of a variable frequency oscillator illustrating one of several ways
55 in which a variometer or variocoupler VC may be used to variably control the frequency of oscillation. As illustrated in Fig. 9, the variometer VC is disposed in the output of the networks N1 and N2 and may consist of inductance windings
60 31, 33 and 35. The windings 31 and 33 may be the fixed crossed primary windings, and the winding 35 may be the rotatable secondary winding of the variometer VC. The coupling between the windings 31 and 35 may be a maximum when the
65 coupling between the windings 33 and 35 is zero, and vice versa. By variation in the rotatable secondary winding 35, the frequency of oscillations may be varied.

It will be understood that in loss-controlled
70 variable frequency oscillators, the variable loss may be controlled not only by variable potentiometers or variable series resistances but alternatively, the loss may be variably controlled by variable inductances such as the variable voltage ratio variometer VC of Fig. 9 which like the

potentiometer may be used as a means of producing a variable voltage ratio to obtain a variable frequency of oscillation. In addition, a variety of other variable loss type means, as described hereinafter, may be used for producing the variable loss or voltage ratio required to obtain a variable frequency of oscillation. As an example, vacuum tube circuits may be arranged in such a way that their through gains or losses may be varied by varying auxiliary control voltages applied thereto in order to obtain variations in the oscillator frequency. This method as more fully described hereinafter offers a convenient means of obtaining frequency modulation of an oscillator or other circuit.

Figs. 10, 11, 12 and 13 are circuit diagrams illustrating examples of systems for converting amplitude modulators into frequency modulators that are relatively simple in construction in comparison with conventional frequency modulation systems using any but very small percentage modulations. In the circuits illustrated in Figs. 10 to 13, variations in or modulation of the frequency of oscillations are obtained by varying the gains or losses in the circuit by means of auxiliary control voltages which may be supplied from an amplitude modulator connected therewith.

Fig. 10 is a circuit diagram illustrating a loss-controlled frequency modulated oscillator in which the loss is variably controlled and the frequency is correspondingly modulated by means of an electronic vacuum tube or other suitable amplitude type modulator M controlled by variable signal control voltages applied by a suitable signal source 36 to the amplitude modulator M. Except for the substitution of the amplitude modulator M in place of the potentiometer P2 of Fig. 5, the circuit of Fig. 10 is similar to that of Fig. 5, the corresponding parts being given like reference characters. In the loss-controlled oscillator circuit of Fig. 5, the loss is variably controlled by a potentiometer P2 which may be used as a means for producing a variable voltage ratio in order to control the frequency of oscillations. In the loss-controlled oscillator of Fig. 10, the loss is variably controlled by the modulator M by varying the auxiliary signal control voltage applied thereto from the signal modulated voltage source 36 in order to correspondingly modulate the frequency of oscillations. The system illustrated in Fig. 10 accordingly offers a simple and convenient means for obtaining frequency modulation of relatively large percentage modulation, and for converting amplitude modulators into frequency modulators.

The modulator M of Fig. 10 may be any suitable amplitude modulator of the linear amplitude type or otherwise. If a known square law type of amplitude modulator M relating its voltage ratio K to the frequency of oscillations in accordance with Equation 7 be used in the circuit of Fig. 10, linear frequency modulation thereof may be obtained. As illustrated in Fig. 10, the input connections to the amplitude modulator M are conventional. The amplitude modulated signal is applied from the signal source 36 to the input grid 10 of the amplitude modulator M. The carrier frequency from the principal feedback circuit 3 is applied to the input grid 9 of the amplitude modulator M. The modulated carrier output is taken from the output plate electrode 12 of the amplitude modulator M and applied to the input of the network N2. The frequency modulated output signal may be taken from the output terminals 1 and 16 which are

connected with the output circuit of the amplifier μ circuit A. The level sensitive amplitude stabilizer S may be in accordance with that of Figs. 1 or 8.

Fig. 11 is a circuit diagram illustrating a loss-controlled frequency-modulated oscillator which is similar to that of Fig. 10 except for using two amplitude modulators M1 and M2 which may be linear type electronic amplitude modulator tubes of conventional design. By using the two amplitude modulators M1 and M2 connected as illustrated in Figs. 11 and 14, linear frequency modulation may be obtained with the conventional linear type of amplitude modulators. As shown in Figs. 11 and 14, the two amplitude modulators M1 and M2 may be connected in tandem, the carrier frequency from the main feedback circuit 3 being applied to the input grid 9 of the first modulator tube M1 and the output of the first modulator tube M1 being applied to the input grid 9 of the second modulator tube M2. The modulated carrier output from the second modulator tube M2 is applied to the input of the network N2. The amplitude modulated signal from the signal source 36 is applied to the grid circuits 10 of both of the amplitude modulator tubes M1 and M2.

Fig. 12 is a circuit diagram illustrating a loss-controlled frequency modulated oscillator which is similar to that of Fig. 11 except for the connection arrangement of the two amplitude modulators M1 and M2, and the addition of a three-winding transformer 37 having a primary winding L1 and two series-connected secondary windings L2 and L3 which are connected in series aiding relation. By using two amplitude modulators M1 and M2 connected as illustrated in Fig. 12, fairly good linear frequency modulation may be obtained even when using linear type amplitude modulator tubes M1 and M2, and fairly large percentages of modulation may be obtained. With the two amplitude modulators M1 and M2 connected in the parallel arms of the two networks N1 and N2 respectively, the carrier frequency from the main feedback circuit 3 is simultaneously applied to the inputs of the amplitude modulators M1 and M2, and the modulated carrier outputs thereof are applied to the networks N1 and N2 respectively. The amplitude modulated signal from the signal source 36 and the transformer 37 is applied to input circuits of both of the modulators M1 and M2.

Fig. 13 is a circuit diagram illustrating a frequency modulated oscillator which is like that of Fig. 12 except for illustrating a particular type of amplitude modulating devices M1 and M2 each in the form of a conventional multigrid mixer tube, which may be of the type disclosed for example in United States Patent No. 1,896,780, issued February 7, 1933, to F. B. Llewellyn. As illustrated in Fig. 13, the carrier frequency from the main feedback circuit 3 may be applied to the inner grid 9 of each of the two modulator tubes M1 and M2, and the amplitude modulated signal from the signal source 36 and from the end windings L2 and L3 of the transformer 37 may be applied to the other grids 10 of the modulator tubes M1 and M2. The modulated carrier outputs from the plate electrodes 12 of the modulator tubes M1 and N2 are applied to the networks N1 and N2 respectively. While particular types of modulator tubes M1 and M2 and circuit connections have been illustrated in Fig. 13, it will be understood that other circuit connections and other types of amplitude modulat-

ing devices M1 and M2 may be used, such as for example known types of copper oxide modulating devices.

Fig. 14 is a circuit diagram of a frequency modulated oscillator similar to that of Fig. 11 but illustrating a modification in which the fixed networks N1 and N2 are adapted for high frequency operation and accordingly are comprised of inductance and capacitance type elements only, such as condenser elements 41 and 44 and inductance elements 42 and 43, rather than resistance and capacitance type elements as used in the low frequency type fixed networks N1 and N2 shown in Figs. 10 to 13. As shown in Fig. 14, a series resistance 46 may be provided in the bridge circuit 4.

Fig. 15 is a circuit diagram of a frequency modulated oscillator similar to that of Fig. 14 but illustrating a modification for obtaining automatic frequency control, as in radio receivers for example. For this purpose, signal currents from the crossed resonant circuits of a conventional audio frequency circuit may be applied to input terminals 51 of rectifiers 52 and 53, and signal currents from the output of rectifiers 52 and 53 may be applied through filters each comprising a series resistance 54 and a shunt condenser 55 to the input grids 10 of the modulator tubes M1 and M2 respectively. The amplitude modulating signal is accordingly applied to one set of the grids 10 of the modulator tubes M1 and M2, and the other grid 9 of each of the mixer type tubes M1 and M2 carries the corresponding frequency modulated signal. In operation, the gain of the tube M1 increases when the frequency is low and thereby increases the frequency again. Similarly, the gain of the tube M2 increases when the frequency is high and thereby reduces the frequency again.

Figs. 16 and 17 are circuit diagrams of variable frequency oscillator circuits illustrating networks N1 and N2 of the inductance-capacitance type for determining the frequency of oscillations, and generalized variable attenuation devices Z1 and Z2 for modulating or variably controlling the oscillation frequency of the circuit. The variable loss type devices Z1 and Z2 may generally represent the variable potentiometers P1 and P2 of Figs. 1, 4, 5 and 8, the variable resistances 21 and 23 of Figs. 6 and 7, the variable inductance variometer V of Fig. 9, the modulators M of Figs. 10 to 15, or other similar variable loss or variable voltage ratio devices which may function to vary the relative transmission levels of the two parallel branch paths 2 and 4 of the circuit in order to correspondingly vary the frequency of oscillations. The network N1 as illustrated in Fig. 16 is a high-pass network consisting of a series condenser 41 and a shunt inductance 42 and the network N2 is a low-pass network consisting of a series inductance 43 and a shunt capacitance 44. Series resistances 45 and 46 may be connected in the parallel transmission paths 2 and 4 respectively.

Fig. 17 is a circuit diagram of a variable frequency oscillator similar to that of Fig. 16 but illustrating a modification in the form of the inductance-capacitance bridge networks which may be used in the frequency determining system. As illustrated in Fig. 17, a series inductance winding 43 and a series condenser 41 may be utilized in the parallel transmission paths 2 and 4 respectively, and a shunt resistance 47 may be connected to ground 1 from the connection joining the inductance 43 and the condenser 41.

The variable loss devices Z1 and Z2, which are variable short-circuit impedances used for frequency control, may be made of relatively small impedance values in order to minimize the amount of output voltage amplitude variation with frequency change produced by the frequency control devices Z1 and Z2. A voltage stabilizer system S may be provided of the loss type.

Figs. 18, 19 and 20 are generalized circuit diagrams of variable frequency oscillation generators which include a μ -circuit amplifier represented as before by the block triangle A, and a β circuit represented by the fixed networks N1 and N2 and variable loss devices Z1 and Z2 connected in the main feedback circuit 1 and 3 of the high gain amplifier A. The bridge type feedback circuit 3 connected between the output 16 of the amplifier A and the input thereof is divided into two parallel-connected transmission paths 2 and 4, one of which includes the fixed network N1 and the variable loss device Z1, and the other of which, namely, the transmission path 2 includes the fixed network N2 and the variable loss device Z2. The variable impedances Z1 and Z2 may be connected in series circuit relation with the fixed networks N1 and N2, respectively, in the parallel transmission paths 4 and 2, respectively.

The variable devices Z1 and Z2 and the fixed networks N1 and N2 of the β -feedback circuit 3 may be connected in circuit by means of the direct parallel connection as illustrated in Fig. 18, or by means of transformers 57 connected as illustrated in Fig. 19, or by a combination of the two as illustrated in Fig. 20. While the specific circuits, given hereinbefore in this specification as examples of specific embodiments for the β circuit fixed networks N1 and N2 and for the variable devices Z1 and Z2, follow the generalized circuit connections given in Fig. 18, it will be understood that they may follow other connections such as those given in Figs. 19 and 20, for example.

The output voltage oscillations may be taken off from any pair of terminals connected with the lines labeled by the reference characters 1 and 16 in Figs. 18 to 20.

The μ -circuit amplifier A comprises a source of gain such as a vacuum tube system of high gain. In the main feedback path 3, there are two parallel transmission paths 2 and 4 each contributing to the total feedback. In the two branch feedback paths 2 and 4, the networks N1 and N2 are fixed frequency determining networks, and the variable loss devices Z1 and Z2 are frequency control devices for varying or modulating or determining the over-all frequency of oscillation by supplying a variable loss or gain to the networks N1 and N2. While both of the networks N1 and N2 make contributions to the total feedback, they have different but related attenuation properties, one being low-pass and the other high-pass and accordingly they produce relative phase shifts therein which differ by an amount which is substantially 180 degrees, independent of frequency change. The relative amplitudes, however, of the voltages supplied by the networks N1 and N2 vary differently with frequency change, the typical difference being a change in the relative levels of the two networks N1 and N2 at a rate which may be substantially of the order of 12 decibels per octave of frequency change. The above conditions apply to the relative phases and amplitude levels of the contributions from the two networks N1 and N2 to the

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total feedback but do not apply to the actual over-all phase and amplitude level of the total feedback. The amplitude levels are equal in the two networks N1 and N2 at a definite frequency which is substantially the frequency of oscillation. In practice, the network N2 may comprise, for example, a T network composed of two series resistances and a shunt capacitance for low frequencies, or a series inductance and a shunt capacitance for high frequencies; and the network N1 may comprise, for example, a T network composed of two series capacitances and a shunt resistance for low frequencies or a series capacitance and a shunt inductance winding for high frequencies, as illustrated in the specific embodiments described hereinbefore.

The frequency control devices Z1 and Z2 represent means for introducing variable flat loss or gain into one or both of the two parallel connected transmission paths 2 and 4 in such a way that the difference between the amplitude levels thereof can be changed at will and in amounts that may be substantially independent of the value of the frequency if the amplitude levels are measured on a logarithmic or decibel basis and in such a way as to leave the phases of the two parallel paths 2 and 4 substantially unchanged. The variation in the flat loss then controls the variation in the frequency of oscillation by varying the frequency at which the amplitude levels of the two parallel paths 2 and 4 are equal. The loss variation devices Z1 and Z2 which produce such variations in the frequency of oscillation of the bridge feedback type oscillators illustrated in Fig. 18 and other figures may be in the form of variable potentiometers, variable resistances, variable inductance windings and modulator tubes for example as illustrated in the several circuits described in this specification. In addition, a voltage stabilizer S is provided for automatically stabilizing the amplitude of the output voltage at a level value over a band or range of operating frequencies of the variable frequency oscillation circuit. Such a voltage stabilizer may operate by producing an effective change in one of the networks N1 or N2, or by changing the feedback in an auxiliary or subsidiary feedback, or by other suitable means.

In the oscillator circuits of Figs. 18 to 20, the forward voltage ratio μ is normally large representing high gain as obtained by the amplifier source of gain A, and the backward or feedback voltage ratio β is normally less than unity representing a loss as obtained from the feedback circuit 3, 2, 4 comprising the network condensers, coils and resistances represented by the devices N1, N2, Z1 and Z2. In the β -circuit 3, the value of β equals 0 at some particular frequency of oscillation $\omega = \omega_0$ when the amplitude level sensitive element S has a particular voltage $V = V_0$. The β -circuit 3 is so arranged that it is a minimum phase network when voltage V is greater than V_0 . The corresponding μ circuit A is such that the gain μ is large at the particular frequency $\omega = \omega_0$, and such that oscillations build up when V is less than V_0 and damp down when V is greater than V_0 . For practical purposes, this amounts designing the μ -circuit A under the minimum phase condition when voltage V is greater than voltage V_0 . The amplitude of the resulting oscillations increases until a point is reached where steady state oscillations at constant amplitude take place, due to the level stabilizing action of the non-linear circuit element S which determines the level of the steady state

oscillations. In practice, the specific element S and connections chosen for stabilizing a specific application depend on such conditions as frequency range, limits on frequency drift, and whether the frequency is to be varied rapidly or slowly. For example, while known thermistor stabilizers yield low harmonic generation they do not follow rapid variations in frequency as well as do gas tubes, or the auxiliary feedback stabilizer S. The most suitable type of amplitude stabilizer may be selected for the particular application. In general, some type of amplifier stabilizer is needed in variable frequency oscillators.

Although this invention has been described and illustrated in relation to specific arrangements, it is to be understood that it is capable of application in other organizations and is therefore not to be limited to the particular embodiments disclosed.

What is claimed is:

1. A variable frequency oscillation system comprising an amplifier or main source of gain, and a main feedback path connected from the output of said amplifier to the input thereof, said main feedback path including a pair of parallel-connected transmission paths, said parallel transmission paths including variable frequency control means and separate frequency sensitive networks contributing to the total feedback in said main feedback path, said pair of parallel transmission paths having related phase and amplitude characteristics, the relative phase characteristics of said pair of parallel transmission paths differing by substantially 180 degrees and being substantially independent of a change in the value of said variable frequency, and the relative amplitude characteristics of said pair of parallel transmission paths varying differently with a change in the value of said variable frequency and being equal in amplitude at a definite frequency corresponding substantially to the value of said oscillation frequency, said frequency control means constituting means separate from said networks for introducing variable attenuation or loss into at least one of said pair of parallel transmission paths for varying said relative amplitude characteristics thereof for thereby varying the value of said variable oscillation frequency in accordance with the value of said variable loss, and means for stabilizing the amplitude level of said oscillations over a range of values for said variable frequencies comprising an auxiliary feedback path for said source of gain, said auxiliary feedback path being in addition to said pair of main parallel-connected feedback paths and comprising means including electron tube apparatus for producing unidirectional currents and yielding variable loss in said auxiliary feedback path in accordance with the amplitude of said oscillations.

2. A system in accordance with claim 1 wherein said frequency control means comprises a variable resistance connected in series circuit relation in at least one of said pair of parallel transmission paths.

3. A system in accordance with claim 1 wherein said frequency control means comprises a variable resistance connected in each of said pair of parallel transmission paths.

4. A system in accordance with claim 1 wherein said frequency control means comprises a variable resistance potentiometer having a tap connection thereon connected in circuit with one of said net-

works disposed in one of said pair of parallel transmission paths.

5. A system in accordance with claim 1 wherein said frequency control means comprises a pair of variable resistance potentiometers each having a tap connection thereon individually connected in circuit with different ones of said networks disposed in said pair of parallel transmission paths.

6. A system in accordance with claim 1 wherein said frequency control means comprises a variable resistance potentiometer connected in circuit with each of said networks disposed in each of said pair of parallel transmission paths, said potentiometers comprising means including simultaneously operated brushes thereon for simultaneously connecting substantially equal resistance portions of each of said potentiometers individually into said pair of respective parallel transmission paths for giving substantially equal short-circuit driving point admittances looking back into said potentiometer brushes, said networks disposed in each of said parallel transmission paths having substantially equal driving point admittances.

7. A system in accordance with claim 1 wherein said frequency control means comprises a pair of variable resistance potentiometers having resistance portions thereof simultaneously connected in each of said pair of parallel transmission paths, and control means for simultaneously operating said potentiometers.

8. A system in accordance with claim 1 wherein said frequency control means comprises a variable resistance potentiometer connected with each of said networks disposed in each of said pair of parallel transmission paths, and means for simultaneously operating said potentiometers, one of said networks comprising two series-connected capacitors and two shunt-connected resistors and the other of said networks comprising three series-connected resistors and two shunt-connected capacitors.

9. A system in accordance with claim 1 wherein said frequency control means comprises a signal-modulated electronic apparatus connected in push-pull relation in each of said pair of parallel transmission paths, said electronic apparatus constituting means for modulating and amplifying said signal transmitted therethrough.

10. A system in accordance with claim 1 wherein said frequency control means introducing said variable loss comprises signal modulated electronic apparatus connected in at least one of said pair of parallel transmission paths, said electronic apparatus constituting means for modulating and amplifying said signal transmitted therethrough.

11. An oscillation generator comprising an amplifier, input and output circuits therefor, a feedback network system connected between said input and output circuits, said network system comprising a pair of fixed T networks connected in separate parallel paths between said input and output circuits, one of said T networks consisting of two series resistances and a shunt capacitance and the other of said T networks consisting of two series capacitances and a shunt resistance, said capacitances and resistances being of related values proportioned to provide substantially maximum attenuation and phase reversal in said networks at a frequency selected by frequency control means for varying said frequency of said oscillations over an operating range of frequencies, said frequency control means comprising signal modulating means comprising an amplifying electron tube connected with at least one of said networks and introducing variable loss in the

transmission in at least one of said parallel paths in accordance with the amplitude of said signal, and amplitude control means for stabilizing the level of amplitude of said oscillations over said range of frequencies.

12. A variable frequency oscillation generator comprising an amplifying electron tube system, a feedback circuit for said tube system, a pair of parallel transmission paths in said feedback circuit, a network disposed in each of said parallel transmission paths, said networks having different attenuation characteristics and having substantially the same phase shift characteristics except for a 180-degree phase difference therein, and attenuation means including signal modulating means comprising an amplifying electron tube connected in circuit relation with at least one of said networks for varying the relative transmission level of one of said networks with respect to the other of said networks and for correspondingly modulating the frequency of said oscillation generator in accordance with the amplitude of said signal.

13. A generator of oscillations comprising a main source of gain, a main feedback path comprising a pair of parallel-connected transmission paths interconnecting the input and output circuits of said source of gain, frequency-sensitive networks disposed in said respective parallel paths and having relatively high-pass and low-pass frequency transmission characteristics with respect to each other, means including a signal responsive modulator disposed in each of said parallel paths in circuit relation with said respective networks for introducing variable loss in said parallel paths and modulating the frequency of said oscillations in accordance with the amplitude of said signal, and means including electron tube apparatus disposed in an additional feedback path interconnecting the input and output circuits of said source of gain for controlling the amplitude of said oscillations.

14. A generator of oscillations comprising a main source of gain, a main feedback path comprising a pair of parallel-connected transmission paths interconnecting the input and output circuits of said source of gain, frequency-sensitive networks disposed in said respective parallel paths and having relatively high-pass and low-pass frequency transmission characteristics with respect to each other, means including a signal responsive modulator disposed in each of said parallel paths in circuit relation with said respective networks for introducing variable loss in said parallel paths and modulating the frequency of said oscillations in accordance with the amplitude of said signal, said modulators comprising amplifying vacuum tubes, and means including electron tube apparatus disposed in an additional feedback path interconnecting the input and output circuits of said source of gain for controlling the amplitude of said oscillations.

15. A generator of oscillations comprising a main source of gain, a main feedback path comprising a pair of parallel-connected transmission paths interconnecting the input and output circuits of said source of gain, frequency-sensitive networks disposed in said respective parallel paths and having relatively high-pass and low-pass frequency transmission characteristics with respect to each other, means including a signal responsive modulator disposed in each of said parallel paths in circuit relation with said respective networks for introducing variable loss in said parallel paths and modulating the frequency

of said oscillations in accordance with the amplitude of said signal, said modulators comprising amplifying vacuum tubes and constituting means for controlling the amplification of said signal transmitted through said modulator vacuum tubes, said vacuum tube modulators being similar in amplification characteristics, and connected in push-pull relation and located symmetrically with respect to each other in said pair of parallel paths for substantially canceling out equal changes in the gain of said modulator tubes, and means including electron tube apparatus disposed in an additional feedback path interconnecting the input and output circuits of said source of gain for controlling the amplitude of said oscillations.

16. A generator of oscillations comprising a main source of gain, a main feedback path comprising a pair of parallel-connected transmission paths interconnecting the input and output circuits of said source of gain, frequency-sensitive networks disposed in said respective parallel paths and having relatively high-pass and low-pass frequency transmission characteristics with respect to each other, means including a signal responsive modulator disposed in each of said parallel paths in circuit relation with said respective networks for introducing variable loss in said parallel paths and modulating the frequency of said oscillations in accordance with the amplitude of said signal, said modulators comprising a pair of vacuum tubes each having an anode electrode connected with one of said networks, a grid electrode connected with said output circuit of said main source of gain and another grid electrode connected with the source for said signal, and means including electron tube apparatus disposed in an additional feedback path interconnecting the input and output circuits of said source of gain for controlling the amplitude of said oscillations.

17. A generator of oscillations comprising a main source of gain, a main feedback path comprising a pair of parallel-connected transmission paths interconnecting the input and output circuits of said source of gain, frequency-sensitive networks disposed in said respective parallel paths and having relatively high-pass and low-pass frequency transmission characteristics with respect to each other, means including a signal responsive modulator disposed in each of said parallel paths in circuit relation with said respective networks for introducing variable loss in said parallel paths and modulating the frequency of said oscillations in accordance with the amplitude of said signal, said modulators comprising a pair of vacuum tubes each having an anode electrode connected with one of said networks, a grid electrode connected with said output circuit of said main source of gain and another grid electrode connected with the source for said signal, said signal source being an alternating current source, and means including electron tube apparatus disposed in an additional feedback path interconnecting the input and output circuits of said source of gain for controlling the amplitude of said oscillations.

18. A generator of oscillations comprising a

main source of gain, a main feedback path comprising a pair of parallel-connected transmission paths interconnecting the input and output circuits of said source of gain, frequency-sensitive networks disposed in said respective parallel paths and having relatively high-pass and low-pass frequency transmission characteristics with respect to each other, means including a signal responsive modulator disposed in each of said parallel paths in circuit relation with said respective networks for introducing variable loss in said parallel paths and modulating the frequency of said oscillations in accordance with the amplitude of said signal, said modulators comprising a pair of vacuum tubes each having an anode electrode connected with one of said networks, a grid electrode connected with said output circuit of said main source of gain and another grid electrode connected with the source for said signal, said signal source being an alternating current source comprising a three-winding transformer having a primary winding and two secondary windings connected in series-aiding relation with respect to each other and connected with said signal grid electrodes of said modulator tubes, and means including electron tube apparatus disposed in an additional feedback path interconnecting the input and output circuits of said source of gain for controlling the amplitude of said oscillations.

19. A generator of oscillations comprising a main source of gain, a main feedback path comprising a pair of parallel-connected transmission paths interconnecting the input and output circuits of said source of gain, frequency-sensitive networks disposed in said respective parallel paths and having relatively high-pass and low-pass frequency transmission characteristics with respect to each other, means including a signal responsive modulator disposed in each of said parallel paths in circuit relation with said respective networks for introducing variable loss in said parallel paths and modulating the frequency of said oscillations in accordance with the amplitude of said signal, said modulators comprising a pair of vacuum tubes having substantially linear amplification characteristics, and means including electron tube apparatus disposed in an additional feedback path interconnecting the input and output circuits of said source of gain for controlling the amplitude of said oscillations.

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