Feb. 8, 1944. R. O. WISE 2,341,067

VARIABLE FREQUENCY BRIDGE STABILIZED OSCILLATOR

Original Filed June 14, 1941 2 Sheets-Sheet 1

FIG. 1

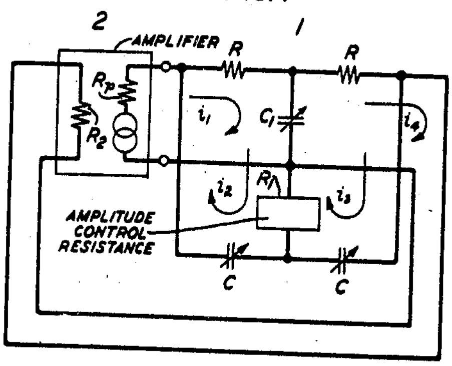


FIG. 3

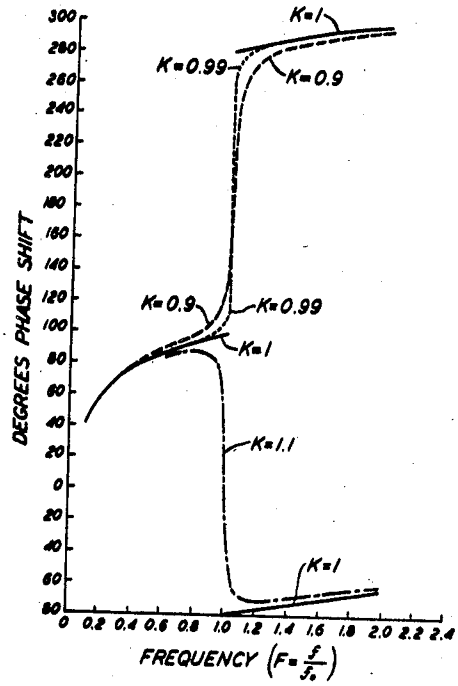
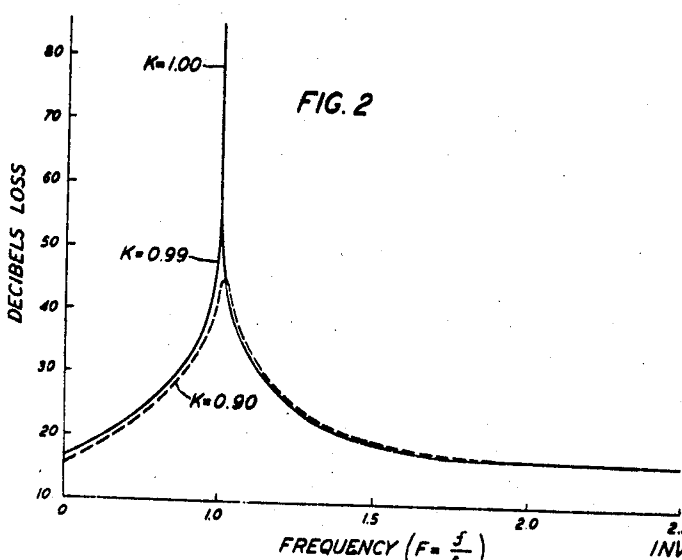


FIG. 2



INVENTOR
R. O. WISE
BY Guy T. Morris
ATTORNEY

Feb. 8, 1944.

R. O. WISE

2,341,067

VARIABLE FREQUENCY BRIDGE STABILIZED OSCILLATOR

Original Filed June 14, 1941

2 Sheets-Sheet 2

FIG. 4

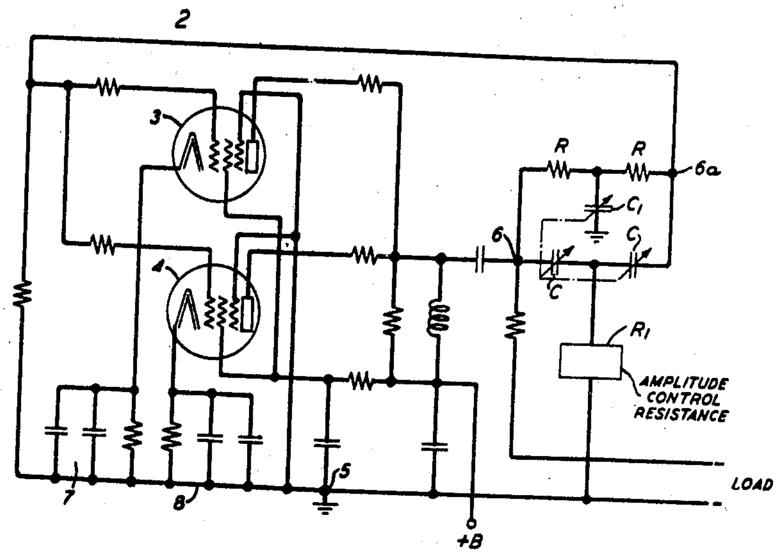
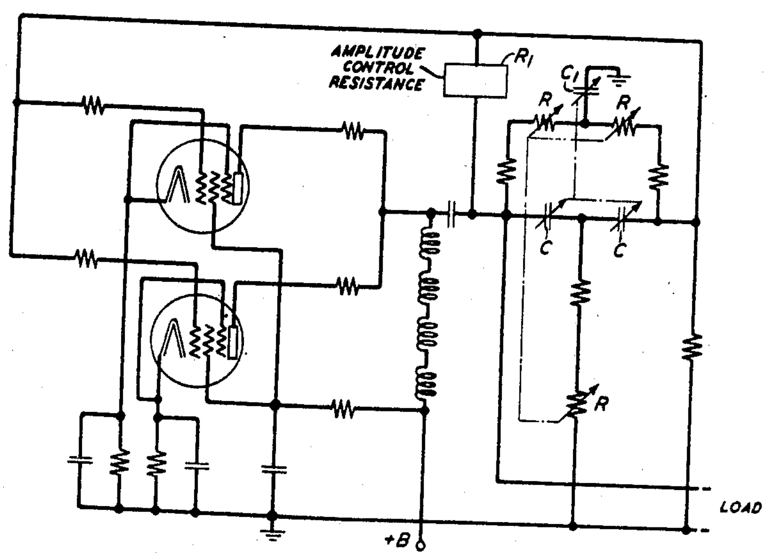


FIG. 5



INVENTOR
R. O. WISE
BY Guy T. Morris

ATTORNEY

Patented Feb. 8, 1944 2,341,067

UNITED STATES PATENT OFFICE

2,341,067

VARIABLE FREQUENCY BRIDGE STABILIZED OSCILLATOR

Raymond O. Wise, Short Hills, N. J., assignor to Bell Telephone Laboratories, Incorporated, New York, N. Y., a corporation of New York

Original application June 14, 1941, Serial No. 398,042, now Patent No. 2,319,965, dated May 25, 1943. Divided and this application May 1, 1942, Serial No. 441,318

7 Claims. (Cl. 250—36)

This invention relates to oscillation generating circuits, that is, oscillators, and more particularly to oscillators of the bridge stabilized type. This application is a division of application Serial No. 398,042, filed June 14, 1941, Patent No. 2,319,965 granted May 25, 1943.

It is the object, generally, of the invention to improve the frequency stability of vacuum tube oscillators, especially as resulting from changes in the various supply voltages and non-linearity in stability-significance circuit elements, either comprised in the circuits immediately associated with the vacuum tube or elsewhere. A subsidiary object is, analogously to the above, to improve the stability of such oscillators with respect to the amplitude of the generated oscillations. Another object is the improvement of the wave form of the oscillations and therefore the suppression of harmonics of the fundamental frequency.

In recent years it has been recognized that it is largely the harmonics produced by the above non-linearity which affect the frequency stability. This affect occurs since any changes in the supply voltages, which because of the non-linearity, will tend to produce relative changes in the amplitudes of the harmonics, will cause a change in the reactance at the fundamental frequency as well and hence a change in that frequency. Various methods of minimizing these affects have been proposed but none have been completely satisfactory. For instance, Llewellyn has shown in his paper "Constant frequency oscillators" in the Proceedings of the Institute of Radio Engineers for December, 1931, that by including appropriate impedances in the grid and plate connections of the tube, the frequency of oscillations may be made independent of changes in the grid and plate resistances of the tube. This method, however, is not adapted to the elimination of instability resulting from harmonic reaction. An approach was made by Arguimbau in his paper "An oscillator having a linear operating characteristic," Proceedings of the Institute of Radio Engineers for January, 1933, by suggesting the operation of the oscillator vacuum tube as a linear amplifier, as by rectifying a portion of the output to furnish a grid bias to the vacuum tube and therefore achieving an automatic volume control which maintains an equality between the gain of the amplifier and the loss of the associated oscillating circuit. A somewhat similar idea was proposed by Groszkowski in a paper "Oscillators with automatic control of the threshold of regeneration," Proceedings of the Institute of Radio Engineers for February, 1934. A major advance along this line was made by Meacham in a paper "The bridge stabilized oscillator," Proceedings of the Institute of Radio Engineers for October, 1938, and in his United States Patent 2,163,403, June 20, 1939. His proposed solution contemplated that the vacuum tube should no longer act in any other capacity than that of an amplifier, the amplitude control which must be inherent in any vacuum tube oscillator being achieved by a circuit element exterior to the tube. This should lead to an increase in stability and experience has demonstrated that it does. The principle which he enunciated was applied by himself to a circuit in which the control unit was incorporated as an element of a bridge network which minimized the changes in the oscillating frequency resulting from any residual variations in the gain or phase of the oscillator amplifier. Applicant believes that the Meacham principle is best capable of being utilized to attain maximum frequency and amplitude stability as effected by the various above considerations.

Applicant's invention hereinafter to be described, contemplates an application of the above principle alternative to that specifically disclosed by Meacham. A specific object of the invention therefore is to improve the operation of bridge stabilized oscillators with respect to frequency and amplitude stability, having regard to the various above considerations especially the effect of harmonic reaction.

Since applicant's invention has to do more particularly with comparatively low frequencies and with a means for readily varying the frequency, an object subsidiary to the above object is to adapt the bridge stabilized oscillators of the prior art, having particularly in mind the Meacham circuits, to relatively low frequency and variable frequency operation. Another subsidiary object is to achieve the first-mentioned objects by bridge networks and their immediately associated circuits which comprise types of impedance means which are themselves inherently stable in function, and which, moreover, so cooperate in the achievement of the desired result so as to necessitate a minimum of circuit complexity and plant cost.

This specification will disclose an oscillator which is particularly applicable to low frequencies because only capacitances and resistances are employed. Specifically the bridge is of a type sometimes known as the parallel-T network. The amplitude is controlled by means of a control resistor the resistance of which is responsive to current changes therein. This control resistor may be an element of the network or, as in the species claimed in this divisional application, it may be outside of the network although still in the feedback path and, specifically, in shunt to the network as a whole. This means may be a thermal device, that is a thermistor. The amplitude control employed permits the analysis of the circuits by linear circuit methods. A like parallel-T network has been described by Augustadt in his United States Patent 2,106,785, February 1, 1938, for use as a filter and by Scott in his United States Patent 2,173,427, September 19, 1939, and in his paper beginning in page 226 of the Proceedings of the Institute of Radio Engineers for February, 1938, for use as a selective analyzer and as an element of an oscillator circuit. In the network as used in the present instance, the elements are so altered as to make any feedback path additional to said network unnecessary and, further, as above indicated a current responsive amplitude control feature is provided. This control adjusts the feedback to the proper value so that the tube functions as a linear amplifier, this therefore resulting in better frequency stability.

In the detailed description of the invention following the brief description of the invention identifying the figures of the drawings which exemplify the invention, the general operating conditions for the oscillator will be treated. A theory for the frequency stability for conditions of large amplifier gain and under actual operating conditions will be developed. A comparison of the theoretical results with those achieved with experimental oscillators will be made and a description will be given of an oscillator found practicable in use which was adapted and modified from the experimental model by use of which the above results were secured.

In the accompanying drawings, Fig. 1 is a schematic showing of an oscillator embodying my invention, the form of the showing being adapted less for a teaching of the details of a complete, practical, embodiment of the circuit, than to provide a simple means for teaching the basic principles of the invention and to make possible the simple analytical treatment which follows;

Figs. 2 and 3 illustrate graphically the performance characteristics of the oscillator schematically shown in Fig. 1; and

Figs. 4 and 5 illustrate two alternative forms of the oscillator in practicable embodiments.

In Fig. 1 which epitomizes a complete oscillator circuit, the round trip energy flow path which principally characterizes any self-contained oscillator circuit comprises the feedback and frequency determining network 1 and the amplifier 2. The network 1 is identical in continuity with that disclosed in the above-mentioned United States patent to Augustadt 2,106,785, February 1, 1938, although for practical convenience and to take account of the somewhat different function and emphasis, it is given a somewhat different analytical treatment herein. It is a parallel-T network comprising two individual T networks one comprising series resistances R and the shunt capacitance C_1 and the other comprising series capacitances C and shunt resistance R_1, which shunt resistance is current responsive so as to enable it to function as an amplitude control resistance as will be hereinafter explained. The two networks are connected as shown in common to the input circuit of the amplifier represented by equivalent resistance R_2 and likewise in common to the output circuit of the amplifier represented by the equivalent resistance R_p. The first identified T network is a filter of the low-pass type which transmits direct currents and low frequency alternating currents with a relatively small loss and attenuates high frequency currents. The last-mentioned T network is a high-pass filter which therefore provides a high attenuation to low frequency currents. The two, in combination, constitute a band elimination filter which may be proportioned to substantially suppress the transmission of alternating currents in any selected frequency range and to effect the complete suppression of a selected frequency in that range, as pointed out specifically in the above Augustadt patent.

The following analysis will treat these two T networks as a whole, that is, as constituting parts of a single network 1. Fig. 1 shows, with identifications, the four significant current flow paths, certain of which will be made use of specifically in the analysis. See the arrows indicating currents i_1, i_2, i_3 and i_4. It is evident that currents i_1 and i_4 are respectively the current entering and leaving the network and that currents i_2 and i_3 are conceived as being confined to meshes of the network, each mesh being made up of a part of each of the component T networks. Specifically, the current paths may be traced as follows: current i_1 follows the path from the resistance R_p through the left-hand resistance R, the capacitance C_1, back to R_p; current i_2 follows a path through left-hand resistance R, capacitance C_1, lamp resistance R_1, left-hand capacitance C back to resistance R; current i_3 follows the path from right-hand resistance R through right-hand capacitance C, the lamp resistance R_1, and capacitance C_1 back to resistance R; and current i_4 follows a path from right-hand resistance R, through, the feedback path including input resistance R_2, and through capacitance C_1 back to said right-hand resistance R.

In the above description of the Fig. 1 circuit, the resistance and capacitance elements perhaps could better be denominated as resistors and capacitors. The selection of the literal labels to identify the elements has purposely been made such that they may be used also to represent the impedance values of the elements as far as may be. For example, the impedances (resistances) of the resistors R have values R. The impedances, that is the reactances, of the capacitors C of course contain the values C of the corresponding capacitances and will be indicated in the analysis as X. Correspondingly the reactance of the capacitance element C_1 will be indicated as X_1. It will be assumed that the network is of symmetrical configuration such that the two resistances are equal and likewise the two capacitances are equal. While other relationships may be used, these do not materially alter the properties of the filter, but the symmetrical arrangement is the one that is preferred in practice because of the resultant simplification both as to structure and as to analysis.

It is useful to consider the conditions which lead to a null for transmission through the network. This is for the reason that with a reasonable amplifier gain, the conditions for generation of oscillations will be approximately satisfied only for such null condition, for it is only a frequency near the null that the phase and amplitude requirements for oscillation may be realized. This will be more clearly evident from the discussion of phase and amplitude characteristics of the network given below. An analysis on this basis may provide approximate criteria for the electrical dimensions of the network for a desired operating condition. Also, the expressions which result are useful in understanding the operation of the circuit.

The conditions for such null transmission through the network of Fig. 1, that is for the relation

$$\frac{i_4}{i_1}=0$$

is found to be

$$X=\pm\sqrt{2RR_1},\quad X_1=\pm\frac{R^2}{2\sqrt{2RR_1}} \qquad (1)$$

The plus or minus signs indicate that the same behavior may be obtained with either inductances or capacitances. For low frequency operation it is advantageous to use capacitances since coils having a sufficiently high Q are difficult and expensive to procure at such frequencies. Assuming that all the reactances are capacitative, Equations 1 may be solved for the frequency of the transmission null, which is very nearly the frequency of oscillations. This frequency is given by the Equation 2

$$f_0^2=\frac{1}{2\pi 2_R 2_{C_1}C} \qquad (2)$$

This equation may be easily proved by conventional analysis but it is believed that sufficient occasion for proof, with the considerable detail involved, is not presented by this specification. A like statement may be made for any other item of analysis to be found in this specification where a complete proof is lacking.

By combining the Equations 1 after expressing the values of reactances X and X_1 by the usual formulae, the conditions for the null in terms of the circuit constants is

$$\frac{4R_1C}{RC_1}=1 \qquad (3)$$

Since an exact null is not possible, if oscillations are to occur, that is, since no oscillations could occur if the effective bridge constituted by the parallel-T network were exactly balanced, let Equation 3 be changed to allow some degree of scope in variation from the conditions taught by said Equation 3 so as to read

$$\frac{4R_1C}{RC_1}=K \qquad (3a)$$

It will be shown below that a value of K less than 1 must be used. Also for the null condition, the input impedance of the network, as expressed in complex notation is,

$$Z_{in}=R\left(1-j\sqrt{\frac{2C}{C_1}}\right)\Big/\left(\frac{1+2C}{C_1}\right) \qquad (4)$$

Figs. 2 and 3 represent respectively the attenuation (insertion loss) and phase characteristic (insertion phase shift) of the parallel-T network. The significance of the ordinates is obvious. The abscissae in both instances are plotted in terms of frequency and specifically in terms of the ratio $F=f/f_0$ between a given frequency f and the frequency for the null condition f_0 as expressed in Equation 2. The various curves in Fig. 2 correspond to different selections of the constant K as expressed in Equation 3a. Although the value of K may be obtained by correspondingly varying the electrical dimensions of any of the several component elements, as is apparent from Equation 3a, it is here assumed that its variation is achieved by variations in the value of the shunt resistance R_1. This is partly because, as will be apparent later, as it is apparent from Equation 2, changes in other elements would, undesirably, also change the frequency and because element R_1 is the element consciously included in the network for this purpose, as the amplitude control resistor. It is intended to and does automatically change the balance of the bridge constituted by the parallel-T network to accord with the desired conditions of operation. However, the curves of Fig. 2, and the same is true of Fig. 3, equally well represent the situation if the factor K were otherwise changed without attendant change of frequency. Of course the change of frequency represented by the abscissae contemplates a change in the frequency of waves incident on a given network and not a change in the frequency dimensions of the network itself as expressed in Equation 2 since at this point it is the network per se which is being subjected to analysis and the conditions for its use in a self-oscillatory circuit have not been explored. It is noted from Fig. 2 that the significant frequency is that where it equals the null frequency, that is, where

$$F=1$$

As indicated by Fig. 3 this is the frequency at which there is a phase shift of 180 degrees.

After the introduction to Figs. 2 and 3 provided by the above, it should be noted that for a critical value of the shunt resistance, the network offers an infinite loss at the frequency for which the phase shift is 180 degrees. For all values of this resistance less than the critical value that is for a value of K less than 1, the loss at this frequency point, that is for a phase shift of 180 degrees, is finite and its value depends upon how nearly the critical condition is approached. It is evident from Fig. 3 that, at the critical (significant) frequency, the phase shift passes through 180 degrees for a shunt resistance either equal to or less than the critical value but that for larger resistances, the phase shift passes through zero instead. This is the reason for the statement made above that a value of K which is less than 1 must be used.

In order to make the operation of the circuit automatic, the shunt resistance arm, that is the element R_1, must include an element whose resistance increases with temperature where this element is positioned as in Fig. 1. The cold resistance must be sufficiently low so that the network loss is less than the amplifier gain at the 180-degree phase point, to respond to the basic condition that there must be sufficient amplification to make up the loss in the remaining part of the round trip path so that the oscillations may be capable of being perpetuated. The amplifier phase shift is also assumed to be nearly 180 degrees. From Fig. 2, it may be noted that the network loss around the critical frequency point increases very greatly with increasing shunt resistance. This means that the automatic control function of the resistance R_1 is most effective at this point and therefore most effective at a near balance of the network. Thus, when the circuit is completed as by energizing the amplifier circuit, oscillations will tend to build up, simultaneously increasing the shunt resistance by the heat produced in R_1 until the network loss is just equal to the amplifier gain.

One of the most desirable characteristics of the circuit, as will be apparent from Figs. 2 and 3, is that because of the very critical conditions as to loss and phase shift around the critical frequency point, all harmonics are fed back degeneratively and therefore in such a manner as to reduce the net value over that present with no degenerative feedback. Since the generated harmonics are at a very low level as a result of this degeneration together with the use of the very sensitive amplitude responsive control means, there results a very nearly sinusoidal output.

From Equation 2 it may be seen that the frequency of oscillation may be varied by changing R, C_1, or C. The amplitude of oscillation will be determined by the value of R_1 required to make the loss through the bridge (network) just equal to the gain of the amplifier. This gain should be constant over the frequency range. The value R_1 therefore depends on the circuit constants as given by Equation 3*a*. Thus if the frequency is varied by changing R, C_1 or C individually R_1 would necessarily be a function of the frequency since these quantities occur in Equation 3*a*. This is true both because the value R_1 to produce the given loss in the network would have to be changed and because the input impedance of the network as given by Equation 4 will change with resulting variations in the amplifier gain which must be balanced by the network loss. However, if the frequency is varied by varying the capacitances C_1 and C with a fixed ratio between them and with a constant value of R, the control resistances R_1 will be independent of frequency. This is true with respect to both of the above reasons for deducing that the value of R_1 is a function of frequency if the frequency is varied by changing the frequency significant element individually, as appears from Equations 3 and 4. It may also be demonstrated that the attenuation and phase characteristics of the circuit depend only on the ratio of these capacitances. As a result the output amplitude and frequency stability of the oscillator will be independent of the frequency provided that the amplifier characteristics are independent of the frequency. For some applications it may not be advantageous to vary frequency by capacitance variation. In such a case the frequency may best be varied by a change of resistance R, but in order to do this most effectively, a wide range control resistance R_1 is required and one which is very sensitive to current changes. Such resistances will be discussed in a later section.

An analysis, not here described, has been made which is valid for the limiting condition of very large amplifier gain. In the analysis it was assumed that the oscillator amplifier is a constant current generator and all parasitic capacitances are neglected. While the analysis covered both the asymptotic and extended theories, the expressions found from the extended analysis are cumbersome although indicating the limiting correctness of the asymptotic expressions. These asymptotic expressions will be given here without derivation and deductions will be made from them to illustrate their application for design purposes. The asymptotic analysis of the oscillator shows that

$$\frac{\delta f}{f_0} = -\frac{9 \sin \theta}{\sqrt{2}Rg} \tag{5}$$

which if differentiated with respect to a variation of supply voltage gives

$$\frac{d}{dE}\frac{\delta f}{f_0} = \frac{9}{\sqrt{2}}\frac{\sin \theta}{Rg^2}\frac{dg}{dE} \tag{6}$$

In Equations 5 and 6

f_0 is the frequency at the null
δf is the operating departure from f_0
θ is the amplifier phase shift
g is the transconductance of the amplifier
E is any supply voltage.

It is to be understood that an asymptotic analysis is intended to show the conditions that are approached as the amplifier gain is allowed to increase without limit. Equation 5 expresses the stability of the oscillator and says that the stability increases as R and g are increased and decreases as the phase shift of the amplifier, as might be caused by plate supply, choke coils and the like, departs from 180 degrees.

Equation 6 indicates that the frequency stability should be independent of the frequency of oscillation if R is fixed and should be infinite if the departure of the phase shift of the amplifier from 180 degrees is zero. This independence of frequency stability and frequency on the assumption that R is fixed also assumes, in the formulation of the Equation 5 wherein the quantity R is used in place of a more complex quantity, that the frequency is changed in the manner above described, namely, by varying capacitances C_1 and C while maintaining a constant ratio between them. It also indicates that the greater the transconductance of the tube, the more stable the oscillator should be with respect to frequency. It should be noted that the phase shift of a vacuum tube per se, is necessarily 180 degrees but that the phase shift of an amplifier, which includes a tube as an element, may easily differ from 180 degrees in either direction and by special design of its circuits external to the tube per se could have any phase shift whatever. In deference to convention in the specification hereafter and in the claims, the term "zero phase shift" or the like will be taken to means a departure from this 180-degree phase shift that is inherent in the tube.

The extended analysis, as distinguished from the asymptotic analysis here given, bears out the above statements. Also curves plotted from the equations resulting from the said extended analysis show that, as above, the optimum phase shift approaches zero (having in mind the definition of zero phase shift a little above) as the transconductance becomes very large and the frequency stability tends to become independent of phase shift. Although it has been stated above and as appears from Equation 6, the frequency stability should be infinite if the phase shift of the amplifier is zero, the above-referred to extended analysis shows that this is not strictly true for small values of transconductance. At least for such small values of transconductance, the optimum amplifier phase shift is not quite zero. Accordingly, it is more nearly correct to say that the frequency stability should be a maximum when there is a condition both of zero amplifier phase shift and very large transconductance.

In order to test the foregoing theory an experimental oscillator was built according to the schematic circuit of Fig. 1. This circuit will be explained in detail later. The control element was composed of twelve Western Electric C-2 switchboard lamps in series, since these lamps were found to have the best characteristics of any available control lamps of this type. The frequency was changed by changing the three capacitances simultaneously with the resistance R fixed. This variation resulted quite closely in the maintenance of a constant ratio between C and C_1 as was desired according to the above analysis.

The amplitude of the oscillations was not absolutely constant with the variation of frequency from 50 to 20,000 cycles although it varied only about 2 decibels. This variation was probably due to phase shift in the amplifier or perhaps to some progressive unbalance in the tuning capacitances. The frequency stability was found to become more nearly independent of frequency as the frequency increased. Over a large part of this range, the stability was as low as about six parts in a million. The frequency stability was with reference to changes in plate voltage supply. The slight deviation from predicted performance may be ascribed to phase shift in the amplifier at low frequency due to the plate supply choke coil, cathode biased network and the like.

Similar experiments were performed to determine the effect of transconductance of the amplifier tube on the frequency stability as effected by changes in plate supply. The transconductance was varied by use of a potentiometer grid leak. Consistently with theory, the frequency stability increased with the transconductance and in fact became a linear function thereof at large values of transconductance.

A similar experimental test was made to demonstrate the frequency stability, as effected by the plate supply voltage as before, as a function of the amplifier phase shift. The amplifier phase shift was assumed to be zero in the normal operating state and phase shift was then produced by shunting the input of the network with reactance. As was to be expected and consistently with the above analysis the stability was found to be a maximum at very nearly zero phase shift and, specifically, and likewise consistently with the theory, when there was a slight lagging phase shift. At the frequency of oscillation, a capacitive reactance appears between the input terminals of the network, as indicated by Equation 4. Therefore, a shunting inductance at this point would tend to increase the amplifier gain and a capacitance would tend to reduce the gain. In any event, it would be difficult to cause a phase shift without an attendant change in gain. However, the test was informative and the use of a shunt capacitance at that point to produce the optimum phase shift was found to offer a convenient and simple method of obtaining high frequency stability even with a comparatively low gain amplifier. Such a shunt capacitance would tend to be most useful for a fixed frequency oscillator since it would be necessary to vary the capacitance inversely with frequency for a variable oscillator. However, for a variable frequency oscillator, the desirable characteristics of an improving frequency stability in the frequency increases may be obtained by shunting the network with a capacitance which will produce the optimum condition at the maximum frequency.

Fig. 4 illustrates a physical embodiment of an oscillator of the invention as constructed for general laboratory purposes and as was used for the above-described tests. The elements of the oscillator which are duplicates of the oscillator illustrated schematically in Fig. 1 have like identifications. Since the feedback network is illustrated in almost complete detail in Fig. 1, the principal additional showing by this Fig. 4 is with reference to the amplifier element. The three capacitances in the network should be equal and, as shown, are caused to vary together simultaneously to change the frequency. This satisfies the condition earlier stated that the ratio of capacitances C and C_1 should be kept constant when the frequency is changed by corresponding change in said capacitances to insure that the control resistance R_1 is independent of frequency. Since said resistance R_1 is independent of frequency, it may be adjusted to remain at the steepest point of its current-resistance characteristic, this resulting in the most sensitive amplitude control. In the oscillator as used this resistance R_1 comprises several Western Electric C-2 lamps in series. These lamps have tungsten filaments and a positive temperature coefficient of resistance. Such resistance constitutes a thermistor since the effect is that of a change in resistance as effected by a change in thermal condition which is in turn induced by the change of the current traversing it. The oscillator is adapted to be varied in frequency from about 40 cycles to 50 kilocycles. The theoretical characteristics of such an oscillator have been indicated in the earlier part of the specification and the operational characteristics of an oscillator corresponding strictly to this particular design have been indicated more recently. The harmonic content of the generated wave was found to be principally the second harmonic and was about 50 decibels below the fundamental over the entire frequency range so that the purity of the output wave was most satisfactory. This result is largely attributable, as previously mentioned, to the linear operation of the amplifier and the feedback properties of the network.

The amplifier element of the above oscillator is effectively a single tube amplifier, that is, the amplification occurs in a single stage rather than in a plurality of stages. For practical reasons, having to do with practical convenience and the obtaining of sufficient power, two tubes **3** and **4** are used, these tubes being in effectively parallel relationship. The various resistances in circuit with the electrodes of the respective tubes have the functions usually attributable to resistances in amplifier tube circuits generally, as well as insuring a proper balance and matching between the two tubes here adapted to function in parallel. The points **5** and **6** may, for practical purposes, be treated as the input terminals of the network and the output terminals of the amplifier since ground point **5** is effectively at cathode potential. The output terminals of the network, and correspondingly the input terminals of the amplifier, are at points **5** and **6**a. The tubes are given the requisite degree of negative feedback to accord with the above-indicated design conditions of the tubes by networks **7** and **8** in the respective cathode leads. The load circuit **9** is connected between point **6** of the network and ground **5** and therefore effectively across the input circuit of the network. The various impedance elements illustrated, exclusive of the resistances in the tube electrode leads which have been separately mentioned above, are conventional in character and function and do not merit specific description.

The use of the tungsten filament lamps for the amplitude control resistance of Fig. 4 places some restrictions upon the oscillator circuit which lead to the use, in some instances, of the alternative circuit arrangement of Fig. 5 presently to be described and claimed specifically herein. The best lamps commercially obtainable require about 2 milliamperes of current flowing through them in order to operate on a steep portion of their resistance-current characteristic. That statement is with reference particularly to the Western Electric C–2 switchboard lamps described as having been used in the circuit of Fig. 4, which type of control resistance means was found to be most effective in practice.

For a permissible plate swing of the oscillator, this limits both the shunt and series resistance arms of the network if the capacitances of the network are to be equal. An inspection of the Equation 5 shows that this tends to restrict the stability of the oscillator, which physically means that an upper limit is imposed by the possible amplifier gain. If the shunt and series capacitances are not made equal, more amplifier gain may be attained but this requires more capacitance to cover a given frequency range, if frequency changes are produced by simultaneous variation of the capacitances.

A solution to the problem of obtaining a highly sensitive control is by fixing R_1 at the largest value possible for the range desired and connecting a resistance having a very large negative temperature coefficient between the grid and the plate of the amplifier as a separate feedback path. This principle will be found to be utilized in the circuit of Fig. 5 to be described later. These negative resistance coefficient thermistors are obtainable in a wide range of impedance and current ratings and have a very high rate of change of resistance with current. Their simplicity and compactness, also, makes them highly useful for circuit applications. The types here had in mind are described in a paper by G. L. Pearson in the Physical Review, volume 57, June 1, 1941, beginning in page 1065. Examples of these thermistors are what are sometimes known as "bead thermistors." Frequently, the "bead" comprises silver sulphide. As used in the present considered circuit as above described the thermistor would have a very high negative temperature coefficient and would be of the high speed type. It would incidentally tend to have a high resistance. In view of the Pearson paper, it is not believed to be necessary to treat the specific characteristics of such thermistors in this specification.

An accurate calculation of the required thermistor resistance is difficult. However, an approximation for design purposes may be calculated by assuming that the impedance looking back into the network from the grid is the value corresponding to a null. Then, as a particular example, given an amplifier gain of 40 decibels, a network loss of 38 decibels, an input impedance at the null of 10,000 $(1-j\sqrt{2})$ and a grid leak resistance of 100,000 ohms, if oscillations are to begin the thermistor impedance must be greater than 4.53 megohms. If the network loss were 36 decibels, the required initial thermistor resistance would drop to 1.32 megohms. As the fixed network loss is made to approach the gain of the amplifier in order to realize the maximum frequency stability, the thermistor resistance at the operating point must increase very rapidly. This tends to result in poorer amplitude stability for it will be apparent that a small perturbation in amplifier gain caused by the voltage variation or the like requires a large change of the thermistor resistance. A compromise between amplitude and frequency stability must therefore be made. In the oscillators described, a K of about .85 was used, resulting in an operating thermistor resistance of a few hundred thousand ohms. A good amplitude control is aided by using the constants of the circuit and of the thermistor to obtain operation in the region where a large change in resistance is effected by a small change in voltage. This may be further improved by the addition of a fixed resistance. If a sufficiently high resistance thermistor is not obtainable, or is otherwise unsuitable, some of the required loss through the thermistor bridge may be taken by a linear resistance attenuator following the thermistor.

The use of these negative temperature coefficient thermistors makes possible an alternative method of control for this oscillator. An example is illustrated by Fig. 5 in the circuit of which the frequency may be changed by varying the series resistance arms simultaneously. As has been pointed out the impedance of the network varies as the resistance, which in turn varies the gain, but with the wide range thermistors control over a 10 to 1 frequency range is readily possible. Since the shunt resistance requires a complementary adjustment it is contemplated that all three resistances be ganged together and be continuously variable over a considerable range. For practical purposes, this system of frequency variation may be superposed on the system applied in the circuit of Fig. 4 wherein the three capacitances are similarly ganged to vary the frequency. Practically it is of convenience to vary the frequency over a 10 to 1 range with the variable resistances and to vary the range by variation of the capacitances together in decade steps. In this way, in an experimental oscillator, a range of from 12 to 50,000 cycles was readily obtained in four steps. This amplitude control made it possible to reduce further the harmonic content of the oscillator circuit to a maximum value of 60 decibels below the fundamental at the high frequency end of each range and about 80 decibels at the low frequency end. These values could be further reduced if needed by choosing a thermistor to control at a lower amplitude. The incorporated feedback amplifier, however, had a harmonic content 60 decibels below the fundamental at maximum output.

In Fig. 5, illustrating the alternative circuit as above, the only significant difference over the circuit of Fig. 4 relates to the network and thermistor as above. Accordingly, no labelling is applied to the amplifier element. Since the shunt resistance is no longer an amplitude control means, but only a variable resistance, it is labeled R similarly as the series resistances with which it is ganged. The amplitude control resistance R_1 is shown, consistently with the above description, connected in circuit between the grid and plate of the parallel connected tubes.

Other modifications of the invention than above described, will occur to a person skilled in the art, and all such are considered to fall within the spirit and scope of the invention, as defined in the appended claims.

What is claimed is:

1. An oscillator comprising an amplifier, input and output circuits therefor, a network connected

between said circuits, said network comprising a pair of symmetrical T-networks connected in separate paths between said 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 proportioned to provide maximum attenuation and phase reversal in the network at substantially the preassigned desired frequency, a current-responsive thermistor having a negative temperature coefficient of resistance connected across the series impedance elements of said T-networks and constituting with said T-networks the sole external coupling between said input and output circuits, and means for continuously varying the frequency of the generated oscillations, comprising means coupling the resistances of said networks and adapted to cause them to be simultaneously varied and similar means coupling the capacitances of said networks and adapted to cause them to be likewise simultaneously varied.

2. An oscillator comprising an amplifier, input and output circuits therefor, a network connected between said circuits, said network comprising a pair of symmetrical T-networks each comprising three impedance elements connected in separate paths between said circuits, the impedance elements of one of said networks consisting of two series resistances and a shunt capacitance, and the impedance elements of the other of said networks consisting of two series capacitances and a shunt resistance, said capacitances and resistances being proportioned to provide maximum attenuation and phase reversal in the network at substantially the preassigned desired frequency, said shunt resistance also having the largest value possible for operation over the desired frequency range, a current responsive thermistor having a very large negative temperature coefficient connected across the series impedence elements of said T-networks and constituting with said T-networks the sole external coupling between said circuits, and means acting unitarily on all the impedance elements consisting of at least one of the two kinds of impedance of said network for continuously adjusting the frequency of the generated oscillations.

3. The combination recited in claim 2 in which the two series elements of each T-network are equal, and in which the three elements consisting of at least one of the two kinds of impedance are variable and initially adjusted and ganged so that the series impedances thereof continue equal in value at all times and have a constant ratio to the shunt impedance, whereby the frequency may be adjusted independently of an amplitude adjustment by adjustment of the other shunt impedance, said other shunt impedance being capable of such adjustment for a desired amplitude.

4. The combination recited in claim 2 in which the series impedances of each T-network are equal and the three impedance elements consisting of at least one of the two kinds of impedance are variable and initially adjusted and ganged so that when varied together the impedances of the series elements maintain equality and a constant relation to the impedance of the corresponding shunt element, whereby the frequency may be changed in such manner that the control of amplitude by the other shunt impedance is independent of frequency, said other shunt impedance being made adjustable for this purpose, the values of all of the impedances of the parallel T-network as a whole being initially adjusted so that the ratio of four times the product of the shunt resistance and one of said series capacitances to the product of one of said series resistances and said shunt capacitance is approximate to but less than 1.

5. The combination recited in claim 2 in which the impedances of the series elements in each said T-network are equal and the three impedance elements of each group comprising like impedance elements are variable and initially adjusted and ganged so that when varied together the impedances of the series elements maintains equality and a constant relation to the impedance of the shunt element, thereby providing two independent means for frequency variation while making possible the adjustment of amplitude by the initial adjustment of the shunt resistance, the values of all of the resistances and capacitances being initially adjusted also so that the ratio of four times the product of said resistance and one of said series capacitances to the product of one of said series resistances and said shunt capacitance is approximate to but less than 1.

6. An oscillator comprising an amplifier, input and output circuits therefor, a network connected between said circuits, said network comprising a pair of parallel symmetrical T-networks each comprising three impedance elements, one of said networks consisting of two series resistances and a shunt capacitance, and the other of said networks consisting of two series capacitances and a shunt resistance, said capacitances and resistances being proportioned to provide maximum attenuation and phase reversal in the network at substantially a preassigned desired frequency, and a current-responsive thermistor having a large negative temperature coefficient of resistance connected across the series impedance elements of said T-networks and constituting with said T-networks the sole external coupling between said circuits, said amplifier having substantially zero phase shift and a very large transconductance, and in which the impedances of the two series elements of each T-network are equal and the three impedance elements consisting of at least one of the two kinds of impedance are variable and initially adjusted and ganged so that the impedances of the series element continue equal in value at all times and have a constant ratio to the impedance of the shunt element whereby the frequency may be adjusted independently of an amplitude adjustment by variation of the impedance of the other shunt element, the impedance of said other shunt element being so adjustable for amplitude change.

7. The combination recited in claim 6 in which the impedances of the three elements of each group comprising like impedance elements are variable and initially adjusted and ganged so that the impedances of the series elements continue equal in value at all times and have a constant ratio to the impedance of the shunt element of the group, whereby the frequency may be adjusted by two independent means independently of the amplitude effect of an adjustment of the impedance of the shunt element of either T-network, the impedance of the shunt element of at least one said T-network being initially adjusted for the desired amplitude value.

RAYMOND O. WISE.