June 2, 1942. J. M. WEST 2,284,855

AMPLIFIER CIRCUITS

Filed Aug. 16, 1939

FIG. 1

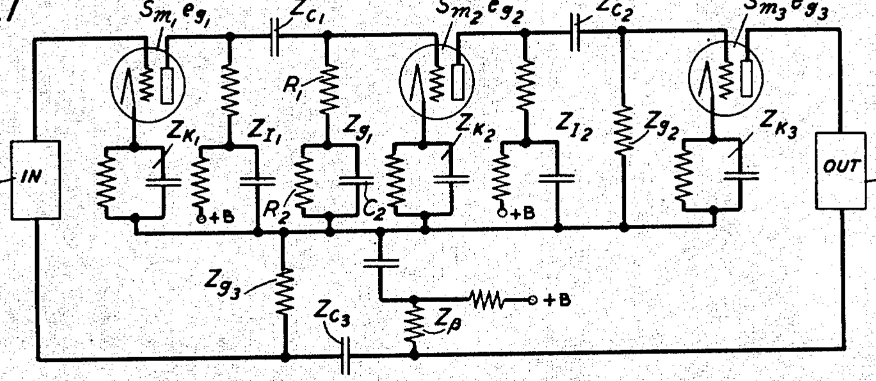


FIG. 2

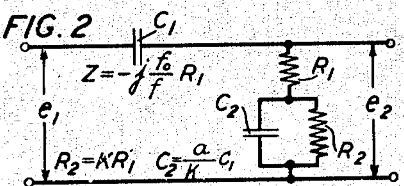


FIG. 3

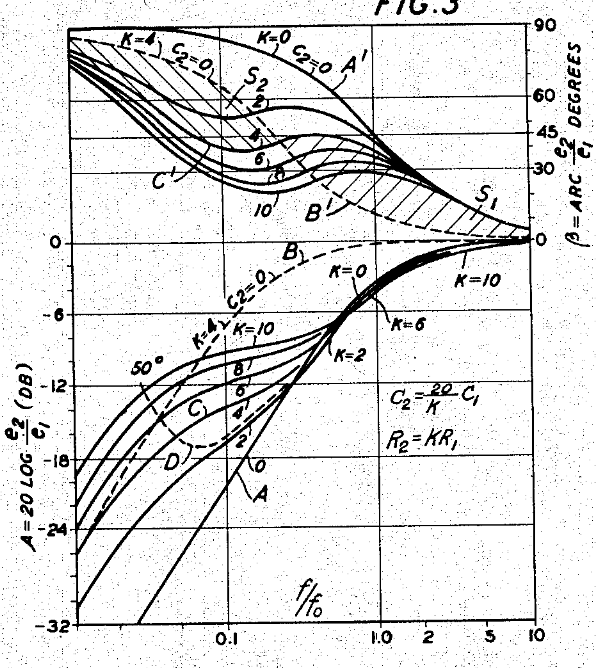


FIG. 4

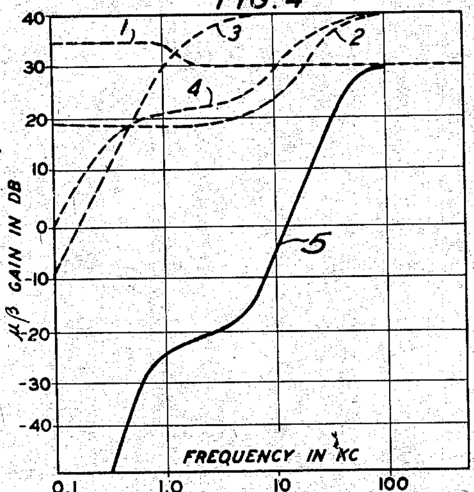
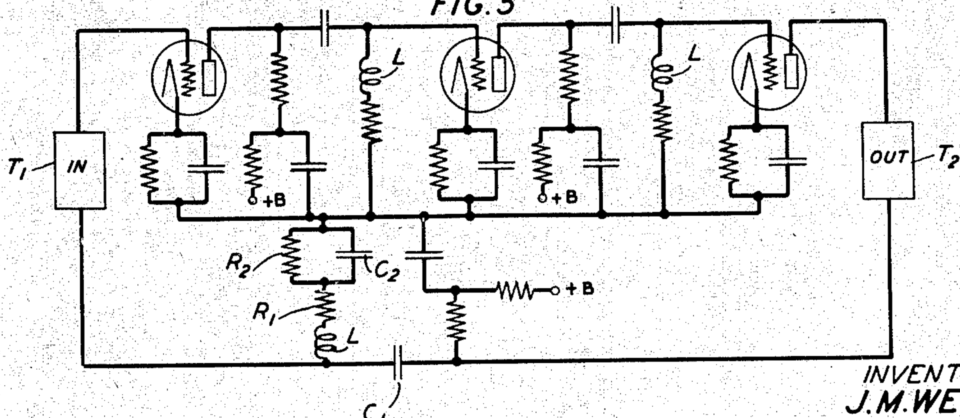


FIG. 5



INVENTOR
J. M. WEST

BY
H. A. Burgess
ATTORNEY

Patented June 2, 1942

2,284,855

UNITED STATES PATENT OFFICE

2,284,855

AMPLIFIER CIRCUITS

Julian M. West, Ridgewood, N. J., assignor to Bell Telephone Laboratories, Incorporated, New York, N. Y., a corporation of New York

Application August 16, 1939, Serial No. 290,354

8 Claims. (Cl. 179—171)

This invention relates to amplifier circuits and more particularly to negative feedback amplifiers so designed as to be effective over a wide band of frequencies. Its object is to provide for a high and uniform internal loop or $\mu\beta$ gain over the whole band of frequencies to be transmitted while still maintaining a substantial margin of safety against singing. More specifically the invention relates to so modifying the cut-off characteristic on the lower side of a band to be transmitted with a negative feedback amplifier that the gain around the feedback loop has a high value over a band to be transmitted and falls to less than unity before a phase shift is introduced which would produce singing.

The benefits to be derived from the use of feedback are described in an article by H. S. Black on "Stabilized feedback amplifiers," Bell System Technical Journal, January 1934. Principal among these benefits are the stabilization of the effective amplification in the presence of variations of the energizing potentials applied to the amplifier vacuum tube and the diminution of the effects of non-linear distortion. These benefits are obtained when the magnitude and phase of the voltage feedback to the amplifier input are such as to produce a diminution of the effective amplification, and the extent to which the benefits are realized is substantially proportional to the amount by which the amplification is diminished.

The degenerative action is provided by so designing the amplifier and the feedback circuit that, when the frequency has some convenient value near the middle of the desired operating range, the effective input voltage is in exact phase opposition to the feedback voltage. The effective input voltage is reduced in comparison with the voltage from the input wave source and the amplification of the system is diminished in proportion. As the frequency departs from this near mid-point value the phase of the feedback voltage changes progressively until at some frequency, usually outside the operating range, it comes into phase coincidence with the effective input voltage. At this point the amplifier may become unstable and develop oscillations. This is most likely to occur at frequencies above the operating range but if suitable precautions have been taken to avoid this then it may occur at some frequency below the operating range. Whether the circuit develops oscillations or not depends upon the relative values of the feedback and the input voltage. If with such phase shift as to bring about phase coincidence the feedback voltage is equal to or greater than the effective input voltage the amplifier is unstable but if the feedback voltage is less than the input voltage, there is no instability. An exceptional case is illustrated in Nyquist Patent 1,915,440 of June 27, 1933. However, the type of stability therein disclosed is not absolute and such systems tend to become unstable when the amplifier gain is reduced.

To insure complete stability of the amplifier against singing it is, therefore, necessary to so proportion the different parts of the circuit that before the condition of phase coincidence is reached, the feedback voltage shall have been attenuated to a value less than the effective input voltage. This requirement may be expressed in terms of certain transmission parameters of the circuit as follows: If the voltage gain or transfer ratio from the input to the output of the amplifier be denoted by μ and the voltage transfer ratio of the feedback path in the reverse direction by β then the vector ratio of the feedback voltage to the resultant input voltage on the amplifier is equal to the product $\mu\beta$ where $\mu\beta$ may have a large value over a wide portion of the frequency spectrum. For complete stability, however, it is necessary that the magnitude of the $\mu\beta$ become less than unity before its phase angle becomes zero.

The vector quantity $\mu\beta$, the real part of which is hereinafter called the loop gain, defines the net change in amplitude and phase experienced by a voltage wave in traversing the amplifier and the feedback path in sequence. This change is made up essentially of two parts, first a constant amplification and a constant phase shift representing the effect of the amplification factors of the vacuum tubes and, second, a variable diminution of the amplitude and a variable phase shift representing the transmission loss in the passive networks of the system. Since the part contributed by the vacuum tubes is constant, all of the variation of $\mu\beta$ is represented by the variation of the transmission loss of the passive networks. These passive networks include, of course, not only the circuit from the output of the last stage to the input of a previous stage but include also the passive coupling and other networks associated with and between the tubes of the amplifier.

By using pure resistances for the feedback and coupling networks it would be possible theoretically to make $\mu\beta$ constant in both amplitude and phase at all frequencies and thereby provide for any desired amount of feedback and an unlimited operating frequency range. In practice, however, this is impossible because of the presence of unavoidable parasitic reactances in the system. At high frequencies these parasitic reactances consist primarily of shunt capacitance, such as those within the tube, or of series inductances; and at low frequencies they consist primarily of series capacitance and shunt inductances. At very high and very low frequencies these parasitic reactances become the sole factors determining the magnitude and the variation of the feedback and are generally productive of phase changes sufficiently great to cause instability. To avoid this it is necessary that the magnitude of the feedback should be reduced to less than unity before the limiting frequency ranges are reached and suitable frequency intervals must be allowed for this reduction.

It is desirable that the cut-off ranges in which the magnitude of the feedback is systematically reduced should be as small as possible in order that the greatest operating frequency range may be conserved. It can be shown that the phase shift component of $\mu\beta$ is greatly influenced by the course that its magnitude follows in the cut-off range and by the rate at which the magnitude diminishes. In the amplifier of this invention the diminution of $\mu\beta$ is made to follow certain preferred courses which represent optimum characteristics in the sense that they permit the use of maximum amounts of feedback and involve the minimum loss of useful frequency range while at the same time insuring complete stability against singing.

The invention will be better understood by reference to the following specification and the accompanying drawing in which:

Fig. 1 is a simplified diagram of a circuit incorporating my invention;

Fig. 2 is a detail of part of the network of Fig. 1;

Figs. 3 and 4 show curves which will explain the manner in which the invention is carried out; and

Fig. 5 shows a modified circuit incorporating my invention.

The general theory underlying the behavior of such circuits as are shown in Fig. 1 is set forth in some detail in the patent to Bode 2,123,178 of July 12, 1938, and need not be included here. The invention therein disclosed is of a broad nature and in one of its specific forms illustrates the design of a feedback circuit which will appreciably improve the cut-off characteristic on the upper side of the transmitted band while still maintaining a safe margin against singing. In this invention the purpose is, specifically, to improve the cut-off characteristics on the lower side of the transmitted band without deleteriously affecting the upper limit and while still maintaining the margin of safety against singing.

One of the advantages which I find accruing from the redesign of the $\mu\beta$ network in accordance with my invention is a large reduction in the size of some of the elements associated with this circuit. This is particularly true of the so-called blocking condensers between adjacent tubes, these in some instances being reduced by a factor of as much as 100. This reduction not only contributes to economy of cost and economy of space, but also contributes to a reduction of the parasitic capacities effective for the high frequencies which are inherent to a greater or less extent in such circuits.

As pointed out in the patent to Bode and elsewhere above it is essential that the phase shift through the $\mu\beta$ circuit shall not exceed 360 degrees. Since for an odd number of amplifier stages, such as in Fig. 1, there will be a net phase change of 180 degrees due to the amplifier tubes themselves, it is evident that the phase shift over the passive networks of the $\mu\beta$ loop must not be allowed to exceed 180 degrees as long as $\mu\beta$ is greater than unity. The extent to which it falls below this at any frequency is a measure of the margin against singing at that frequency. A full understanding of the behavior of a given circuit requires then, among other things, a knowledge of the $\mu\beta$ gain around the loop and a knowledge of the phase shift both as functions of frequency.

In the circuit of Fig. 1, there are shown three stages of amplification, with the usual batteries omitted for simplicity, and associated with the three tubes are cathode networks Z_{K_1}, Z_{K_2}, Z_{K_3}, primarily for establishing a self-bias. There are also the load circuits Z_{I_1}, Z_{I_2}, Z_β, the latter also including the output transformer T_2. There are condenser-resistance coupling between stages 1 and 2, between 2 and 3 and between 3 and 1, the last constituting essentially the β or feedback circuit. The load circuits are shown as resistive with a resistance-condenser filter, all well known in the art, but it is to be understood that these may take on the form of any suitable impedance.

At any one frequency one may find the insertion loss due to each of these networks and find the phase shift due to each network. The total loss, preferably expressed on a decibel scale, and the total phase shift due to the passive part of the $\mu\beta$ circuit will be the sum of these. If this shift is near to but below 180 degrees, then one may set the total gain due to the tubes equal to the total loss of the coupling and other networks at that frequency so that the net gain around the loop at this frequency is equal to unity. If for any frequency the shift is substantially below 180 degrees then the gain may be correspondingly raised and the amount of feedback correspondingly increased, that is, the insertion loss due to the passive networks may be reduced. However, such a change usually brings in an additional change in phase so that the permissible increase in gain is limited. It is essential that at no part of the frequency spectrum where the phase shift due to these networks exceeds 180 degrees shall the magnitude of $\mu\beta$ exceed unity.

While the three coupling circuits are shown as much the same as condenser-resistance combinations they will, in general, be of different impedance. In this figure, moreover, it will be noted that one of the couplings, namely, that between stages 1 and 2, has the resistive portion made up of two series resistances R_1 and R_2, the latter of which is shunted by a condenser C_2, these being proportioned as will be described below, the combination and proportioning in this coupling being illustrative of and constituting an important part of my invention.

For a better understanding of the ideas underlying this invention reference may be made to Fig. 2 showing a four-terminal network corresponding to the coupling circuit just mentioned. If e_1 is the impressed and e_2 the output voltage over this coupling circuit, then e_2/e_1 is the loss or attenuation or the transfer ratio due to the network and this will be a function of frequency. This attenuation may conveniently be expressed in decibels by using the logarithm of the real part of e_2/e_1 in the relation.

$$A = 20 \log \left| \frac{e_2}{e_1} \right| \text{ (db)} \qquad (1)$$

Also in the analysis of the network it will be convenient to express R_1 in terms of an arbitrary unit such that R_1 is taken as unity and to represent R_2 by KR_1. It will also be convenient to express the impedance of the condenser c_1 in terms of a reference capacitance c_0 whose impedance at a reference frequency f_0 is

$$-j\frac{1}{2\pi f_0 c_0} = -jR_1 \qquad (2)$$

Then at any other frequency, taking $R_1=1$, the impedance of c_0 is

$$-j\frac{f_0}{f}$$

The capacitance of c_2 represented by

$$\frac{a}{K}c_1$$

will be taken, for illustrative purposes,

$$\frac{20}{K}c_1$$

For $K=0$ (resistance $=R_1$ and $c_2=0$) the attenuation is given by curve A of Fig. 3. For $K=4$, but still keeping $c_2=0$, the attenuation is decreased, the curve being shifted toward lower frequencies as shown by curve B. If, on the other hand, the grid leak is made up of R_1 and KR_1, the latter shunted by capacitance c_2, then the attenuation curve may with proper proportioning take the form of curve C. This is evident, for at high frequencies the portion KR_1 is virtually shorted by c_2 so that the effective grid impedance is R_1 and the attenuation is relatively high. As the frequency lowers, the impedance of c_1 increases, which would increase the attenuation but the impedance of c_2 also increases, thus retarding the increase of attenuation. At still lower frequencies the impedance of c_1 becomes controlling and the attenuation curve approaches curve B.

The effect of the shunt capacity c_2 is then to increase the loss at higher frequencies but at low frequencies it becomes practically ineffective. At intermediate frequencies there is a transition from curve A to curve B with a portion the slope of which is substantially reduced.

Curve C correspondents to a value of $K=4$ but a series of curves for other values of K are also shown and in every case it will be noted that the effect of c_2 is to decrease the slope of the attenuation curve for the intermediate frequencies below that for which c_2 is virtually a short circuit KR_1. Other similar families of curves could be plotted for other values of c_1.

These changes in the form of the attenuation curves are accompanied by changes in the phase shift curves as shown in the upper portion of Fig. 3. Curves A′, B′ and C′ correspond to curves A, B and C. From curve B′ it is seen that the mere increase of K (or, equivalently, the mere increase of c_1) delays somewhat the building up of phase shift as one goes to lower frequencies, but the form of the curve is not altered. The addition of capacitance c_2 to the circuit corresponding to curves B and B′ brings about, however, a substantial change in form, characterized by a substantial increase in phase shift in the higher frequencies, but a substantial decrease at the lower frequencies. In fact, it can be shown that the area s_1 for the region of increase of phase shift taken to infinite frequency is equal to the area s_2 for the region of decrease of phase shift taken to zero frequency.

It is this phenomenon that I use in my invention. The delay in building up of phase shift as one goes to lower frequencies permits a substantial increase in the attenuation before the critical phase shift of 180 degrees is reached. This in turn permits a larger amount of gain by by means of the tubes and thus a larger value of $\mu\beta$ over the transmission band, there still being a fall in value of $\mu\beta$ to less than unity before a phase shift of 180 degrees is brought about.

By direct transfer from the set of phase shift to the set of attenuation curves of Fig. 3

$$\left(\text{for which } c_2 = \frac{20}{K}c_1\right)$$

one can readily, for any assumed value of phase shift of a single coupling network of this type, plot a curve showing the amount of loss due to the network for different values of K. Such a curve is shown at D for $\varphi=50$ degrees and it will be observed that there is a maximum loss of about 17 decibels at $K=2.5$ and that this occurs at about

$$\frac{f}{f_0} = .06$$

If now for a total maximum phase shift of the combined passing networks of less than 180 degrees it is agreed that 50 degrees of the phase shift may be permitted in one of the coupling networks, then it is seen from curve D that the maximum insertion loss for such a coupling network will be obtained for a value of $K=2.5$. A comparison with the corresponding case $c_2=0$ shows that the loss for $\varphi=50$ degrees, amounts to only 4 decibels. Similar curves could be plotted for other values of φ and some of the results are tabulated below. From this one obtains additional information for the design of the amplifier circuits:

ϕ	K	Maximum loss in decibels	Maximum loss for $c_2=0$
40°	4	14	3
50°	2.5	17	4
60°	1+	21	6
70°	1	24	9

Supplementing the theory presented in the Bode patent referred to above, it may be here pointed out, without derivation, that the expression for the $\mu\beta$ gain for the circuit of Fig. 1 in the low frequency range where the high frequency parasitic capacities may be neglected is given by

$$\mu\beta = \left[\frac{1}{1+S_{m_1}Z_{K_1}} \cdot S_{m_1}Z_{I_1} \cdot \frac{Z_{g_1}}{Z_{g_1}+Zc_1+Z_{I_1}}\right] \left[\frac{1}{1+S_{m_2}Z_{K_2}} \cdot S_{m_2}Z_{I_2} \cdot \frac{Z_{g_2}}{Z_{g_2}+Zc_2+Z_{I_2}}\right] \left[\frac{1}{1+S_{m_3}Z_{K_3}} \cdot S_{m_3}Z_{\beta} \cdot \frac{Z_{g_3}}{Z_{g_3}+Zc_3+Z_{\beta}}\right] \qquad (3)$$

where S_m is the transconductance of a tube. The only assumptions involved are that the internal grid and plate resistances are substantially infinite and that the source of B supply is of zero impedance.

A convenient design procedure is to break this expression for $\mu\beta$ into four factors as follows:

$$\text{1st} \quad (S_{m_1}Z_{I_1}.S_{m_2}Z_{I_2}.S_{m_3}Z_\beta) \tag{4}$$

To a very close approximation this is the only significant term at the middle of the useful band of the amplifier. It, therefore, gives the nominal $\mu\beta$ gain. The use of condenser-resistance filters in the B supply leads causes this gain to rise a few decibels at very low frequencies. The effect of this factor on the $\mu\beta$ gain of the circuit is shown by curve 1 of Fig. 4 plotted with reference to a nominal gain of 40 decibels and against the logarithm of frequency.

$$\text{2nd} \quad \frac{1}{(1+S_{m_1}Z_{K_1})(1+S_{m_2}Z_{K_2})(1+S_{m_3}Z_{K_3})} \tag{5}$$

This term represents the effect of the local feedback produced by the resistances used for self-biasing the tubes. In general, this term contributes a major part of the low frequency cut-off. For example, with one type of tube known as Western Electric 7650P tube and the value of resistance to give the correct bias voltage, the gain is reduced 21 decibels at low frequencies by this factor. The effect of this factor on the $\mu\beta$ gain is shown by curve 2 of Fig. 4.

$$\text{3rd} \quad \left[\frac{Z_{g_2}}{Z_{g_2}+Z_{C_2}+Z_{I_2}}\cdot\frac{Z_{g_3}}{Z_{g_3}+Z_{C_3}+Z_\beta}\right] \tag{6}$$

This term represents the effect of any two of the simple blocking condenser grid leak combinations. The blocking condensers are preferably made so small that this factor begins to reduce the $\mu\beta$ gain at the point where the 2nd factor becomes less effective. The effect of this factor on the $\mu\beta$ gain is shown by curve 3 of Fig. 4, reaching a slope of 12 decibels per octave.

$$\text{4th} \quad \left[\frac{Z_{g_1}}{Z_{g_1}+Z_{C_1}+Z_{I_1}}\right] \tag{7}$$

This term represents the effect of the remaining grid blocking condenser grid leak combination with shunt capacitance, the results being shown by curve 4 of Fig. 4 and corresponding closely in form to the attenuation curves of Fig. 3. It is this configuration that I find effective in the low frequency design. If the coupling network to which this factor relates had been of the normal blocking condenser grid leak combination, the gain throughout the circuit due to this coupling would have been reduced only 3 decibels when the phase change introduced by it reaches 45 degrees. With this combination, however, one finds that the gain is reduced about 15 decibels before appreciably exceeding 45 degrees. Curve 5 of Fig. 4 represents the attenuation curve for the whole passive circuit, this being made up of the sum of the individual attenuation curves.

In view of the delay in the building up of the phase shift which accompanies the introduction of the capacitance c_2 it becomes evident that the condenser c_1 may be substantially reduced without introducing an excessive additional phase shift. This is a marked advantage in that the use of a smaller condenser is cheaper and more economical of space, but still further because the reduction in the physical dimensions of the condenser c_1 reduces its capacity to ground, which latter contributes to the parasitic capacities at high frequencies. Thus the combination reacts favorably on the cut-off characteristic at the upper side of the frequency band to be transmitted. In practice I find that this capacity may be decreased by a factor of as much as 10 or 100.

While the description thus far has been on the basis of altering the coupling network between stages 1 and 2 of Fig. 1, it should be pointed out that the total alteration desired may be accomplished at one step as described or may be distributed in any desired portions among the plurality of coupling circuits. Certain advantages would accrue from such distribution, for each of the blocking condensers could then be reduced with a corresponding reduction in parasitic capacities for high frequencies. On the other hand, it will be noted that such distribution will call for a larger number of elements, represented by the plurality of capacitances c_2. Depending upon the circumstances, therefore, it may be advantageous in the one case to distribute the adjusting factors and in some other cases to confine it to one of the coupling networks. Also, while in the description the coupling network has been shown between stages 1 and 2 of Fig. 1, there would be certain advantages at times in placing it in one of the other positions. In some cases, for example, I would prefer to place it in the β circuit representing the coupling between stages 3 and 1 (see Fig. 5). Such a location has a certain advantage for in general the impedance Z_β in the output circuit would be small compared to Z_{I_1} and Z_{I_2}. In that case the resistance R_1 of Fig. 2 may be made appreciably smaller without this portion of the circuit acting as an appreciable shunt to Z_β.

Thus far in the description reference has been largely to pure ohmic resistances. It is to be understood, however, that the impedances R_1 and KR_1 of Fig. 2, as well as the resistances of the other coupling circuits and of the loads, may be reactive. As a matter of fact, I find it advantageous at times to include with R_1 a suitable amount of inductive reactance as shown in Fig. 5. The principles described above hold equally well and the application of the invention to any one particular case would then involve the plotting of attenuation and phase shifting curves which involve more, but only slightly more, labor.

What is claimed is:

1. A wave amplifying system comprising a plurality of vacuum tube amplifying devices, impedance networks coupling said devices in tandem, a feedback path coupling the output circuit of the last of said tandem connected devices with the input circuit of the first of said devices and an impedance network included in said feedback path, said impedance networks each comprising a plate circuit branch and a grid circuit branch, said networks having a combined attenuation which is substantially constant and small relative to the combined gain of said amplifying devices at frequencies in an assigned operating range, which increases with the decrease of frequency just below said operating range, and which has a progressively changing phase as the frequency is still further lowered, means for delaying the building up of phase shift with decreasing frequencies, said means comprising capacitance shunting a portion of the grid circuit branch of one of said coupling impedances, the magnitude of the capacitance being such as to delay the building up of phase shift with decreasing frequencies until a frequency-level is reached at which the net gain around the feedback loop is less than unity.

2. A wave amplifying system comprising a plu-

rality of vacuum tube amplifying devices, impedance networks coupling said devices in tandem, a feedback path coupling the output circuit of the last of said tandem connected devices with the input circuit of the first of said devices and an impedance network included in said feedback path, said impedance networks having a combined attentuation which is substantially constant and small relative to the combined gain of said amplifying devices at frequencies in an assigned operating range and which increases as the frequency decreases below said operating range, the rate of increase of attenuation being at first increasingly high near the operating range, then low at lower frequencies, and then high at still lower frequencies.

3. A negative feedback amplifier comprising a plurality of amplification stages and a feedback path around said stages for feeding back waves in gain reducing phase within a transmission frequency band, said amplifier including a stopping condenser grid leak resistance coupling network between two stages, the stopping condenser being of such magnitude as to produce substantially no reduction of transfer ratio across the resistance portion for frequencies in the transmission band but to reduce this ratio rapidly below the transmission band, a condenser shunting a portion of the resistance and of such magnitude as to substantially short-circuit the portion of resistance for frequencies in the transmission band but to permit substantial increase of the transfer ratio at lower frequencies, the combined effect of the two condensers being to rapidly reduce the transfer ratio for the first two octaves of frequency below the transmission band but to reduce it at a much lower rate for the next few octaves and then to reduce it rapidly for still lower frequencies.

4. A loop circuit comprising two electric space discharge devices, one having an anode and a cathode, and the other having a grid and a cathode, connections for feeding waves from the output of the other device to the input of said one device, and means for coupling the anode and cathode of said one device to the grid and cathode of said other device, said means comprising an impedance connected between the plate and the cathode of said one device, a stopping condenser and a grid leak impedance connected in series across said first impedance, means connecting said grid leak impedance between the grid and the cathode of said second device, and a condenser shunting a portion of said grid leak impedance and having its capacity value such as to increase the phase margin of stability of said loop against singing below an operating frequency range for which transmission around the loop has a substantial gain without materially reducing the gain around the loop in said range.

5. A wave amplifying system comprising an electric space discharge amplifier, a feedback path extending between the output and input of said amplifier and forming therewith a closed loop circuit, frequency selective impedance networks included in said loop circuit, said networks being proportioned to provide in combination with the gain of the amplifier a relative large loop gain at frequencies above a given frequency and to diminish the loop gain at frequencies below said given frequency at a rate between 6 and 12 decibels per octave of frequency until the loop gain attains negative value not greatly different from zero, and said networks being further proportioned to diminish the loop gain at a rate between zero and 6 decibels per octave at still lower frequencies for a short frequency interval whereby the total loop phase shift is held substantially different from zero at the frequency of zero decibel loop gain, at least one of said networks comprising a stopping condenser and a grid leak impedance including in series a resistance and a parallel combination of a resistor and a condenser.

6. An amplifier comprising three cascaded stages of vacuum tubes forming a closed feedback loop with each interstage circuit including a coupling network comprising a stopping condenser and a grid leak impedance, in the case of each coupling network said grid leak impedance comprising in series a resistance and a parallel combination of a resistance and a shunting condenser, said shunting condenser having the value of its capacity such as to increase at least several fold in comparison with conditions in the absence of the shunting condenser the decibel value reached by the insertion loss of the coupling network before its phase shift exceeds 60 degrees with diminishing frequency below the operating frequency range of the amplifier.

7. In a negative feedback amplifier including a plurality of stages with a feedback path for feeding waves from the output to the input in gain reducing phase throughout an operating frequency range, and including interstage networks each comprising series coupling capacity with shunt impedance branches on both sides of the series capacity, the network branch connected across the input of one of said stages between a terminal of the series capacity and the grid of said one stage comprising a grid leak resistance and a capacity in shunt relation to a part only of said leak resistance, said shunting capacity, said grid leak resistance, the portion of said resistance that is shunted by said capacity, and the series coupling capacity connected to said last-mentioned impedance branch being proportioned with respect to one another to cause said interstage network to have increasing attenuation with decreasing frequencies near the lower edge of the operating frequency range to a much higher value of attenuation than would be attained without said shunting capacity, before the phase shift produced by said interstage network has reached 60 degrees and to delay increase of said phase shift above 60 degrees with further decrease of frequency until the attenuation has attained a still larger value.

8. The combination of claim 7 in which said shunting capacity, said grid leak resistance, the portion of said resistance that is shunted by said capacity, and said series coupling capacity are proportioned with respect to one another to cause said phase shift produced by said interstage network to have approximately the constant value of 45 degrees over a range of low frequencies and to cause the attenuation to increase with decreasing frequency within said range of low frequencies by a factor of more than 2 to 1 on a logarithmic scale.

JULIAN M. WEST.