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WAVE TRANSLATION SYSTEM

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FIG. 1

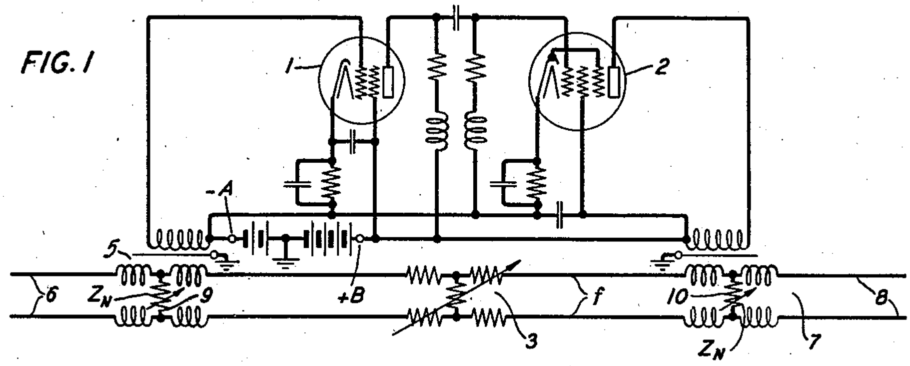

FIG. 2

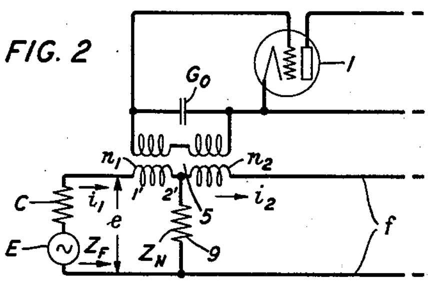

FIG. 3

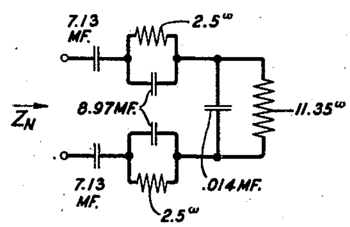

FIG. 4

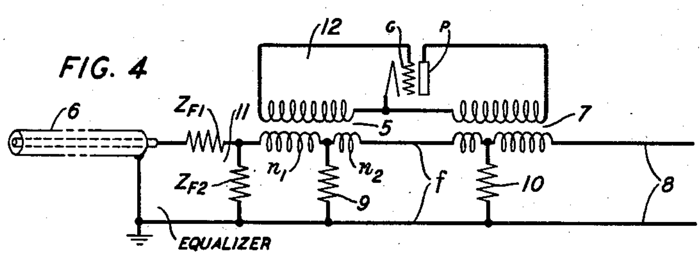

FIG. 5

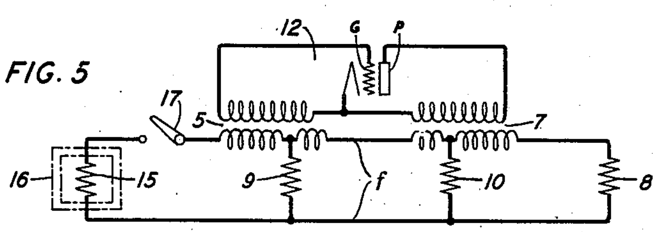

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FIG. 6

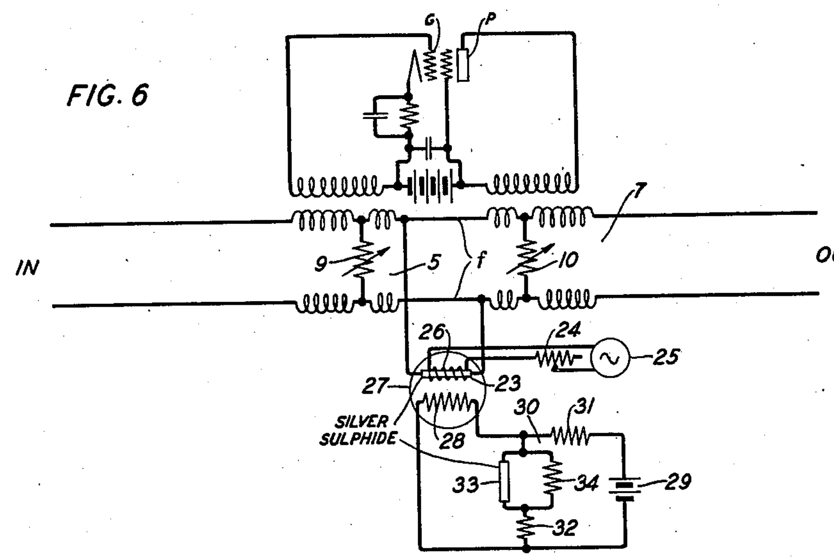

FIG. 7

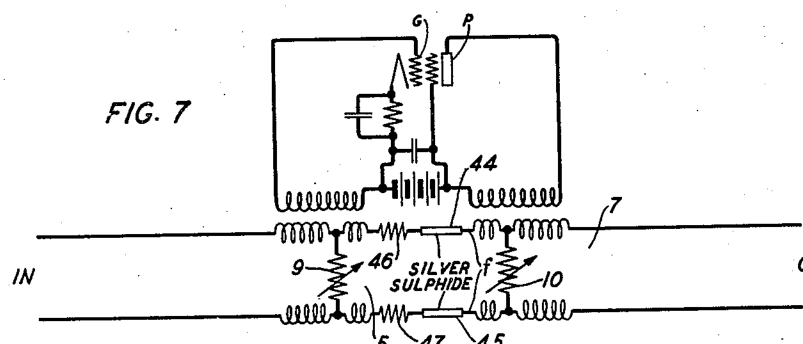

FIG. 8

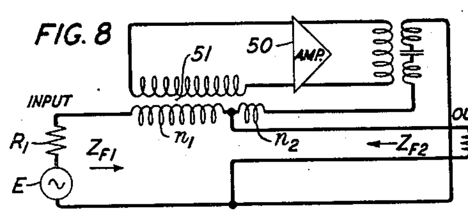

FIG. 9

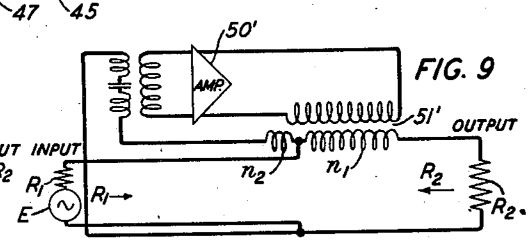

FIG. 10

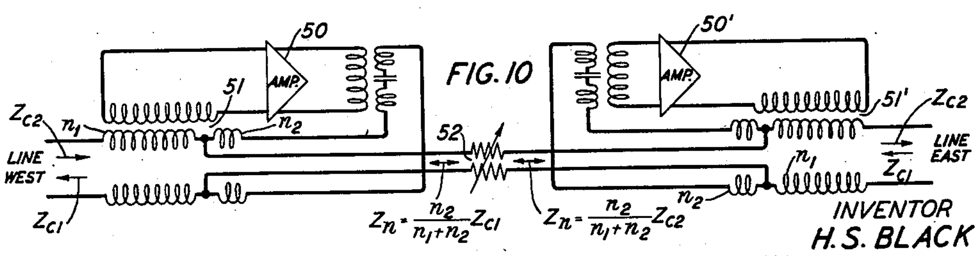

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FIG. 11

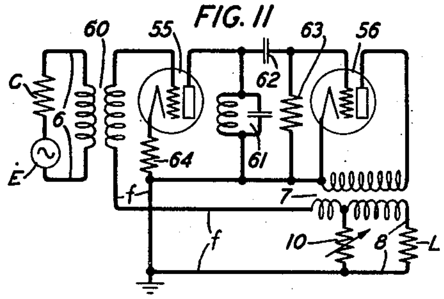

FIG. 11A

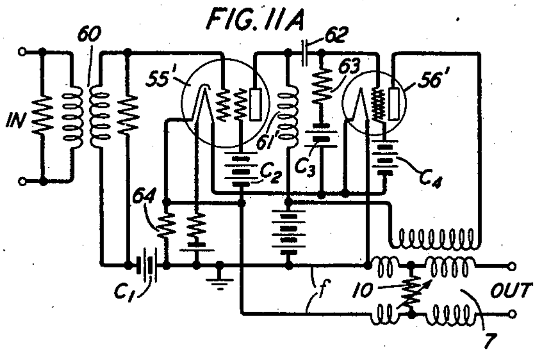

FIG. 12

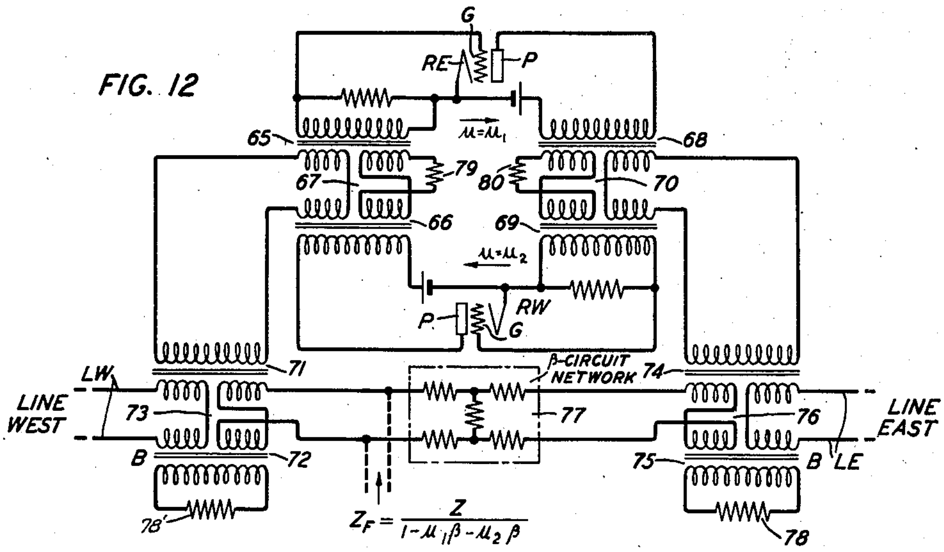

FIG. 13

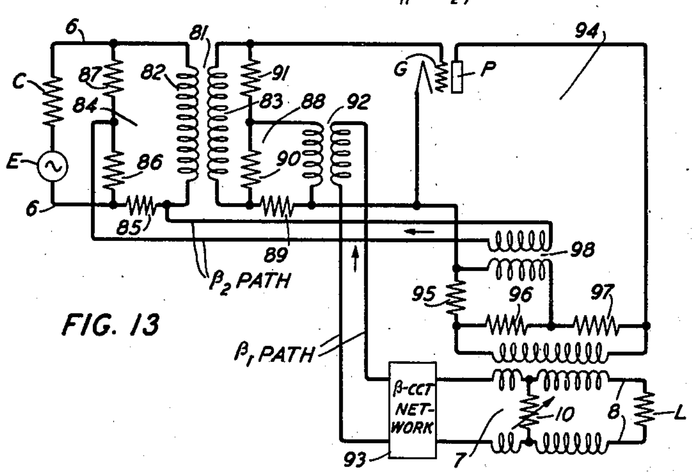

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FIG. 14

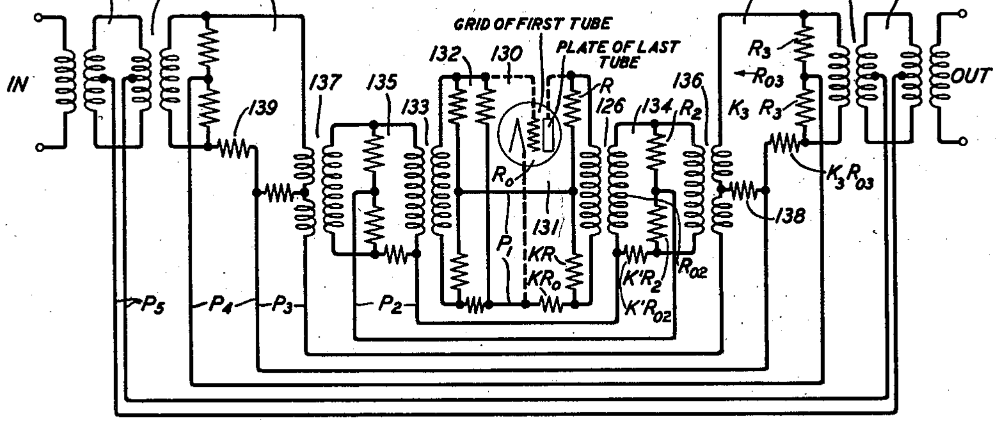

FIG. 15

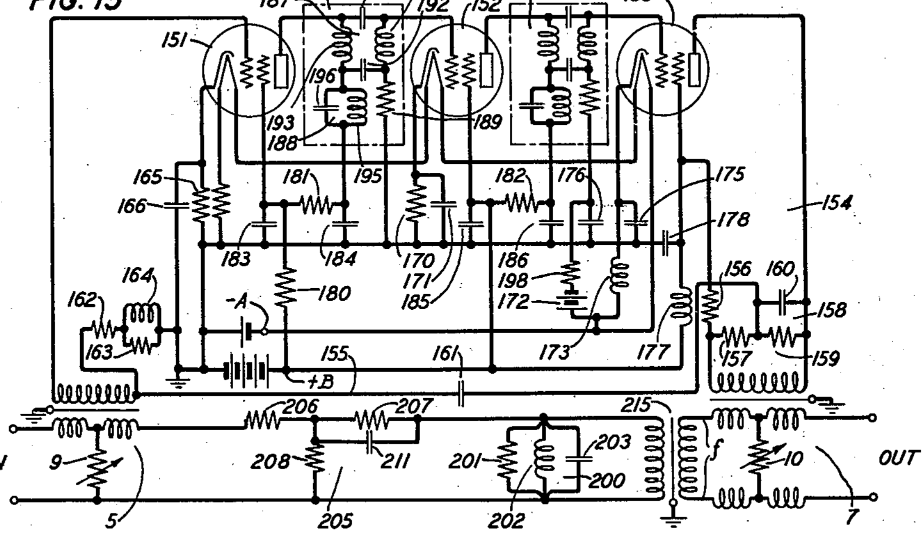

FIG. 15A

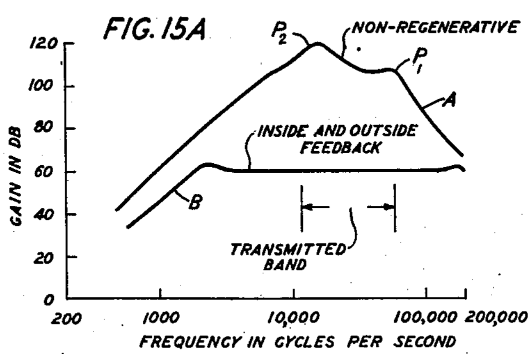

FIG. 15B

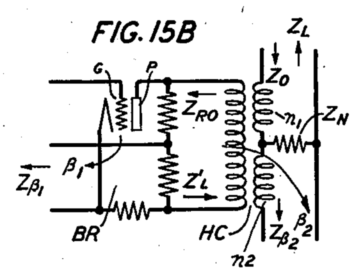

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FIG. 16

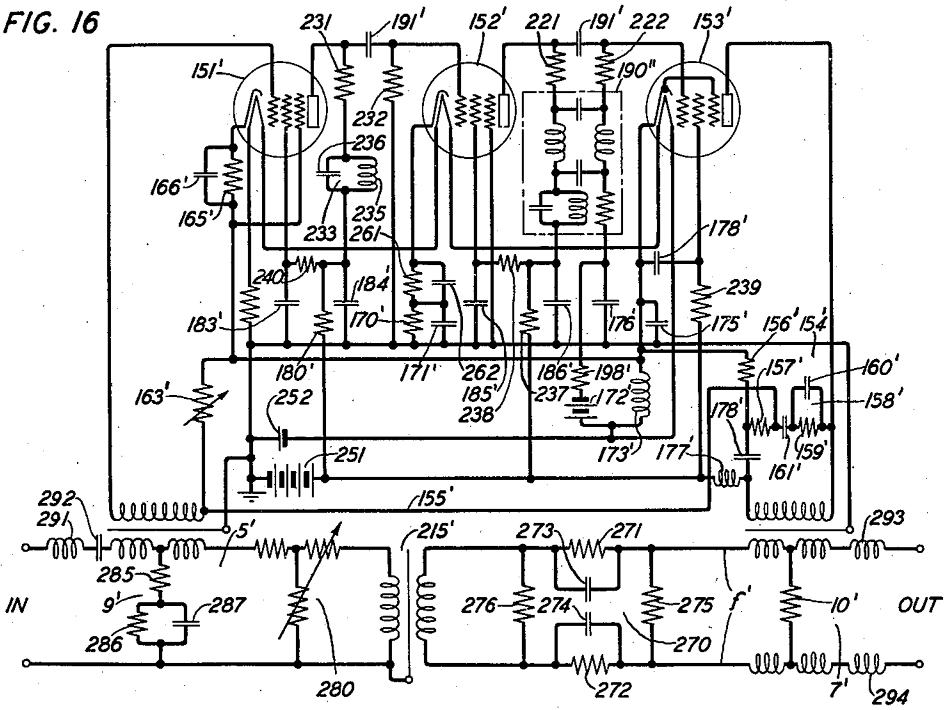

FIG. 18

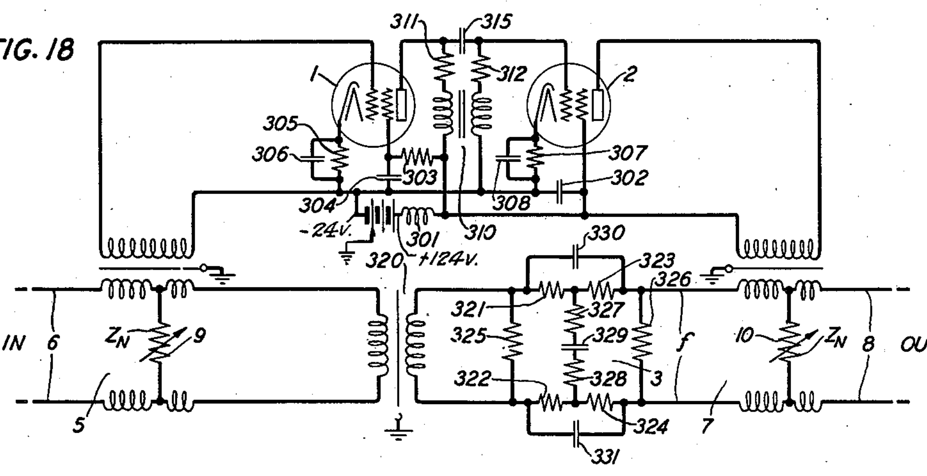

FIG. 19

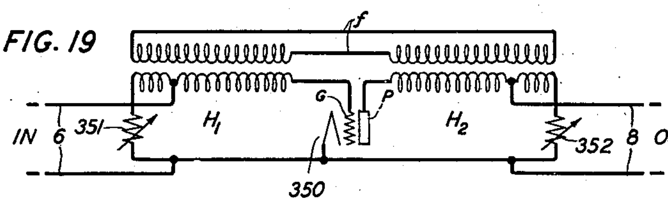

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FIG. 17

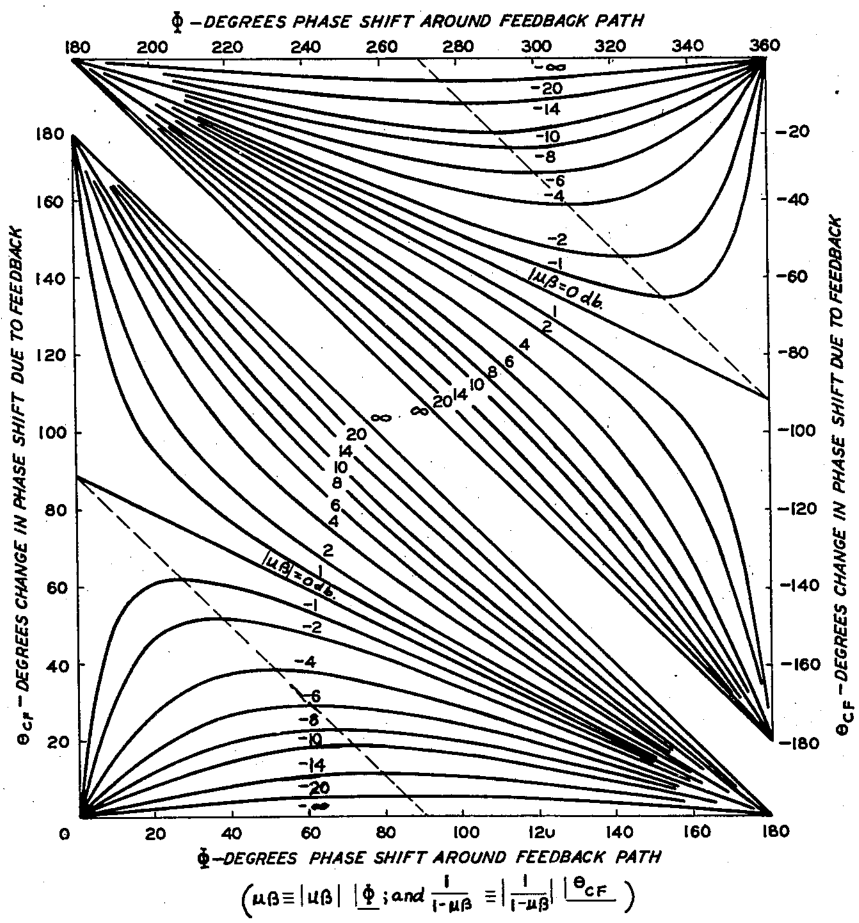

FIG. 16A

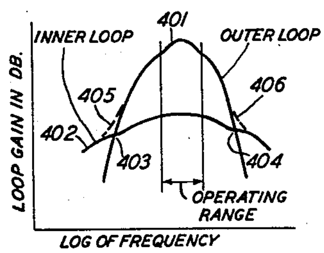

FIG. 16B

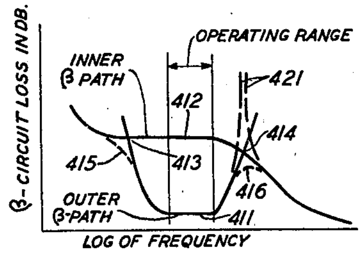

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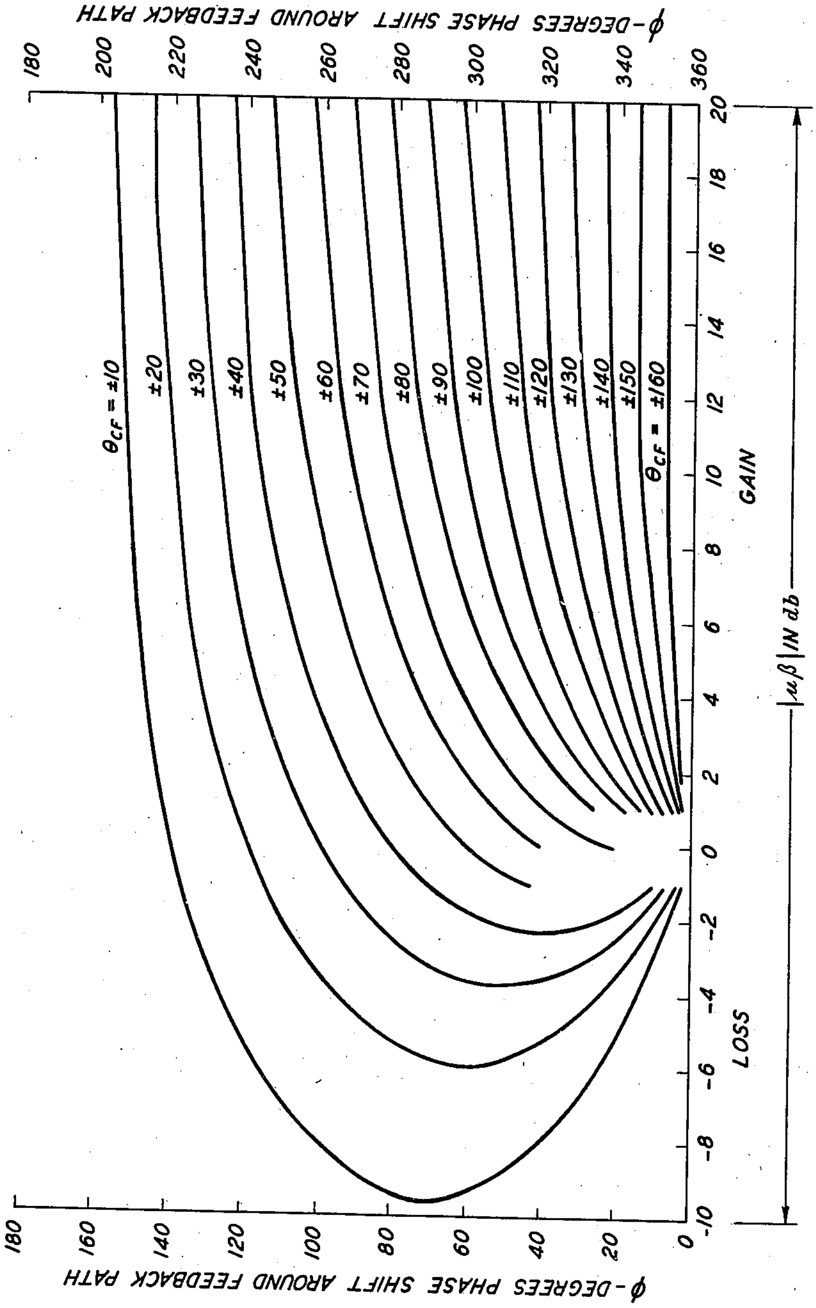
FIG. 17A



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Patented Aug. 6, 1940

2,209,955

UNITED STATES PATENT OFFICE

2,209,955

WAVE TRANSLATION SYSTEM

Harold S. Black, Elmhurst, N. Y., assignor to Bell Telephone Laboratories, Incorporated, New York, N. Y., a corporation of New York

Application December 5, 1936, Serial No. 114,390

17 Claims. (Cl. 178—44)

This application is a continuation in part of my copending applications Serial No. 606,871 filed April 22, 1932, for Wave translation system, and Serial No. 663,317, filed March 29, 1933, for Wave translation systems which issued as U. S. Patent 2,102,671 December 21, 1937 and U. S. Patent 2,131,365 September 27, 1938 respectively.

This invention relates to wave translation and aims to control transmission of waves with regard to amplitude or phase relations or both.

The invention also aims to control transmission properties of wave translating systems, as for example, systems involving means for increasing the power level of waves.

Representative objects of the invention are to control modulation, stability, impedance relations, wave reflections, cross-talk, resistance noise, signal-to-noise ratio, heat transfer, temperature, transmission efficiency, gain-frequency and phase relations in such systems.

A feature of the invention relates to effecting such control by feedback.

Objects of the invention are also to control feedback and facilitate application of feedback.

The feedback may be, for example, feedback of a portion of the output wave of a system in gain-reducing phase and in amount sufficient to reduce distortion below the distortion level without feedback. Such feedback is disclosed, for example, in the above-mentioned copending applications and in my article on Stabilized feedback amplifiers, published in Electrical Engineering, January 1934, pages 114 to 120.

In one specific aspect the invention is an amplifier having a path which produces such feedback, the input (or output) end of the amplifying path being joined to the feedback path and the incoming (or outgoing) circuit of the amplifier by an interconnecting network that has branches with mutual impedance and causes the input (or output) impedance of the amplifier to stabilize around a fixed value which, with considerable amounts of feedback, is independent of variations within the amplifier, its gain, or the amount of feedback.

The interconnecting circuit at the amplifier input, or the interconnecting circuit at the amplifier output, or each interconnecting circuit, may be for instance a hybrid coil or bridge transformer with four pairs of terminals—one pair connecting to the incoming (or outgoing) lines, a second pair to the feedback path, a third pair to the input (or output) circuit of the amplifying element, and the remaining pair to a two-terminal impedance, the hybrid coil net.

Considering, for example, the case in which the hybrid coil is a three-winding transformer with a line winding connected between the line and the hybrid coil net, and a feedback winding connected between the feedback path and the hybrid coil net (the net thus being across the "bridge points" of the hybrid coil) it is shown hereinafter that, in the case of either an equality ratio or an inequality ratio between the turns of the line and feedback windings, the input or output impedance of the amplifier, according to whether the hybrid coil is the input hybrid coil or the output hybrid coil, can be made to equal the impedance of the hybrid coil net multiplied by one plus the turns ratio, regardless of whether the hybrid coil is balanced in the sense of giving passive conjugacy. Thus, for any chosen turns ratio between the line and feedback windings, the desired amplifier input or output impedance can be realized by choice of the impedance of the hybrid coil net.

This same turns ratio controls the transmission loss from the incoming line to the amplifying path, in the case of the input hybrid coil, or from the amplifying path to the outgoing line, in the case of the output hybrid coil; and if the turns ratio exceeds unity say by several times, varying the amplifier impedance by changing the net impedance will not materially change the transmission loss, which moreover can readily be kept below a few tenths of decibel, and, further, the amount of feedback will be influenced but little by the line impedance facing the amplifier input or output impedance and likewise the transmission between the line and the amplifier will be but slightly influenced by the magnitude of the impedance of the feedback path seen from the input hybrid coil or the output hybrid coil.

The input and output impedances of the amplifier are independent of one another and may have the same value or different values.

By making the input hybrid coil so that without feedback the amplifier input impedance is high compared to the impedance of the incoming circuit instead of being a match for the latter impedance, the ratio of signal to resistance-noise in the amplifier output energy can be made as much as three decibels greater than for the matched impedance condition; and with proper choice of the value of impedance for the net of the input hybrid coil, feedback can make the amplifier input impedance match the impedance of the incoming circuit while the improvement in the ratio of signal to resistance-noise is retained. Indeed, the improvement that feedback can produce in the ratio of signal to resistance-noise is by no means limited to three decibels, and, as will now be explained, by the aid of feedback action an impedance from which to work the amplifier can be created which does not produce resistance-noise.

The relation of the amplifier input or output impedance to the impedance of the input or output hybrid coil net can be used for transforming impedances, i. e., for obtaining from the net impedances an impedance equal to the net impedance multiplied by a constant (unity plus the turns ratio), provided the amount of negative feedback is sufficient. Thus, the impedance obtained is an enlarged copy of whatever impedance is used as the hybrid coil net (the impedance across the bridge points of the hybrid coil) and the latter impedance may be of any suitable type. For example, it may be an ordinary impedance such as a resistance, an inductance, a capacity or any complex form of (two-terminal) impedance constructed of resistances, inductances and capacities.

I have discovered that feedback action can produce resistances or generalized impedances free from resistance-noise—that feedback action can transform ordinary resistances (or generalized impedances) to resistances (or generalized impedances) that are free of all noise, including thermal agitation. For example, if the net across the bridge points of the input hybrid coil be a resistance, the feedback can transform this resistance by producing an enlarged copy of it as the amplifier input impedance, this copy having the remarkable property of freedom from resistance-noise. This property of freedom from resistance-noise or thermal agitation electromotive force obtains likewise for the case in which the input impedance is a generalized impedance (which may include inductance or capacity or both) produced from a corresponding generalized impedance across the hybrid coil bridge points by the feedback action. Thus, any desired simple or complex (two-terminal) impedances (resistances, capacities, inductances or generalized impedances) with the extraordinary property that they are theoretically free from thermal agitation electromotive force and practically free at least to a considerable degree, can be obtained from similar smaller impedances which may if desired be of ordinary type, or which, on the other hand, may themselves in turn be impedances similarly obtained free of resistance-noise by feedback action.

Any physically realizable impedance can thus be reproduced free of resistance-noise; and of such impedances free from resistance-noice can be constructed all manner of impedance networks or active or passive transducers free from resistance-noise, for example, filters, equalizers, phase shifters, delay distortion correctors, impedance correctors, artificial cables or lines, amplifiers and systems in general.

Such impedances and networks that do not produce resistance-noise have a wide scope of practical application. For instance, in the case of the amplifier referred to above, a transmission equalizing network constructed of such impedances can be connected between the incoming line and the input hybrid coil. The incoming line may be, for example, a coaxial system or other cable delivering signals whose lower limit of transmission level is fixed by the cable resistance noise; the equalizer may be, for example, a constant resistance equalizer; the hybrid coil may be such that the amplifier input impedance without feedback would be high compared to the equalizer impedance from which the amplifier works; and the hybrid coil net may be a resistance or impedance of such value that the feedback matches the amplifier input impedance to the equalizer impedance from which the amplifier works. Notwithstanding the fact that the equalizer may include elements having resistance, the equalizer and amplifier need not materially increase the ratio of resistance-noise to signal; and thus the desired equalization and amplification of the signal can be accomplished without decreasing the ratio of signal to resistance-noise.

The resistances and impedances free from resistance-noise render possible large improvements in signal-to-noise ratio (by no means limited to improvements of three decibels) and the penetration of substantial amounts, theoretically as great as desired, below the noise level heretofore considered as an inevitable, natural, impenetrable limit established by thermal agitation.

Features of the invention include resistances, generalized impedances, networks of generalized impedances, and active and passive transducers, free from resistance-noise; and a further feature of the invention is production of such devices by feedback action.

I have discovered that feedback action can abstract heat from a body. When a resistance is connected to an amplifier, feedback action can be made to abstract heat from the resistance or cool it. For example, if an electric conductor or resistance be connected across resistance of the type described above as free from resistance-noise, the effect of making the connection is to abstract heat from the ordinary resistance or cool it, the ordinary resistance receiving no energy from the other resistance but giving up energy of thermal agitation to the other resistance in the form of an electric current. To observe the cooling effect the resistance to be cooled can be heat-insulated. If it is not insulated, the small losses due to thermal agitation are readily replaced from the relatively vast reservoir of heat surrounding the unit.

The variation of the amplifier output impedance by changing the impedance of the net of the output hybrid coil can be accomplished without materiall changing the impedance into which the output tube (or the amplifying path) works; and thus the output tube, though its impedance may differ from its optimum load impedance, can be worked into an impedance having substantially the optimum value and yet be made to appear from the outgoing line to be equal to the impedance of the outgoing line. In other words, the feedback cannot only transform the net impedance to the desired value of amplifier output impedance, but can do so without materially affecting the impedance into which the output tube works; and this ability of the feedback to change the difference between two impedances (such as the tube impedance and the impedance of the outgoing line) by a different amount for one direction of transmission between them than for the other direction, makes it possible to work the tube into its optimum load impedance and at the same time match the amplifier output impedance and the line impedance, without entailing the transmission loss entailed in doing this without feedback, (as for instance in shunting across the tube impedance a resistance which in combination with this parallel tube impedance is equal to twice the optimum load impedance of the tube and connecting the resistance and the tube to the line with a transformer that changes the difference between the impedances which it connects the same amount for both directions of transmission).

Since, as noted above, with considerable amounts of feedback the input and output hybrid coils can render the amplifier input and output impedances independent of the amount of feedback, the amplifier gain can readily be varied without varying the amplifier input and output impedances, by varying the attenuation of the feedback path, as for example by an adjustable series or shunt resistance in the feedback path. The resistance may be, for instance, a thermo-responsive resistance, as for example a silver sulphide resistance adjustable by control of its temperature. If desired, the attenuation change made in the feedback path for gain control can be made automatically, as for example in response to transmission level. For instance, a silver sulphide resistance unit may be connected in series in the feedback path, to render the amplifier a volume limiter by virtue of decrease in resistance of the silver sulphide due to heating of the silver sulphide resulting from increase of current through it in response to increase in output voltage of the amplifying path.

Specific aspects of the invention also embrace feedback systems, including multiple feedback systems and systems involving repetition of the feedback process, with various forms of hybrid coil feedback connections.

Other objects and aspects of the invention will be apparent from the following description and claims.

In the drawings, Fig. 1 shows an amplifier of a specific form referred to above;

Fig. 2 is a diagram for facilitating explanation of operation of such amplifiers;

Fig. 3 shows a specific form of network which may be used in such an amplifier to match the amplifier input or output impedance to the impedance of a specific type of attached cable circuit;

Fig. 4 shows a cable terminated in an equalizer working into such an amplifier, the equalizer and amplifier being constructed, in accordance with the invention, to avoid increasing the ratio of resistance-noise to signal;

Fig. 5 illustrates cooling a resistance by feedback action produced for example by such an amplifier;

Fig. 6 shows a gain control for such an amplifier;

Fig. 7 shows a volume limiting circuit embodying a specific aspect of the invention;

Figs. 8 to 10 show feedback amplifier systems with a single hybrid coil or bridge transformer network interconnecting the input and the output of an amplifying element with an incoming circuit and an outgoing circuit;

Fig. 11 shows a multiple feedback amplifier with a hybrid coil feedback connection;

Fig. 11A shows an amplifier which is a modification of that of Fig. 11;

Fig. 12 shows a two-way transmitting, single loop feedback system with hybrid coil feedback connections through two feedback paths having a common portion;

Fig. 13 shows a triple loop feedback system with a hybrid coil feedback connection;

Figs. 14, 15 and 16 show feedback amplifiers with hybrid coil feedback connections and with repetition of the feedback process;

Fig. 15A shows gain-frequency characteristics of the amplifier of Fig. 15;

Fig. 15B shows a type of feedback connection used in Fig. 15, for facilitating explanation of operation of such connections;

Figs. 16A and 16B show curves facilitating explanation of the operation of the amplifier of Fig. 16;

Figs. 17 and 17A show curves for facilitating design of feedback systems;

Fig. 18 shows a specific form of amplifier of the general type of the amplifier of Fig. 1;

Fig. 19 shows an amplifier or system with feedback through hybrid coil connections that are modifications of those of Fig. 1.

The amplifier of Fig. 1 may be a stabilized feedback amplifier of the general type disclosed, for example, in the copending applications and published article mentioned above. It comprises an amplifying path or element shown as including tandem connected vacuum tubes **1** and **2**, and comprises a feedback path f shown as including a transmission control network **3** of generalized impedances. The amplifying path or element may be referred to as the μ-circuit, and the feedback path may be referred to as the β-circuit, the significance of μ and β being as indicated in the applications and article just mentioned. The network **3** may be referred to as the β-circuit network.

An input hybrid coil **5** couples the incoming circuit **6** and the feedback path f to the input end of the amplifying path; and an output hybrid coil **7** couples the output end of the amplifying path to the outgoing circuit **8** and the feedback path. One of the important advantages of this type of feedback circuit, with considerable amounts of feedback, is that the input and output impedances of the amplifier stabilize around fixed values that are independent of variations within the amplifier, its gain, or the amount of feedback, regardless of whether, in the passive condition of the amplifier, the hybrid coils are balanced, or in other words, regardless of whether the impedance Z_n of the hybrid coil net (i. e., the impedance **9** or **10** across the bridge points of the hybrid coil) is such as to give passive conjugacy (i. e., conjugacy in the absence of feedback) between the line and the feedback path. Inasmuch as the case of relatively large amounts of feedback (relatively large values of $\mu\beta$) in a case of great practical importance, the derivation of the amplifier input impedance for that case is given by reference to Fig. 2. The equation for the output impedance is similar.

In Fig. 2, for simplicity, the hybrid coil is shown in the unsymmetrical form (i. e., as unbalanced to ground). The source of electromotive force E and impedance C represent or replace line **6** of Fig. 1. The capacity G_0 represents the effective grid-cathode capacity of tube **1**. The voltage across the input terminals of the amplifier is designated e. The input impedance of the amplifier is designated Z_F. The impedance of the hybrid coil net is designated Z_N. The numbers of turns in the line and feedback windings of the hybrid coil are designated n_1 and n_2, respectively. The turns ratio

$$\frac{n_1}{n_2}$$

will be called t.

$$e=i_1Z_F \qquad (1)$$

If $\mu\beta$ is very large compared with 1, there will be current i_2 return to the feedback side of the coil sufficient to reduce the flux in the coil 5 by a factor equal to

$$\frac{1}{1-\mu\beta}$$

This means that the voltage across terminals 1', 2' will approach zero as compared to the value of e. The fed back current, i_2, to accomplish this flux cancellation will be equal to

$$\frac{n_1}{n_2}\cdot\frac{\mu\beta}{1-\mu\beta}i_1$$

If the voltage across 1', 2' is negligible, e approaches (i_1-i_2). Z_N=drop across the network.

$$\therefore e=i_1Z_F \doteq i_1\left(1-\frac{n_1}{n_2}\cdot\frac{\mu\beta}{1-\mu\beta}\right)Z_N \qquad (2)$$

and, for large amounts of feedback,

$$Z_F \doteq \left(1+\frac{n_1}{n_2}\right)Z_N \qquad (3)$$

It is thus seen that, for large values of $\mu\beta$

$$Z_F=\left(\frac{n_1+n_2}{n_2}\right)Z_N=kZ_N,$$

k representing the constant

$$\frac{n_1+n_2}{n_2}=(1+t).$$

This gives the very valuable result that the impedance of the amplifier is equal to the impedance of the net connected across the hybrid coil bridge-points multiplied by one plus the turns ratio of the equality ratio or usually inequality ratio hybrid coil. Thus, assuming $\mu\beta$ is large, the impedance of the amplifier can be made to approach what is wanted as closely as can the net. Using this procedure, amplifiers have been built and used having remarkably good impedances. It has been observed that the input or output impedance of the amplifier can easily be varied, for example, in the ratio of 100:1 by merely changing the impedance of the net. Moreover, if the hybrid coil is an inequality ratio hybrid coil with the turns ratio t sufficiently large, so that $n_1>n_2$ say by several times, then varying the impedance in this manner will hardly vary the input or output loss at all, and further, each of these losses can readily be kept below a few tenths of a decibel (instead of the usual three decibels for an equality ratio hybrid coil). The ability to vary the amplifier input or output impedance (or both) in such an easy manner without much affecting the transmission is a highly desirable feature. Z_F is a surprisingly accurate copy of Z_N, either enlarged or attenuated according as n_1 is greater than or less than n_2.

Thus, if Z_N is for example a capacity, the (input or output) impedance of the amplifier is a capacity, etc.

Regarding conjugacy relations with the feedback through hybrid coils, the transmission between the (incoming or outgoing) line and the amplifying path or μ-circuit is affected by impedance of the feedback path f or β-circuit as seen from the (input or output) hybrid coil. Also, the transmission between the amplifying path and the feedback path is affected by the impedance of the (incoming or outgoing) connecting line; and as a corollary the amount of feedback obtained for any setting of the β-circuit network such as network 3 is somewhat dependent upon the impedance of the connecting line. However, by making the hybrid coils inequality ratio hybrid coils with one of the quantities n_1 or n_2 sufficiently exceeding the other, as for example, with one of these quantities several times as large as the other, the effect of the line impedances upon the amount of feedback can be made negligibly small and likewise the effect that the value of the impedances of the feedback path as seen from the hybrid coils has upon the transmission between the incoming or outgoing line and the amplifying path can be made negligibly small.

Thus, regarding conjugacy relations with feedback through hybrid coils, an indication of the degree of conjugacy may be obtained by noting the change in i_1 as the β-circuit impedance is changed. Changing the β-circuit impedance (generator impedance producing i_2) will change the value of $\mu\beta$. But it can be seen from equation (2) above, that Z_F is dependent on

$$\frac{\mu\beta}{1-\mu\beta}$$

For values of $\mu\beta$ which are sufficiently larger than 1 variations in

$$\frac{\mu\beta}{1-\mu\beta}$$

due to changes in $\mu\beta$ are nil, or Z_F is independent of $\mu\beta$. Therefore, regardless of the turns ratio,

$$\frac{n_1}{n_2}$$

the input impedance presented to the connecting circuit is rendered independent of the feedback path. But the converse is not true, i. e., the value of $\mu\beta$ is not independent of the connecting circuit, unless the hybrid coil possess a passive impedance balance.

The incoming and outgoing lines are in conjugacy with the feedback path when the proper hybrid coil balancing nets 9 and 10 are used. For example, the amplifier output impedance is not a function of the impedance of the feedback path by direct transmission, but is controlled by the apparent plate impedance resulting from the feedback that the impedance of the feedback path provides. That is, feedback causes the apparent plate impedance to approach such a value that the output hybrid coil will be in dynamic balance. It is this property of hybrid coils, in conjunction with sufficient negative feedback, that causes the nets 9 and 10 of the input and output hybrid coils to determine the amplifier input and output impedances.

Since either an input hybrid coil or an output hybrid coil can render the amplifier input and output impedances independent of one another (regardless of whether the hybrid coil is in passive balance), it results that if an input hybrid coil is used, or an output hybrid coil is used, or both are used, the amplifier input and output impedances can be given equal values or altogether different values, at will.

Particularly in the communication field, there are many applications of amplifiers requiring the input (and also the output) impedance of the amplifier to match the impedance of the circuit it joins. If the input impedance of the amplifier is required to match, instead of being permitted to greatly exceed the impedance of the cable or circuit to which it connects, then for the same insertion gain the amplified noise that is due to thermal agitation and appears in the output of the amplifier will be about three decibels more than for the case in which the amplifier input impedance is relatively high, assuming a proper and well designed input circuit. However, with an input hybrid coil for example as shown in Fig. 1 or Fig. 2, the coil (and its terminating impedance, if any, across its winding attached to the grid and cathode of the first tube) can be chosen so that the amplifier input impedance without feedback is high, and at the same time the value Z_N of the impedance of the net of the input hybrid coil can be chosen so that the negative feedback will cause the amplifier input impedance to be improved and stabilized around a proper value that will match the input connecting circuit. As a result, when the insertion gain is the same as that of an amplifier without feedback but whose input impedance is high, the amplified resistance-noise at the output of the two amplifiers will be the same (because in either amplifier the noise in question depends upon the resistance component of the passive impedance without feedback between the grid and cathode of the first tube), and yet the feedback amplifier will have a matched impedance instead of a very high impedance. For systems of this character whose general noise level is of the order of magnitude of resistance-noise, it can be shown, other things being equal, that under certain circumstances this may amount to a 2:1 saving in the amount of output power on the basis of comparable signal-to-noise ratios. Moreover, as will now be explained, the improvement that feedback can produce in the ratio of signal to resistance-noise is by no means limited to three decibels, and feedback action in circuits such for example as those of Figs. 1 and 2 can produce stable impedances which do not create resistance-noise and from which the amplifier can be worked.

In circuits such for example as that of Fig. 1 the relation of the amplifier input or output impedance to the impedance Z_N of the input or output hybrid coil net 9 or 10 can be used for transforming impedances, or in other words for producing from the net impedance an impedance equal to the net impedance multiplied by the constant

$$\left(1+\frac{n_1}{n_2}\right)$$

provided the amount of negative feedback is sufficient. Thus, the impedance produced as the amplifier input or output impedance is an enlarged copy of whatever impedance Z_N is used as the net 9 or 10 of the input or output hybrid coil, and the impedances 9 and 10 may be of any suitable types. For example, either may be an ordinary impedance such as a resistance, an inductance, a capacity or any complex form of (two-terminal) impedance constructed of resistances, inductances and capacities.

Further, in circuits such as those of Figs. 1 and 2, the feedback action can transform ordinary resistances or generalized impedances that produce resistance-noise to corresponding resistances or generalized impedances that are free of all noise, including thermal agitation. For example, if the net 9 across the bridge-points of the input hybrid coil 5 be a resistance, the feedback can, as just noted, transform this resistance by producing an enlarged copy of it as the amplifier input impedance, and this copy will be free from resistance-noise. This property of freedom from resistance-noise or thermal agitation electromotive force obtains likewise for the case in which the amplifier input impedance is a generalized impedance produced from a corresponding generalized impedance across the hybrid coil bridge-points by the feedback action.

For instance, the impedance Z_N of the net 9 of hybrid coil 5 in Figs. 1 and 2 may be the impedance Z_N shown in Fig. 3, which, with

$$\frac{n_1+n_2}{n_2}=10 \text{ and } \mu\beta>>1$$

causes the stabilized input impedance of the amplifier to be an extremely close match, over the 12 kilocycle to 60 kilocycle frequency range, for the impedance of a 19 gauge, non-loaded, .062 capacity standard toll cable; and then this input impedance (the input to the amplifier) is free from noise due to thermal agitation.

In general, in thus building an impedance equal to $(t+1)Z_N$, free from thermal agitation, there are no restrictions on Z_N, which may be any combination of coils, resistances or condensers, or a generalized impedance. Moreover, of such impedances free from resistance-poise can be constructed all manner of impedance networks or active or passive transducers free from resistance-noise.

For example, Fig. 4 shows an equalizer 11 and an amplifier 12 which are substantially free from resistance-noise notwithstanding the fact that the equalizer may, if desired, include resistance. The equalizer may be, for instance, a constant-resistance equalizer for equalizing the cable attenuation. The amplifier may be of the type shown in Fig. 1, and may have any desired number of stages, G and P designating the grid of the first tube and the plate of the last tube (in this figure and also in other figures of the drawings). The incoming line 6 may be, for example, a coaxial system or other cable delivering to the equalizer signals whose lower limit of transmission level is fixed by the cable resistance-noise. The input hybrid coil 5 may be such that the amplifier input impedance without feedback would be high compared to the equalizer impedance from which the amplifier works; and the net 9 may be a resistance or impedance of such value that the feedback matches the amplifier input impedance to the equalizer impedance from which the amplifier works. Preferably the turns ratio

$$\frac{n_1}{n_2}$$

of the hybrid coil 5 is large, giving low loss for transmission from the equalizer to the grid G, (and correspondingly a high loss results for transmission from the feedback path f to the grid G). Then if the impedance elements of which the equalizer is constructed, shown for example as a series arm of impedance Z_{F1} and a shunt arm of impedance Z_{F2}, are quiet impedances of the type described above (impedances free from resistance-noise, which may be obtained as input impedances of negative feedback amplifiers), the equalizer can produce the desired equalization without adding resistance-noise. The amplifier then restores the signal level, amplifying the signal without introducing resistance-noise. Thus the desired equalization and amplification of the signal can be accomplished without decreasing the ratio of signal to resistance-noise.

It will be understood that the impedances Z_{F1} and Z_{F2} are input impedances of suitable feedback amplifiers. For instance, they may be input impedances of amplifiers such as the amplifier of Fig. 1 or Fig. 2, and may be obtained by giving the nets of the input hybrid coils of the amplifiers the impedance values

$$\frac{Z_{F1}}{1+t} \text{ and } \frac{Z_{F2}}{1+t}$$

respectively,

Fig. 5 shows how a suitable feedback amplifier such, for example, as amplifier 12 can cool a resistance 15 by the feedback action. The resistance is of the ordinary type producing thermal agitation electromotive force. It is shown in a heat-insulated chamber 16, and a switch 17 is shown by which the resistance can be connected to the input impedance of the amplifier. Since, as indicated above, the input impedance of the amplifier can be made a resistance free from thermal agitation by having the net 9 of the input hybrid coil 5 a resistance, it results that with switch 17 closed power is abstracted from resistance 15 without power being returned, and hence resistance 15 loses heat.

Since, with considerable amounts of negative feedback in the amplifier of Fig. 1, the impedance of net 10 will not materially affect the impedance into which tube 2 works, the tube, though its impedance may differ from its optimum load impedance, can be worked into an impedance having the optimum value and yet the net impedance can be given the value required to cause the impedance of line 8 to match the amplifier output impedance. Perhaps the clearest illustration of the advantageous nature of such operation is when the output tube 2 is a pentode. Suppose the output impedance R_0 of the pentode is 1,000,000 ohms, and it delivers maximum power when the output impedance it works into is 25,000 ohms. Then if it were to connect through a transformer, say a two-winding transformer, to an output cable or line having a resistance of 100 ohms, a 25,000:100 impedance ratio output transformer would be required. Usually, in such cases, it is also a requirement that the output impedance of the transformer on its low side match the impedance of the connected line or amplifier load. In the example given, without feedback this requirement could not be met because the output impedance of the amplifier would be the low side impedance of the coil when its high side was terminated by 1,000,000 ohms or practically open, whereas the high side winding of the coil would have to be terminated by 25,000 ohms for the low side impedance to be 100 ohms. Of course, by using a 50,000:100 impedance ratio output transformer and by adding a 50,000 ohm resistance across the high side winding in parallel with the tube, both these requirements (a 100 ohm output impedance for the amplifier and a 25,000 ohm load for the tube) could be satisfied simultaneously, but in this case, one-half of the output power delivered by the tube would be wasted.

However, by using at the output of the amplifier an inequality ratio hybrid coil connection which in itself introduces only a very slight insertion loss, both these requirements can be met simultaneously and practically all of the available tube power is useful. For example, in the amplifier circuit of Fig. 1, if the impedance of the net 10 were such as to cause the amplifier output impedance to equal 3,500 ohms, and the tube 2 were, say a coplanar grid tube having its impedance R_0 equal to 3,500 ohms and working into 3,500 ohms, and if then the coplanar grid tube were replaced by a power pentode whose impedance R_0 was 75,000 ohms and whose optimum load impedance was 3,500 ohms, this pentode would work into 3,500 ohms, thus satisfying the first requirement. It would also be found that substituting the pentode produced practically no change in the amplifier gain, because the amount of negative feedback was already great and substituting a still higher gain tube merely further increased the feedback. The amplifier output impedance if measured would likewise be found not to have changed, thus satisfying the second requirement, because as explained above, the value of the amplifier output impedance is set by the impedance of the net 10, which was not changed.

Thus, it can be seen that by using the hybrid coil connection, the pentode can be worked into its optimum load impedance and at the same time the output impedance as presented by the amplifier can be kept on a matched impedance basis and hence, since practically no power is wasted, the available output is doubled as compared to the case without feedback.

Especially in amplifiers without feedback, if the gain-load curve of the amplifier is scrutinized very closely, it will usually be discovered even at light loads, that the gain of the amplifier varies slightly with changing load. When a great number of such amplifiers are in tandem as, for example, on a telephone circuit having many repeaters, this effect, if systematic, tends to be proportional to the number of amplifiers. It leads to a degradation in the quality of the speech transmitted and this particular kind of deterioration has been termed "pep effect." Even when the number of amplifiers in tandem is very large and the characteristics of the amplifiers without feedback are unsuitable from this standpoint, by using considerable amounts of negative feedback the practical difficulties of pep effect are readily avoided. When the number of amplifiers involved is very large, the transmission characteristics of the input and output transformers may show this effect and in such instances the design of these coils is simplified by feeding back around the transformers as, for example, in the circuit of Fig. 1 or the circuit of Fig. 14 described hereinafter.

Since, with considerable amounts of negative feedback in the amplifier of Fig. 1, the impedances of network 3 will not substantially affect the amplifier input or output impedance, the network may be any desired form suitable for controlling transmission. For example, it may be a series or shunt resistance adjustable for giving amplifier gain changes independent of frequency, or a network adjustable for giving variable equalization or changes of amplifier gain dependent on frequency. When it is desired that gain characteristics be parallel and flat, they are preferably made as parallel as required or possible, and, in addition, as flat as possible. This distinction is made because for applications requiring great refinement, if the curves are parallel and almost flat, then the slight lack of flatness can be corrected by the addition of a fixed equalizer.

Fig. 6 shows an amplifier similar to that of Fig. 1, but shows the β-circuit network by way of example as a thermo-sensitive shunt element, preferably a resistance 23 of silver sulphide for controlling the gain of the amplifier. G and P in this figure, and wherever appearing in other figures of the drawings, designate the first grid of the first stage and the plate of the last stage of the amplifier, and indicate that the amplifier may have any suitable number of stages. To vary the resistance, for changing the amplifier gain, the temperature of the resistance is varied. This temperature variation is accomplished by adjusting a resistance 24 which controls heating current supplied from an alternating current or direct current power source 25 to a heating element 26 for the silver sulphide resistance 23.

Ordinarily the amount of power required to heat the silver sulphide resistance so as to produce a gain change of a few to as much as 80 to 100 decibels would not exceed a small fraction of a watt and in many instances could be as little as one milliwatt or even a fraction of a milliwatt.

Used in this manner it would often be required that the silver sulphide be stable with time and humidity. Where effects of room temperature variations tend to be objectionable, the silver sulphide resistance may be enclosed in a heat insulated chamber 27. The heat insulated container, possessing sufficient heat capacity with respect to the size and heat capacity of the silver sulphide resistance, is very helpful in reducing effects of variations in ambient or room temperatures upon the silver sulphide resistance. If desired, to compensate for effect of room temperature on operation of the silver sulphide unit, the heat insulated container can be maintained at a constant temperature, above the highest room temperature.

This could be done by a thermostatic control, but in Fig. 6 is accomplished by a chamber-heating element 28 supplied with heating current from alternating current or direct current power source 29 through a regulating network such for example as network 30. The network 30 is shown as having a series resistance arm 31 and a shunt arm comprising a resistance 32 in series with two parallel resistances 33 and 34, resistance 33 being a silver sulphide resistance. The constants of the network depend upon the thermal and other properties of the particular silver sulphide unit 33. With the temperature of the chamber 27 elevated above the highest room temperature by the chamber-heating unit 28, if the room temperature rises the resistance of the silver sulphide element 33 falls sufficiently to reduce the heating current in the element 28 so that the temperature of the chamber 27 is held constant.

The voltages or currents from the power supply sources 25 and 29 should be relatively stable.

With the gain control in the feedback path of a stabilized feedback amplifier, as indicated for example at 3 in Fig. 1 and at 23 in Fig. 6, reducing the gain is accomplished by correspondingly increasing the amount of negative feedback; and this improves the amplifier performance accordingly, for example reducing modulation, increasing gain stability, and, in the case of hybrid coils that are in passive unbalance, increasing the independence of the amplifier input and output impedances with respect to the impedances of the network such as 3. Since the amplifier gain practically equals the loss in the feedback circuit and the working gain is usually appreciable, a loss is usually required in the feedback path; and consequently, considerable loss may occur in the gain control device in the feedback path without necessarily being a disadvantage.

The gain control 3 in Fig. 1, or the adjustable gain-control contact of resistance 24 in Fig. 6, may be operated manually; or if desired it may be operated automatically, for example as the gain-varying element is operated by the pilot apparatus in the transmission control system disclosed in H. S. Black Patent 1,956,547, May 1, 1934; E. I. Green Patent 1,918,390, July 18, 1933; or J. R. Fisher-C. O. Mallinckrodt Patent 2,116,-600, May 10, 1938.

Silver sulphide is preferred as the temperature responsive transmission control element because of its large (negative) temperature coefficient of resistance, constancy and uniformity of performance as disclosed more fully in the Fisher and Mallinckrodt application just mentioned. The specific form of the silver sulphide element may be, for instance, the form disclosed therein; and the preparation of the element may be, for example, as disclosed in J. R. Fisher Patent 2,082,102, June 1, 1937.

Instead of having the silver sulphide element shunted across the feedback path, it can be placed in series in the feedback path in the manner indicated in Fig. 7, about to be described; and if it is then desired that the circuit be symmetrical (balanced to ground), two such elements can be used, as indicated in Fig. 7. Both can be in the same heat chamber, both heated by the same heater 26 or each heated by an individual heater such as 26.

Fig. 7 shows an amplifier which may be a stabilized amplifier, having considerable negative feedback, similar to the amplifiers of Fig. 1 and Fig. 6. However, the β-circuit network is shown as two silver sulphide resistances 44 and 45 and two ordinary resistances 46 and 47. This network automatically controls the amplifier gain to maintain constant output. It causes part of the output wave to vary the amplifier gain so that the amplifier output level is maintained constant although the signal input level may vary over a considerable range of values. The resistances 44 and 46 are in series in one side of the feedback path f and the resistances 45 and 47 are in the other side. Where symmetry (balance to ground) is not required, the resistances in either side may be omitted.

The resistance of silver sulphide has such a large negative temperature coefficient that passage of even a small current will heat it up sufficiently to reduce its resistance as compared to the value of its resistance at ambient temperatures. In the operation of the system, at sufficiently light loads the feedback currents do not heat the silver sulphide appreciably; and, as a result, the silver sulphide resistance introduces nearly a constant insertion loss and, therefore, will not appreciably vary the amplifier gain.

As the applied input is further increased, due to self-heating, the resistance of the silver sulphide decreases; and this reduces its insertion loss and therefore the amplifier gain. Thus, there is a tendency for the gain to decrease as the input increases, which effect in the proper proportion can hold the output constant. By adding small resistances 46 and 47 in series with the silver sulphide resistances, this proper proportion is maintained, so that there is a considerable range of values of input for which the output is independent of the input, and in these instances the gain of the feedback amplifier goes down decibel for decibel for each decibel increase in input. In a specific amplifier having a feedback path connected across the secondary winding of the output transformer and across the primary winding of the input transformer, and with a silver sulphide resistance and a constant resistance connected in series with each other and in series in the feedback path for controlling volume in the same general way as that indicated in Fig. 7, the output level was found independent of the input level to a precision of 0.2 decibel over a range of 1 to 15 decibels, and to a precision within $\pm.05$ decibel over a range from $+1$ to $+14$ decibels, and to a precision within $\pm.01$ decibel from $+2$ to $+9.5$ decibels. This was for a particular frequency, 5000 cycles per second. However, inasmuch as silver sulphide resistances are independent of frequency from very low frequencies to very high radio frequencies, the characteristic effect described as obtaining for one frequency is the same for other frequencies in the useful range of the feedback amplifier. Therefore, in a well-designed amplifier, the gain-frequency curves for various specified values can be made parallel and flat over the useful range.

To obtain the precisions mentioned above, the constant resistance in series with the silver sulphide resistance was given an optimum value, which in this particular instance was 210 ohms and amounted to about 3 per cent of the resistance of the silver sulphide unit at room temperature.

While silver sulphide resistance and a constant resistance in series, bridged across a circuit, can hold the circuit voltage constant over a considerable range of variation of the internal voltage of the generator feeding the circuit, the range over which the volume can be limited is greater in the case of a circuit including an amplifier in the general manner indicated in Fig. 7. That is, the amplifier increases the useful range of the volume limiting circuit. Further, the useful range of operation, and also the precision of the control, can be considerably extended by refinements in the design of the silver sulphide resistance and in the design of the feedback circuit associated with it.

The output voltage at which regulation takes place in Fig. 7 can be reduced by decreasing the volume of the silver sulphide unit. Another way to increase the sensitivity is to elevate the temperature of the unit, for example to a temperature just below that at which it begins to regulate, by an auxiliary or indirect heater such for instance as the heater **26** shown in Fig. 6.

The useful range of the volume limiting system of Fig. 7 can be increased by connecting volume limiting devices, such as the device shown, in tandem with each other in the feedback path f, and adjusting one to operate when the limit of the useful range of the preceding device has been reached; or the resistances **46** and **47** may each be made a silver sulphide resistance and a constant resistance in series, these additional silver sulphide resistances then being adjusted to begin their regulating or limiting action when the volume limiting device shown has reached the limit of its useful range.

Aside from its simplicity the circuit of Fig. 7 has a desirable property not possessed by the ordinary current or volume limiter type of circuit. In the circuit of Fig. 7, as the limiting increases, i. e., gain decreases, the harmonic level also decreases. In other words, the greater the limiting, the better the transmission performance generally. This is a very desirable action and is opposite to the action of the ordinary type of circuits of this class.

Since modulation originating in the feedback path or β-circuit is not improved by feedback and appears in the amplifier output, it might be questioned whether, as the alternating current flowing through the silver sulphide in Fig. 7 varies from instant to instant, the resistance of the silver sulphide might not vary correspondingly and thereby produce serious modulation. Fundamentally, the answer depends upon the speed with which the silver sulphide heats up and cools off. For instances where, in the heating and cooling of the silver sulphide, the steady state is reached in several milliseconds, or where the unit heats faster but takes much longer to cool off, measurements indicate that the modulation is about 100 decibels down on the fundamental, at least for frequencies that have been measured, which includes a frequency as low as 1000 cycles per second.

The speed with which a circuit of this character may be made to operate automatically to control the gain of the feedback system may be very fast or very slow, for example, apparently from as fast as of the order of 1/10,000 second to as slow as more than ten minutes. The speed of operation is controlled chiefly by changing the physical size, and mechanical and thermal design of the silver sulphide resistance unit itself.

Silver sulphide changes its resistance with temperature apparently faster than any other known material which simultaneously would satisfy the additional practical requirements that such a resistance be stable, that its temperature effects and behavior be accurately reproducible and that it be capable of being manufactured cheaply to fall within closely specified limits. Silver sulphide meets these latter specifications and in addition, for all temperatures below 179° C., its resistance is halved every time its temperature is increased approximately 13.9° C. At 179° C. its resistance is abruptly divided by more than 40, and for still higher temperatures its resistance increases.

Due, for temperatures below 179° C., to the large and negative temperature coefficient of the resistance of silver sulphide, after a voltage applied to such a resistance exceeds a certain critical value depending upon the temperature and material surrounding the silver sulphide and upon the thermal properties of the silver sulphide unit itself, the current flowing will commence to increase abruptly, due to excessive self-heating of the silver sulphide, and this increase will continue until finally limited by any resistance or impedance in series with the applied voltage or until the temperature of the unit has reached 179° C.

For example, for a fixed ambient temperature, if a very small voltage be applied in series with a silver sulphide resistance, a very minute current will flow in response to the applied voltage. This flow of current will cause power i^2 R, to be dissipated as heat which will raise the temperature of the silver sulphide and thus reduce its resistance, the offset of this can be shown to increase the current and, therefore, further raise the temperature of the silver sulphide. The total energy this represents in any arbitrary interval of time will be in the form of heat energy, and, if by conduction, convection and radiation this heat is transferred to the surroundings at a faster rate than it is supplied, the dynamic system will be stable and the silver sulphide will assume an equilibrium temperature slightly in excess of its ambient temperature. For further increase in voltage, the equilibrium temperature is further increased and the rate of increase of the latter can be shown to exceed the former. The net result of all this is that if the current is plotted against voltage applied, the resulting curve is concave upward and lies above the straight line representing the relationship between current and voltage for a fixed resistance whose value is equal to that of the resistance of the silver sulphide at its ambient temperature.

However, as the applied voltage is further increased, the main effect occurs. That is, at some critical value of voltage the heat cannot be dissipated by conduction, convection and radiation as rapidly as it accumulates and at this trigger value of voltage the current begins to increase without limit until the temperature of the silver sulphide reaches and exceeds 179° C. at which temperature its positive coefficient eventually operates to prevent further increase in temperature, reduction of resistance, and increase of current.

Appropriate analysis disclosed that the trigger action above described is duplicated in a general sort of way if impedance is inserted in series with the applied voltage source provided the added impedance is not obviously too great. Theoretical analysis also showed that a resistance, such as 46 or 47, of proper value and in series with the silver sulphide greatly improves the operation of the circuit of Fig. 7, and experiment demonstrated the improvement.

The input or trigger voltage at which the automatic regulating system of Fig. 7 operates and, therefore, also the value of fixed output, is somewhat reduced as the room temperature increases. For many applications this effect is not enough to be troublesome. Where the operation needs to be precise, the silver sulphide resistance can be housed in a container maintained at constant temperature in the manner indicated in Fig. 6.

When the volume of the container is sufficiently small it will not greatly affect the heating time but will increase the cooling time to a value as great as desired, depending upon the insulation. For example, a volume limiter to take the place of the present seven-tube limiter used in commercial type C carrier communication systems would have to operate in two to three milliseconds and release in more than 0.8 second. In a circuit such as shown in Fig. 7 this fast-operate time would be obtained by making the volume of the silver sulphide unit sufficiently small and the release time adjusted by the amount of heat insulation immediately surrounding the unit. When the surrounding heat insulation is sufficiently far from the silver sulphide it merely has the effect of putting the silver sulphide in a heat-insulated room and the times of heating and cooling are not affected. However, placing the silver sulphide in an insulated chamber sufficiently large (relative to the silver sulphide unit itself and the power it dissipates) greatly reduces the effect upon it of fluctuations in room temperature, and materially reduces its slow temperature changes, also.

The volume limiting circuit of Fig. 7 is of general application. For example, it is suitable for use as a receiving amplifier in a carrier transmission system, automatically keeping the voice frequency equivalent constant independent of high frequency variations; as a substitute for the volume control circuits now used in radio receiving circuits; as a volume limiter in front of a loudspeaker, to prevent overloading; as a volume limiter to prevent overloading a detector, a group modulator or a broad band amplifier carrying a number of channels (the effect in this case resulting in an important increase in the useful operating level of the amplifier or repeater); as a control to keep the pilot current fixed at the sending end of a pilot channel system; as a control to maintain the output of multi-frequency carrier supply systems at a constant voltage; as a control to limit the volume of voice frequency telegraph signals superimposed on carrier telephone systems; or as a control for an alternating current supply voltage to render the voltage constant for operation of alternating current apparatus or for testing or measuring purposes.

The hybrid coils in Fig. 7 not only render the amplifier input and output impedances independent of the connecting circuits, but, in this volume limiter, the hybrid coils prevent the amplifier impedances from being changed by changes in the amount of feedback that result from changes of input level.

It is noted as to the operation of the hybrid coils in a circuit such as Fig. 1, that when the β-circuit network 3 is a network for amplitude or phase equalization or correction of distortion (in the general manner disclosed for example in my above-mentioned article, Patent 1,956,547, or copending application 606,871, or in British Patent 371,887), the hybrid coils render the equalization independent of the flexibility of the amplifier input and output impedances, the impedances of the amplifier being adjustable through very wide ranges by means of the hybrid coil nets 9 and 10 without affecting the equalization or distortion correction. This is especially noteworthy since locating the equalizing or transmission controlling network in the feedback path or β-circuit has important advantages. For example, whereas in certain instances it is practical to design a corrective network having the same transmission characteristic as the apparatus or system to be equalized, the design of a network having the inverse characteristic may involve negative elements; and therefore by equalizing in the feedback path compensation can be effected for transmission distortion which, without resort to feedback action, could not be corrected in any known manner using physically realizable elements.

Another advantage concerns effect upon signal-to-noise ratio. If the received signal has been propagated over a transmitting medium having more loss at some frequencies than others, and if the general noise level is of the order of the noise due to thermal agitation, a substantial improvement results from performing the equalization or other transmission control in the feedback path. This can be appreciated by observing that the introduction of a loss of x decibel in front of the input to the amplifier degrades the signal-to-noise ratio x decibel, assuming the introduction of such loss leaves the resistance component of the equivalent passive series impedance from grid to filament of the first tube unchanged. Thus, if the decibel attenuation of the transmitted signal in being propagated over the transmitting medium is, for example, 30 decibels greater at the highest frequency than at the lowest, a line equalizer in front of the amplifier would be required to have 30 decibels more loss at the lowest than at the highest frequency. In a constant-resistance type of equalizer, this would degrade the signal-to-noise ratio of the lower frequencies something in excess of 30 decibels. These relations only hold for resistance noise which is unaffected by insertion of a constant-resistance network in front of the amplifier.

A further advantage is increase of allowable "head-end" loss of the equalizer or corrective network. In practical designs, at the frequency at which the insertion loss is required to be least, actually there is a finite decibel loss. Usually, when the equalizer is located in the line, this loss is required to be a minimum. However, if the equalizer is located in the β-circuit, in most instances this loss can be considerable without affecting the overall transmission performance. Generally speaking, increasing "head-end" loss will permit appreciable economies in size and cost of the parts.

In the case of pre-equalization there is an advantage with respect to load rating. If the equalizer is located in the β-circuit and, in addition, the cable or transmitting medium is free from practically all noise but resistance noise, pre-equalization will result in a substantial number of decibels of improvement in the level rating of the amplifier depending upon the band width, number of channels, and attenuation-frequency characteristic of the medium. As compared to locating the equalizer in front of the amplifier, the input levels can be reduced by the total amount of loss of the equalizer for either high or low frequencies. For the case of a cable this would mean reducing the levels at the lower frequencies by amounts corresponding to the difference between the loss at the highest frequency and the lesser loss at the lower frequencies plus an additional reduction at all frequencies equal to the head-end loss. As a result of reducing the level of some frequencies more than others, the load rating of the amplifier is improved.

Another advantage is reduced modulation requirements. If the pre-equalization is one-half of the cable slope, attenuation vs. frequency, instead of the full amount as just discussed, a worthwhile portion of the improvement in level rating of the amplifier is still retained. However, with pre-equalization, if and only if the equalizer is in the β-circuit, the modulation requirements are uniformly reduced at all frequencies in the transmitted band by an amount approximately equal to one-half the difference in the cable attenuation at the highest and lowest used frequencies. For the same performance as before, this leads to a worthwhile reduction in the gain without feedback which is more important the wider the frequency band transmitted and the higher the top frequency.

Even without pre-equalization, the performance can be improved by the β-circuit location of the equalizer. If the cable noise varies with frequency and increases for each frequency the same number of decibels the cable attenuation decreases, pre-equalization is not possible. In this case the improvement in level rating and reduction in modulation requirements as just mentioned cannot be obtained. However, the performance can be improved, or for equal performance less feedback will be required. For example, if the attenuation is 30 decibels less at the lowest than at the highest frequency, the negative feedback is 30 decibels greater at this frequency due to the presence of an equalizer in the β-circuit. The result, as compared to not locating the equalizer in the β-circuit, is the gain without feedback for equal performance does not have to be held at so high a value over so broad a frequency band. This leads to a higher gain per stage for the top frequency and this gain increase is comparable to that of the preceding paragraph.

An outstanding advantage of the β-circuit location for the corrective network relates to speeding up transmission. As shown in my above-mentioned article or copending application 606,871, or in British Patent 371,887, the amplification of a feedback amplifier with $\mu\beta >> 1$ is approximately

$$\frac{1}{-\beta}$$

so when the β-circuit is made such that its propagation is the same as that of the line or cable between a source of voltage e and the input to the μ-circuit, the transmission from the source of voltage e to the output of the μ-circuit is

$$\frac{\beta}{-\beta} = -1$$

(as brought out for example in the disclosures just mentioned or in my copending Patent 2,002,499, May 28, 1935, or F. A. Cowan Patent 2,017,180, October 15, 1935), and hence there is no delay nor distortion and the amplified signal appears instantly at the output of the μ-circuit as a replica of the signal e applied to the line or cable by the source, except reversed in sign. (There is no restriction on β other than that the amplifier comply with Nyquist's rule.) This result is obtained from theoretical considerations. The equations which are available for arriving at this conclusion are rigorous only for systems containing lumped constants. The transmission through systems made up of continuous elements is only approximated from the lumped constant equations. Hence to this extent the above procedure will correct for amplitude distortion and phase distortion. (Applications are encountered where there are likewise important advantages in having the β-circuit correct solely for phase distortion.)

Compared to the customary way of improving cable distortion this procedure speeds up transmission. The customary way, where the network for correcting attenuation and phase over the frequency band of interest is not in a feedback path, delays the time of transmission by a period that exceeds the time required for a particular frequency to travel down the cable, this particular frequency generally being that one in the band of interest which is most slowly transmitted. In contrast to this, for the described procedure of correcting in the β-circuit, the time is the time for a particular frequency, which is the one most rapidly transmitted, or in other words, it is the time required for current no matter how trivial to make its appearance at the receiving end. All other velocities for all other frequencies for which the circuit is properly operative are made equal to this most rapid speed. In the case of a cable, this speed apparently corresponds to a speed less than the velocity of light in a vacuum, depending upon the dielectric constant and permeability of the cable. It should be noted that this time is independent of the wave form impressed at the sending end. It is also independent of lumped series inductance, either positive or negative, and likewise is independent of lumped shunt resistance, either positive or negative. Thus, by adding apparatus (the feedback amplifier) to a heavily loaded voice frequency cable, a wave can be propagated over the cable at the same speed as over the same cable non-loaded.

Assuming the cable is many wave-lengths long, preferably the β-circuit instead of being a like section of cable may have a different indicial admittance, as now to be explained.

Considering indicial admittance of a transmission system as a transfer admittance, it is the current as a function of time at the far end in response to a unit voltage suddenly applied at time $t=0$ at the sending end, and two systems having like indicial admittances will behave identically at the receiving end for like excitations at the sending end. If two systems have like indicial admittances their steady state amplitude and phase characteristics are identical. If two systems have like amplitude and phase characteristics (that is, steady state attenuation and phase vs. frequency characteristics) over a specified band of frequencies, for example f_1 to f_2, then the received currents within this band probably will be the same for each system for like excitations at the sending end.

Nevertheless two systems can have very nearly identical steady state attenuation and phase characteristics from $f=0$ to $f=\infty$ and yet the time taken for some current, no matter how little, to arrive at the receiving end can be entirely different.

For example, two such systems are on the one hand a cable of length l where l is many wavelengths and on the other hand a lattice network of one section whose four elements are two impedances each equal to the cable impedance of length

$$\frac{l}{2}$$

with far end short-circuited and two other impedances each equal to the cable impedance of length

$$\frac{l}{2}$$

with far end open. The lattice network could be made to match the cable over a specified band as regards steady state amplitude and phase, and yet have the time for the first current to arrive relatively short compared to the case of the cable. (This assumes, for the case of the actual cable, that current can not appear quicker than would correspond to the velocity of light.)

The real time of transmission of a cable will be designated τ and equals

$$\frac{l}{v}$$

where v is the velocity of propagation referred to above, i. e., τ is the time for the first current to appear at the distant end.

When a feedback amplifier is to correct for a section of cable in front of the amplifier in the general manner referred to above, preferably the β-circuit should have an indicial admittance equal to the indicial admittance of the cable when, referring to the admittance characteristic of the latter, t is replaced by $(t-\tau)$. If this is done and $\mu\beta>>1$, the output of the amplifier will be a replica of the input at the sending end except it will appear τ seconds later.

If the β-circuit be a like section of cable (assuming Nyquist's rule is satisfied by appropriately modifying μ, or the cable itself is modified by making additions) the time of transmission is still τ but for an interval τ at the beginning and a corresponding interval at the end, the wave form apparently will be incorrect. This effect would be something more than multiplying the time of transmission by two as compared to the method of the preceding paragraph. It would make the received wave distorted in a manner analogous to that of a loaded cable.

To give a physical concept of the method described in the second preceding paragraph for correcting phase distortion the response without feedback to the excitation can be viewed as the forced or steady state response plus the free flow or transient response. As a result of feedback action the amplifier does not amplify the transient response for the conditions above described. Thus, the first currents to arrive of necessity carry information as to the signal impressed so that the amplifier sends out a copy of the signal wave form applied at the sending end, and, when the main body of the transmitted signal arrives later and appears at the input to the amplifier, it just is not amplified.

When an attenuation equalizer or other corrective network is located in the feedback path between the hybrid coils as in the case of the β-circuit network 3 of Fig. 1, since the corrective network does not affect the amplifier input or output impedance the equalization or correction can be accomplished by merely inserting a simple series or shunt arm between two arbitrary fixed impedances. Usually this requires less than half as many elements as the corrective network requires when it is so located that it affects the amplifier impedance seen from the line.

Moreover, when an attenuation equalizer or phase corrector is to be located between the hybrid coils as in Fig. 1, the impedances between which it is to insert a specified loss or phase shift can be conveniently caused to assume a very wide range of values, as much as 1000:1. This element of flexibility is an aid in producing a low cost design.

In Fig. 2, if $\mu\beta>>1$, assuming C is made equal to Z_F there is a decibel gain from C to Z_N equal to

$$10 \log_{10}\frac{C}{Z_N}.$$

Therefore, a system, such for example, as that shown in Fig. 8 (which is dominated by R. S. Caruthers application Serial No. 114,409, Patent 2,166,929, issued July 25, 1939, entitled Electric wave amplifying systems, filed of even date herewith) can be built, comprising an amplifier (with $\mu\beta>>1$) that works on a matched impedance basis from a resistance R_1 into a greater resistance C_2 and has its gain

$$G_P=10 \log_{10}\frac{R_1}{R_2}$$

In Fig. 8 the amplifying element is designated **50**. It may be of any suitable type, as for example, a vacuum tube device such as the amplifying device indicated in Fig. 2. Hybrid coil 51 in Fig. 8 corresponds to hydrid coil 5 of Fig. 2, the load or receiving circuit R_2 in Fig. 8 corresponding to the network Z_N of Fig. 2.

In the circuit of Fig. 8,

$$Z_{F1}=\frac{n_1+n_2}{n_2}R_2$$

and

$$Z_{F2}=\frac{n_2}{n_1+n_2}R_1$$

So if

$$\frac{n_1+n_2}{n_2}$$

be made equal to

$$\frac{R_1}{R_2}$$

then

$$Z_{F1}=R_1,\ Z_{F2}=R_2,\ \text{and}\ G_F=10\log_{10}\frac{R_1}{R_2}$$

In transmitting in the opposite direction through the system, that it, from R_2 to R_1, the loss is precisely the same as the previous gain (i. e., from R_1 to R_2). In either direction, the gain or loss, as the case may be, is independent of frequency by virtue of the method of operation of the circuit.

The circuit of Fig. 8 can be rearranged as in Fig. 9, wherein

$$\frac{n_1+n_2}{n_2}=\frac{R_2}{R_1}$$

to work from a resistance R_1 into a resistance R_2 that is greater than R_1, giving a gain

$$G_F=10\log_{10}\frac{R_2}{R_1}$$

and giving a loss from R_2 to R_1 equal to this gain G_F. In Fig. 9 the amplifier **50'** and the hybrid coil **51'** correspond to the amplifier **50** and hybrid coil **51** of Fig. 8.

Figs. 8 and 9 can transform impedances, in the manner indicated in the discussion above of Figs. 1 and 2, the circuit of Fig. 9 dividing an impedance by a number greater than unity.

Fig. 10 shows an amplifier circuit obtained by combining two circuits such as those of Figs. 8 and 9. The input impedance Z_{c2} and output impedance Z_{c1} of the amplifying system equal respectively the impedance Z_{c2} of the output connecting circuit and the impedance Z_{c1} of the input connecting circuit. Without the pad **52** shown in the path joining the two amplifying elements, the amplifying system has a gain (from west to east) equal to

$$20\log_{10}\frac{n_1+n_2}{n_2}$$

and has a loss (from east to west) equal to

$$20\log_{10}\frac{n_1+n_2}{n_2}$$

If desired, to reduce the gain without affecting the impedance relationships, the pad **52** of impedance level Z_n can be inserted as shown.

The circuits of Figs. 8 to 10 show that power is dissipated in stabilizing an impedance by feedback as in Fig. 2. In the case of Figs. 8 to 10 (with network **52** omitted) all such power is dissipated in the output load.

Certain of the circuits described hereinafter employing hybrid coil feedback connections, such for example, as those shown in Fig. 1, employ types of feedback that will be called multiple feedback and repetition of the feedback process. The significance of these terms will now be indicated.

As discussed in my above-mentioned copending application 606,871, it is often desirable, especially in analysis and design of feedback systems, to distinguish between systems restricted to a single feedback loop and systems having a plurality of loops or in other words multiple loop feedback systems. A feedback loop with more than one path from the same output to the same input is a multiple loop, but as noted in that copending application, if the multiple loop can be theoretically replaced by a single loop, the feedback is regarded as parallel feedback, except that a special case of parallel feedback is termed repetition of the feedback process (referred to below). As also noted in that copending application, multiple feedback is arbitrarily viewed as including those cases of feedback in which a single $\mu\beta$-loop includes additional local feedback loops and thus comprises a collection of feedback systems having elements in common and satisfying the condition that when the over-all system is viewed analytically in the form of an equivalent circuit, it is not possible to obtain further reduction in the number of local feedback loops surrounded by a single $\mu\beta$-loop. As further noted in that copending application, a repetition of the feedback process is considered to occur whenever a complete feedback system (single loop or multiple feedback) may be viewed and treated as a unit and, in addition, is used to form a new $\mu\beta$-path which path in every way is independent of the first feedback system except in so far as it utilizes the original over-all properties of the first feedback system. This requires conjugacies or their equivalent and the use of transformers, assuming no new unilateral devices or their equivalent be added. As shown hereinafter, repetitions of the feedback process are but a special case of parallel feedback, namely, with an added restriction regarding conjugacy. Accordingly, unless active elements are used to separate the feedback paths the various paths or loops theoretically can be replaced by a single path or loop. However, repetition of the feedback process can have practical advantages, for example, as pointed out hereinafter.

Suppose, for example, it is desired to feed back through the input and output transformers and suppose further that 60 decibels of feedback is required but that with the particular transformers being used only 25 decibels of feedback is possible through the coils without getting into singing troubles. Now by putting 35 decibels of feedback inside the coils and 25 decibels outside the amplifier can be designed so as not to sing. Suppose further it is desired that the input and output impedances be exceptionally good. By using repetitions of the feedback process instead of some other type of multiple feedback, the deviations of these impedances from the ideal sought will be improved roughly fifty-fold.

Fig. 11 is a simple example of a multiple feedback amplifier employing a hybrid coil feedback connection, the feedback path f connecting to the output of the amplifier through hybrid coil **7** and being in series with the secondary winding of a two-winding input transformer **60** that connects the incoming line **6** to the amplifier. Thus this feedback through path f is a series or current feedback at the input of the amplifier and a hybrid coil feedback at the output of the amplifier and may be referred to as a series-hybrid feedback. The amplifier has tubes **55** and **56** coupled in tandem through an interstage coupling circuit comprising tuned circuit **61**, stopping condenser **62** and grid leak **63**. A resistance or impedance **64** common to the input and output circuits of tube **55** provides a local or internal series-series feedback around tube **55**. Thus this local series-series feedback loop or system is surrounded by the main and single $\mu\beta$-loop. The series-series internal loop around tube **55** is paralleled by a second loop comprising tube **56** and its output hybrid coil, etc. Therefore, with respect to the input and output of tube **55**, the circuit of Fig. 11 is an example of a double loop multiple feedback system. Although double loop systems are often multiple feedback systems, it is evident that from some aspects this particular double loop system around tube 55 could be replaced by an equivalent single loop system. However, in the case of an equivalent single loop system, the modulation at the output of tube 55 would be different and, of course, with respect to the over-all amplifier from C to the load L fed by the outgoing circuit 8, the entire situation would be very much altered.

The local feedback provided by resistance 64 may be negative feedback with $\mu\beta>>1$ for the local loop. The feedback from hybrid coil 7 around the main feedback loop may be negative feedback with $\mu\beta>>1$ for this main loop; or, if desired, it may be made positive feedback by the poling of the hybrid coil. The positive feedback may be desired, for example, to produce over an appreciable frequency interval an increase in gain or to cause the change in phase due to feedback to increase at low frequencies in order to reduce delay distortion. The internal negative feedback, (and also the external feedback, if negative) may be desired, for example, to flatten the gain-frequency characteristic, as for example, in the case of a voice frequency amplifier.

The procedure for derivation of equations for the transmission performance of multiple feedback systems can be along exactly the same lines as followed, for example, in my above-mentioned article and copending application 606,871. However, the criterion for singing may be different. For example, in certain multiple feedback systems, to avoid singing of the over-all system, or in other words to meet the requirement that the free response of the system as a whole will not be a wave expanding with time, it is necessary that the $\mu\beta$ characteristic of an individual single loop do not inclose a unit circle about the point $1,<0$ as a center instead of the point $1,<0$ itself as given by Nyquist's rule for a single loop. In other examples, certain loops will be unstable if other loops are opened, etc.

Fig. 11A shows an amplifier which is a modification of the circuit of Fig. 11, the potential, with respect to ground, that is fed back from the hybrid coil to the grid of the first tube being applied across resistance 64. The voltage across resistance 64 is transmitted also to the grid of the second tube through the plate-cathode path in the first tube and the interstage coupling circuit, before phase reversal in the first tube, but the voltage thus applied to the grid of the second tube is small in view of the high impedance of that plate-cathode path. The amplifier is suitable, for example, as a high quality voice frequency amplifier for program transmission circuits. The tubes are shown as a screen grid tube 55′ and a coplanar grid tube 56′ which may be Western Electric Company 259A and 281A tubes, for example. The space current for tube 55′ is supplied through a choke coil 61′ and the negative bias for the grid of tube is supplied partly by battery C_1 and partly by the resistor 64 which is unby-passed and, therefore, produces local negative feedback around tube 55′. Battery C_2 supplies the screen biasing voltage for this tube, the control grid and coplanar grid biasing voltages for tube 56′ being supplied by batteries C_3 and C_4. As in the case of Fig. 11, the outer feedback may be made either negative or positive by poling of the hybrid coil.

Fig. 12 shows a repeater with negative feedback through hybrid coils, the system being an example of a system not viewed as a multiple feedback system though a portion of the β-circuit is common to two feedback paths. Viewing the system as a single loop feedback system, the value of $\mu\beta$ is other than zero for each direction of transmission, clockwise and counterclockwise. RE is an amplifier for transmission from line LW to line LE; and RW is an amplifier for transmission from line LE to line LW. Transformers 65 and 66 are connected to operate as a hybrid coil 67; and, similarly, transformers 68 and 69 form a hybrid coil 70, transformers 71 and 72 form a hybrid coil 73; and transformers 74 and 75 form a hybrid coil 76. A β-circuit network 77 may provide any suitable control for transmission through the repeater. Networks 78′ and 78 are balancing networks of hybrid coils 73 and 76, respectively; and networks 79 and 80 are balancing networks for hybrid coils 67 and 70, respectively.

Transmission from line LW arriving at hybrid coil 73 is partly dissipated in network 78′ and partly transmitted through hybrid coil 73 to hybrid coil 67. The transmission arriving at hybrid coil 67 is partly dissipated in the output of amplifier RW and partly transmitted through the hybrid coil 65 to amplifier RE and thence to hybrid coil 70. Part of the transmission thus reaching hybrid coil 70 is dissipated in network 80 and part is transmitted to hybrid coil 76. Of this latter part, one portion goes through the hybrid coil to line LE, and another portion goes through the hybrid coil and the β-circuit network 77 to the hybrid coil 73, there dividing between network 78′ and hybrid coil 67. Thus, waves from the output of amplifier RE are fed back through hybrid coil 76, β-circuit network 77 and hybrid coil 73 to the input of amplifier RE.

Similarly, transmission from line LE is amplified by amplifier RW and transmitted to line LW, and the amplifier RW feeds back through hybrid coil 73, β-circuit network 77 and hybrid coil 76 to amplifier RW.

Fig. 13 shows a second example of a multiple feed back amplifier with a feedback connection from an output hybrid coil 7. There is a triple loop path (and, as in the case of the circuit of Fig. 11, it is not apparent how the multiple loop could be replaced by a single loop connection).

The amplifier has a two-winding input transformer 81 with primary and secondary windings 82 and 83. Winding 82 forms one arm of an input bridge 84, the other arms being resistances or impedances 85, 86 and 87. The incoming line 6 is in one diagonal of this bridge, and the other diagonal includes a feedback path, which will be called the β_2 path.

The grid-cathode impedance of the first tube of the amplifier forms one arm of a second input bridge 88, the other arms being resistances 89, 90 and 91. The winding 83 is in one diagonal of this bridge, and the other diagonal includes a feedback path, which will be called the β_1 path, this feedback path comprising a two-winding transformer 92 and also comprising a β-circuit network 93 for controlling transmission through this feedback path in the same general manner as explained above for the case of network 3 in the feedback path in Fig. 1.

Bridges 88 and 84 need not be balanced. As pointed out in my above-mentioned copending application Serial No. 663,317, there is a 6-decibel advantage (3 decibels at the input and 3 decibels at the output) in unbalancing the bridges. Considerable amounts of negative feedback around the $\mu\beta_1$ path and the $\mu\beta_2$ path can give practical conjugacy between the diagonals of either or both of these bridges even if one or both of the bridges are unbalanced.

The plate-cathode impedance of the last tube of the amplifier forms one arm of an output bridge **94**, the other arms being resistances **95**, **96** and **97**. The β_2 path, including a two-winding transformer **98** is connected across two opposite corners of this bridge, and the output hybrid coil **7** is connected across the other corners.

It can be seen that, in addition to the $\mu\beta_1$ path and the $\mu\beta_2$ path, there is a third path, which may be called the $\mu^2\beta_1\beta_2$ path, this third path surrounding the other two. That is, transmission through this third path is through the μ path, thence through the β_1 path, thence through the μ path again, and thence through the β_2 path back to the μ path.

The amplification with feedback is

$$A_F=\frac{\mu}{1-\mu^2\beta_1\beta_2-\mu\beta_1-\mu\beta_2}$$

From this it is to be noted that if $\mu\beta_1$ and $\mu\beta_2$ are both much greater than unity, the operation may be viewed merely as using the amplifying path or μ path twice, that is, once for μ and once to increase β; and in this case A_F is approximately

$$\frac{1}{\mu\beta_1\beta_2}$$

The transformers **92** and **98** may increase the longitudinal balance of the circuit. Moreover, since they insulate the incoming line, the amplifier and the outgoing line from one another for direct current, any desired combination of these three circuits can be grounded for direct current, and further the network **93** or other such transmission control networks in β_1 or β_2 circuits can be balanced or unbalanced, such for example as T or H networks or lattice networks, if desired.

Fig. 14 illustrates a repetition of the feedback process five times. This figure is Fig. 38 of my above-mentioned copending application Serial No. 606,871 with the reference numerals increased 100. In using feedback to improve modulation (i. e., to reduce modulation products relative to fundamental), the amount of improvement that a single feedback process can yield depends upon how well the specific amplifier or system employed is adapted to this purpose. For example, one amplifier gave an improvement of 60 decibels, yet another amplifier, with only one feedback path, corresponding to path P_1 of Fig. 14, gave 20 decibels improvement and then reducing the gain beyond 20 decibels did not result in an improvement in hormonics corresponding to each decibel the gain was further reduced. However, by stopping the first step of gain reduction and distortion improvement at 20 decibels, and employing a second feedback path P_2, as shown in Fig. 14, to secure a second step of gain reduction and distortion improvement, this second step of improvement will now be effective starting with 20 decibels (the end of step 1) and in this case will now carry on effectively over a range of more than 20 decibels additional reduction in gain. The reason for this may be seen from the following considerations. The effectiveness of the first step in improving the distortion of the circuit is limited by the distortion present in the output transformer **126**. By including this source of distortion in the $\mu\beta$ path of the second feedback path P_2, the second feedback path can not only continue to reduce the distortion in the tube but can also reduce that in the coil, giving a greater proportionate improvement than could result from merely increasing the feedback of P_1. This supplemental action is conditioned upon an effective conjugacy between the feedback paths P_1 and P_2. The feedback P_1 aids in enabling this conjugacy to be realized, for it gives an improved impedance on the secondary side of transformer **126**, looking back toward R_0, with which to balance the bridge across which feedback P_2 is connected. This impedance is indicated in Fig. 14 as R_{02} and is, in effect, a new "R_0" for the feedback P_2 analogous to the R_0 of the tube for the feedback P_1. Thus, repetitions of the feedback process actually differ from feedback around a single path, and even make it practicable to extend the advantages of feedback to a degree otherwise unattainable. In applying repetitions, it is not necessary that each step be directed toward the same objective.

In Fig. 14, the feedback process with amplifier **130** is carried out a third time, through feedback path P_3, a fourth time through feedback path P_4 and a fifth time through feedback path P_5. The amplifier **130** has output and input bridge circuits **131** and **132**. These bridges render path P_1 and transformer **126** conjugate, and render path P_1 and transformer **133** conjugate, and consequently path P_1 is conjugate to paths P_2, P_3, P_4 and P_5 (and vice versa) since they are connected to path P_1 through the transformers **126** and **133**.

Similarly, bridges **134** and **135** render path P_2 conjugate to (transformers **136** and **137** and consequently to) paths P_3, P_4 and P_5 (and vice versa); bridge transformers **136** and **137** and their bridge impedances **138** and **139** render path P_3 conjugate to paths P_4 and P_5 (and vice versa); bridges **140** and **141** render path P_4 conjugate to (transformers **142** and **143** and consequently to) path P_5; and the networks **144** and **145** (which are forms of bridge circuits with their ratio arms and diagonals composed of transformer windings) render path P_5 conjugate to the incoming and outgoing circuits of the amplifier (and vice versa). Since the conjugacy of P_4 to transformers **142** and **143** satisfies the requirement for repetition of the feedback process by P_5, conjugacy of P_5 to the incoming and outgoing lines need obtain only to the extent desired for reducing line-to-line transmission through P_5 and effects of line impedances on feedback and of feedback on amplifier impedances. Thus the turns ratio of the two portions of the divided winding of transformer **142** may differ from that of the two portions of the divided winding of the output transformer, and similarly the turns ratio of the two parts of the divided winding of transformer **143** may differ from the corresponding turns ratio for the input transformer secondary, sufficiently to provide the desired amount of feedback through P_5.

The ratio arms R_0, KR_0, KR and R in bridge **131** correspond to the arms R_{02}, $K'R_{02}$, $K'R_2$ and R_2 respectively in bridge **134**, and to the arms R_{03}, K_3R_{03}, K_3R_3 and R_3 respectively in bridge **140**.

Repetition of the feedback process is not limited to negative feedback. For example, positive feedback for increasing gain and gain stability can be repeated to obtain greater increases than are feasible without the repetition.

The bridge transformers or hybrid coils **136** and **137** with their nets **138** and **139** operate in the general manner described for the hybrid coils in preceding figures; as for example, Figs. 1 and 2.

Fig. 15 shows a second example of repetition of the feedback process, with the repetition taking place through an outer feedback path *f* from an output hybrid coil 7 to an input hybrid coil 5. The amplifier comprises tandem connected tubes 151, 152 and 153, and has an output bridge 154 joining tube 153 to the hybrid coil 7 and to an inner feedback connection 155. This connection 155 applies the voltage from the bridge 154 across a feedback impedance in series with the secondary winding of the input hybrid coil with respect to grid-cathode impedance of tube 151. Thus, the inner feedback is a series-bridge feedback. If the feedback lead 155 be connected to the top, instead of the bottom, of the secondary winding of the input hybrid coil, the inner feedback becomes a shunt-bridge feedback, the system still illustrating repetition of the feedback process. However, in either case, if the bridge 154 be omitted, the feedback lead 155 being connected for example to the plate of tube 153, the system no longer exemplifies repetition of the feedback process.

The four ratio arms of the bridge 154 are the anode-cathode impedance of tube 153, resistances 156 and 157, and impedance 158. The arm 158 comprises a resistance 159 in parallel with a capacity 160 (of the order of 50 micromicrofarads, for example) for controlling the slope of the attenuation-frequency characteristic of the inner feedback path, to control the phase shift above the transmitted band. The bridge 154 may be unbalanced, if desired, with resulting advantages brought out, for example, in my above-mentioned copending application Serial No. 663,317.

The feedback connection 155 includes a stopping condenser 161 for preventing passage of direct current through this connection. The feedback impedance in series with the secondary winding of hybrid coil 5 and included in the feedback diagonal of bridge 154 comprises two series connected elements: first, a feedback resistance 162; and second, a feedback resistor 163 in parallel with an inductance 164, this inductance serving to cancel the phase shift introduced by the blocking condenser 161 in the inner feedback path at low frequencies. A grid bias resistor 165 and by-pass condenser 166 are shown in the grid circuit of tube 151.

A 24 volt A battery or other suitable direct current source with its positive pole grounded and its negative pole connected to the —A terminal supplies heating current to the filaments of tubes 151, 152 and 153 in series. A 130 volt B battery or other suitable direct current source connected from ground to the +B terminal supplies space current for tubes 151 and 152; and these A and B sources in series supply space current for tube 153 in the general manner disclosed in J. O. Edson and I. G. Wilson Patent 2,191,167, February 20, 1940.

A grid bias resistor 170 by-passed by a condenser 171 supplies grid bias voltage for tube 152. Grid bias voltage for tube 153 is supplied through grid filter resistance 198 by grid biasing battery 172 and any direct current drop in inductance 173 which is in series with battery 172 and the direct current path from the cathode to the grid. This inductance serves, with condenser 175 which is a by-pass from the cathode of tube 153 to ground, as a filter to prevent the alternating plate current of tube 153 from flowing through the A battery. Condenser 176 and resistance 198 filter the biasing voltage for the grid of tube 153; and the condenser 176 provides a by-pass around battery 172, resistance 198 and the A battery in series. A choke 177 and condenser 178 serves as a filter isolating the B battery with respect to alternating current; and condensers 175 and 178 by-pass alternating current between the cathode of tube 153 and the bridge arm 156. Resistance-capacity filters for the direct current plate supply circuits and screen grid biasing circuits of tubes 151 and 152 comprise resistors 180, 181 and 182 and capacities 183, 184, 185 and 186.

Fig. 15A shows the gain frequency characteristics A and B of the amplifier of Fig. 15 for operation without feedback and with feedback respectively, the particular amplifier for which these characteristics were measured being a 12 to 60 kilocycle transmitting amplifier for a twelve-channel cable carrier telephone system.

In this amplifier, tubes 151 and 152 are coupled by an interstage coupling network 190 comprising an air core or magnetic core transformer 187, a parallel-resonant circuit 188, a resistance 189, and stopping condensers or by-pass condensers 191 and 192. The transformer has a primary winding 193 and a secondary winding 194 which may have, for example, a 1:1 turns ratio, these windings being connected at one end by the stopping condenser 191, which connects the plate of tube 151 to the grid of tube 152, and at the other end by the condenser 192. Each of these condensers 191 and 192 may have a capacity of the order of .1 microfarad, for example. Resistor 189 is a grid leak having a resistance, for example, of the order of 60,000 ohms, providing a direct current path for biasing the grid of tube 152 without producing undue shunting effect upon the resonant circuit 188. This resonant circuit 188 is shown as a coil 195 and a capacity 196 which may be the self-capacity of the coil or may include a condenser separate from the coil. The interstage network 190 becomes parallel resonant with the interstage shunt capacity, including the self-capacity of the transformer, at a frequency in the neighborhood of the 60 kilocycle peak P_1 in curve A; and the resonant circuit 188 resonates at a frequency in the neighborhood of the 15 kilocycle peak P_2 in curve A. The transformer 187 renders the coupling impedance of the interstage network 190 high over the upper portion of the utilized frequency range and the resonant circuit 188 renders the impedance of the interstage network 190 high over the lower portion of the utilized frequency range, the two cooperating to maintaian the impedance of the coupling circuit high over the whole of the utilized frequency range, to the end that the amplification of the amplifier without feedback may be high and not unduly variable with frequency over the utilized frequency range.

Tubes 152 and 153 are coupled by an interstage network 190′ structurally and functionally similar to the interstage network 190. The frequency of the peak P_1 in curve A is determined by the resonance frequencies of transformer 187 and the corresponding transformer in network 190′, and these resonance frequencies may differ somewhat. Similarly, the frequency of the peak P_2 in curve B is determined by the resonance frequencies of circuit 188 and the corresponding resonant circuit in network 190′, and these two resonance frequencies may be staggered.

In the feedback path *f* is a deviation equalizer

200 comprising a parallel-connected resistance 201, inductance 202, and a capacity 203, shunted across the path f. This deviation equalizer in the outer feedback path renders the over-all gain-frequency characteristic of the amplifier practically exactly flat over the utilized frequency range. For example, in the 12 kilocycle to 60 kilocycle transmitting amplifier for the twelve-channel cable carrier telephone system, this equalizer rendered the gain flat within about .01 decibel over the transmission band. Where it is desirable to be able to reduce the gain of an amplifier with feedback to as low a value as possible and an equalizer is to be used in the feedback path, it is advantageous to have the $\mu\beta$ loop include the input and output transformers in the general manner of Fig. 15, for example.

In the feedback path f is a network 205, comprising resistances 206, 207 and 208, and the capacity 211, which functions to increase the transmission loss in path f in the transmitted band of the amplifier, thus increasing the amplifier gain over that obtained with no transmission loss in the feedback path. The condenser 211 in parallel with resistance 207 is used to improve the phase shift at high frequencies and to decrease the loss of network 205 at frequencies above the utilized frequency range.

A transformer 215 which may have a turns ratio of 1:1, or any desired ratio, is shown in the feedback path f. The transformer improves the longitudinal balance of the system, and enables the circuits connected to its two windings to be both unsymmetrical (unbalanced-to-ground) or both symmetrical, or to be either one symmetrical and the other unsymmetrical.

If bridge 154 be assumed passively balanced (i. e., balanced even in the absence of feedback), the inner and outer feedback paths will be conjugate to each other. Then, representing by β_1 and β_2 the values of β for the respective loops, the amplification A_F from the grid of the first tube to the plate circuit generator of the last tube is

$$A_F=\frac{\frac{\mu}{1-\mu\beta_1}}{1-\frac{\mu}{1-\mu\beta_1}\beta_2}=\frac{\mu}{1-\mu\beta_1-\mu\beta_2}=\frac{\mu}{1-\mu(\beta_1+\beta_2)}$$

or in general

$$A_F=\frac{\mu}{1-\mu\beta_1-\mu\beta_2-\ .\ .\ .\ \mu\beta_n}\text{ for } n \text{ loops}$$

Thus, as noted above, repetitions are but a special case of parallel feedback, namely, with an added restriction regarding conjugacy. In all cases of repetitions of the feedback process, if $\beta_1, \beta_2\ .\ .\ .\ \beta_n$, are thought of and defined with respect to the single loop which results from viewing the entire arrangement as a simple case of single loop parallel feedback, then the requirement for conjugacy merely becomes an additional requirement relating to the parallel paths, and

$$A_F=\frac{\mu}{1-\mu\beta_1-\mu\beta_2\ .\ .\ .}$$

with no restrictions as to approximation. For example, viewing Fig. 15 in this manner, β_2 includes the insertion loss of the bridge 154 and the insertion loss of the output coil, circuit f, and transmission through input coil to the grid of tube 151, and

$$A_F=\frac{\mu}{1-\mu\beta_1-\mu\beta_2}$$

Unbalance of the bridge 154 as referred to above will not affect the performance of the amplifier as regards the output impedance as long as there is adequate feedback through the output transformer. Feedback will reduce the unbalance of the bridge, thereby producing a definite impedance termination for the transformer. The amplifier output impedance is, however, a function of the turns ratio and the impedance of the network 10, for large amounts of feedback. With the bridge 154 unbalanced the feedback in the amplifier circuit is still an example of repetition of the feedback process. The inside loop makes the outside feedback path conjugate to the outgoing line over the frequency band throughout which there is appreciable negative feedback in the inner loop, and if this exceeds the pass-band of the transformers, the singing difficulty is alleviated.

In the amplifier whose gain-frequency characteristics are shown in Fig. 15A, the entire 45 to 60 decibels of negative feedback was not possible through the hybrid coils. However, 20 decibels was, and therefore the remaining 25 to 40 decibels was transferred to the inside loop. The feedback through the hybrid coils practically removed impedance and transmission requirements on the transformers and caused their modulation requirements to be 20 decibels more lenient.

Fig. 15B shows a circuit of a general type shown in Fig. 15, including at the output of the amplifier a resistance bridge BR and hybrid coil output transformer HC. When the bridge BR is passively balanced (i. e., balanced without feedback), the following conditions obtain:

1. Changes in Z_{β_1} will not affect Z_{Ro}.
2. Changes in $\mu\beta_1$ will not affect Z_{Ro}.
3. Changes in $Z_{L'}$ will not affect β_1.
4. Z_{Ro} is a definite value so that the output hybrid coil can be passively balanced.
5. Z_L and Z_{β_2} can have a ratio such that no current will flow through Z_N.
6. Z_{Ro} and Z_N can have a ratio such that no current will flow through Z_{β_2} when a voltage is applied to Z_0.
7. Z_{Ro} is not a function of $\mu\beta_1$ (see "2" above).
8. A terminating impedance with its power loss may be required if the tube is to work into its optimum impedance and Z_{Ro} is to match $Z_{L'}$.

If the output bridge BR is not passively balanced but is dynamically balanced (i. e., balanced by means of feedback), the following conditions obtain:

1'. Changes in Z_{β_1} will not affect Z_{Ro}.
2'. Changes in $\mu\beta_1$ will not affect Z_{Ro}.
3'. Changes in $Z_{L'}$ will affect β_1.
4'. A suitable impedance Z_{Ro} is presented to the hybrid coil so that it can be passively balanced (i. e., balanced without feedback through β_2).
5'. Z_L and Z_{β_2} can have the right ratio so that no current will flow through Z_N when voltage is applied to $Z_{L'}$.
6'. Z_{Ro} and Z_N can have the right ratio so that no current will flow through Z_{β_2} when a voltage is applied to Z_0.
7'. Having Z_{Ro} independent of changes in $\mu\beta_1$ depends upon having $\mu\beta_1$ sufficiently large (see "2'" above).
8'. The terminating resistance and its transmission loss referred to at "8" in the preceding paragraph will not be required in order to make the tube work into its optimum impedance and at the same time match the coil impedance.

With the particular connection of hybrid coil output transformer indicated in Fig. 15B, there are two ways in which to balance the coil and produce Z_0, as follows:

A. The coil may be passively balanced, i. e., the ratio of Z_{R_0} to Z_N may be given such value that no current flows in Z_{β_2} when a voltage is applied across the output Z_0. Then the impedance Z_0 depends upon the degree of this passive balance of the coil and approaches

$$Z_N\left(1+\frac{n_1}{n_2}\right)$$

B. The coil may be balanced by feedback around through β_2. Then Z_0 becomes independent of the values of Z_{R_0} and Z_{β_2}. As in case "A", Z_0 approaches the value

$$Z_N\left(1+\frac{n_1}{n_2}\right)$$

However, Z_0 now depends not upon the degree or accuracy of the balance between Z_{R_0} and Z_N, but only upon the value of Z_N and upon having a sufficiently large value of $\mu\beta_2$.

If the inner loop feedback is used to produce Z_{R_0} and the outer feedback is also used, there will be two factors tending to produce conjugacy or render Z_0 independent of Z_{β_2} and changes in the outer feedback. The accuracy with which a desired value of Z_{R_0} is obtained will depend upon the design of the unbalanced resistance bridge and the value of $\mu\beta_1$, while $\mu\beta_2$ exerts a further influence to correct Z_0 or stabilize it at the desired value.

When Z_{R_0} does not balance Z_N in Fig. 15B, changes in the value of Z_L affect the transmission from Z_{R_0} to Z_{β_2} whether there is feedback through the β_2 path or not. Hence it may be necessary to consider the effects of changes in output termination in the design of $\mu\beta$ to insure against possibility of a singing condition with some particular termination.

Fig. 16 shows a 12–60 kilocycle amplifier which is a modified form of the amplifier of Fig. 15. Elements corresponding structurally and functionally to elements of Fig. 15 are designated by the same reference characters primed. The interstage circuit coupling tubes 152′ and 153′ comprises, in addition to the network 190″ which corresponds to network 190′ of Fig. 15, resistance 221 in series in the plate circuit of tube 152′ and resistance 222 in series in the grid circuit of tube 153′. These resistances may be, for example, 20,000 ohms each. They serve to reduce the phase shift of the interstage coupling at low frequencies and at high frequencies, as for example at frequencies well below and well above the utilized frequency range, and thus facilitate compliance with Nyquist's rule.

The interstage circuit coupling tubes 151′ and 152′ comprises a plate circuit resistor 231 of, for example, 40,000 ohms resistance, the stopping condenser 191′ and a grid leak resistor 232 of, for example, a half megohm resistance, and further comprises a parallel-resonant circuit 233 connected in series with the plate resistor 231 to form part of the plate coupling impedance. The circuit 233 comprises an inductance 235 and a capacity 236 whose values may be, for example, 60 millihenries and .00017 microfarad respectively, this circuit resonating at a frequency in the neighborhood of 50 kilocycles for example to maintain the gain of the amplifier without feedback high in that frequency region.

Resistances 237, 238, 239 and 240 serve as elements of plate or screen grid potentiometers and filters.

In Fig. 16, battery 251 supplies space current for tube 152′ and battery 252 supplies filament heating current for all of the tubes; and these two batteries in series not only supply space current for the last tube as in Fig. 15 but also supply space current for the first tube. The cathodes of the first and last tubes are then not separated by the impedance of the branched circuit comprising in one branch the coil 173′ and the filament battery and in the other branch the by-pass condenser 175′ (whereas in Fig. 15 the cathodes of the first and last tubes are separated by the impedance of the branched circuit comprising in one branch the coil 173 and the filament battery and in the other branch the by-pass condenser 175). In other words, in Fig. 16 the current fed back from the last tube to the first tube through the inner feedback path does not have to pass through this impedance, so any phase shifts that might be produced in this feedback current if it passed through such impedance are obviated. Further, with the direct current potential thus eliminated in the cathode connection of the inner feedback path, no stopping condenser corresponding to the condenser 161 of Fig. 15 need be used. This elimination of the stopping condenser is advantageous since the condenser is a source of undesirable phase shift.

A parallel connected resistance 261 and capacity 262 in the cathode lead of the tube 152′ produces local negative feedback around tube 152′, serving to introduce in the second stage a corrective phase shift at frequencies between 60 kilocycles and 100 kilocycles for increasing singing margin as regards singing around the feedback loops having the feedback paths 155′ and f'. The resistance 261 and capacity 262 may have values of 200 ohms and .008 microfarad respectively, by way of example. It is seen that the amplifier of Fig. 16 involves multiple feedback, as well as repetition of the feedback process.

In the amplifier system of Fig. 16 the inner feedback around the three tubes, i. e., the feedback through the path 155′ increases singing margin, or facilitates compliance with Nyquist's rule by reducing departure from 180 degrees of the net or total loop phase shift, e. g., of the phase shift for propagation from the first grid once through the complete feedback loop or system (including paths 155′ and f') back to that grid. By way of explanation it may be said that in general in circuits having smooth transmission characteristics over a considerable frequency range which do not have large changes in the rate of change of loss with frequency in such range, the faster the attenuation changes with frequency change at any frequency well within such range, the greater will be the phase shift at that frequency. Thus, when feedback loops are such circuits, the faster the loop gain (i. e., the decibel gain in propagation once around the feedback loop) decreases with frequency-increase at any frequency well within such range, as for example at such a frequency above the highest utilized frequency, the greater is the shunt capacity loop phase shift (i. e., the shunt capacity phase shift in propagation once around the feedback loop) at that frequency. Similarly, the faster the loop gain decreases with frequency-decrease at any frequency well within such range, as for example, at such a frequency below the lowest utilized frequency, the greater is the shunt inductance loop phase shift at that frequency. For frequencies well within a range in which the attenuation change (or gain change) is at a rate of 12 decibels per octave of frequency the minimum phase shift possible is 180 degrees; where the attenuation change (or gain change) is at the rate of 6 decibels per octave the minimum possible phase shift is 90 degrees; where the attenuation change is 10 decibels per octave the minimum possible phase shift is $^{10}/_{12}$ of 180 degrees=150 degrees; etc. In an amplifier circuit such for example as that of Fig. 16, in order to facilitate complying with Nyquist's rule and obtaining sufficient singing margin, it may be desirable to avoid having the reactive phase shift of the propagation from the first grid once through the complete feedback loop or system back to that grid (that is, the phase shift of this loop due to reactance and exclusive of the constant phase shift through the tubes, for example), exceed an angle of say 150 degrees. Then it becomes desirable to prevent the rate of change of the loop gain from exceeding say 10 decibels per octave of frequency, and this is accomplished with the aid of the feedback through path 155'. In other words, without this inner feedback, at some frequency region perhaps well above the utilized frequency range the slope of the curve of loop gain versus frequency would become so great, due for example to the rapidly increasing loss through the input and output transformers, that the loop phase shift would become excessive, with the result that the singing margin would become insufficient, or singing take place, at a frequency in such region. However, feedback through path 155' can prevent this; for, as such frequency region is approached or before it is reached the control of the net or resultant fed-back voltage (amplitude and phase) is transferred from the outer feedback path to the inner feedback path. This may be explained with reference to Figs. 16A and 16B. The amount of the negative feedback through the outer feedback path may be represented by some such curve as the solid line curve 401 of Fig. 16A; and the amount of the negative feedback through the inner feedback path may be represented by some such curve as the solid line curve 402 of that figure. The outer feedback may exceed the inner feedback by, say, some 20 decibels over the utilized frequency range, but fall below the inner feedback at some frequency such, for example, as the frequency indicated by point 403 below the operating frequency range, and also fall below the inner feedback at some frequency such, for example, as that indicated at point 404 above the operating frequency range. In the frequency region of point 403, the magnitude of the net or total or resultant feedback, i. e., the sum of the magnitudes of the inner and outer feedbacks, is given by the dotted curve 405; and in the frequency region of point 404 the total feedback is given by the dotted curve 406. For convenience, the frequency region of point 403, (i. e., the frequency region coextensive with the frequency range of the dotted curve 405), may be called the lower transfer frequency region; the frequency region of point 404, (i. e., the frequency region coextensive with the frequency range of the dotted curve 406), may be called the upper transfer frequency region; and the frequencies at these points 403 and 404 may be called respectively the lower cross-over frequency and the upper cross-over frequency.

Between the upper and lower transfer frequency regions, the outer feedback is the predominant feedback or substantially the total feedback, the inner feedback being relatively very small or insignificant; but below the lower transfer frequency region and above the upper transfer frequency region the outer feedback is negligibly small in comparison to the inner feedback and therefore is inconsequential, the inner feedback being substantially the total feedback. Thus, in passing from the operating frequency range to lower or higher frequencies, dominance or control of the total feedback is transferred from the outer to the inner feedback path (i. e., in effect the total feedback is transferred from the outer to the inner feedback path) as the lower transfer frequency region or the upper transfer frequency region is traversed.

Factors operative in effecting this transfer are, for example, the transmission efficiencies of the input and output transformers and the transmission efficiency of condenser 160'. Fig. 16B may be referred to in this connection. In this figure the β-circuit losses through the outer and the inner feedback paths, (i. e., the decibel losses from the last plate generator to the first grid in propagation through the outer and the inner feedback paths), may be represented by some such curves as 411 and 412, respectively. Regarding curve 411, the input and output transformers contribute largely to the high losses through the outer feedback path that exist both below some such frequency as that indicated by point 413 below the operating frequency range and at frequencies above some such frequency as that indicated by point 414 above the operating frequency range. In the frequency region of the point 413, the net or total or resultant β-circuit loss, i. e., the combined losses of the two feedback paths, is given by the dotted curve 415; and in the frequency region of the point 414 the resultant β-circuit loss is given by the dotted curve 416. The frequency region coextensive with the frequency range of the dotted curve 415 is the lower transfer frequency region referred to above; the region coextensive with the dotted curve 416 is the upper transfer frequency region referred to above; and the frequencies at points 413 and 414 are the lower and upper cross-over frequencies referred to above. Regarding curve 412, the condenser 161' contributes largely to the high losses through the inner feedback path that exist below the lower cross-over frequency. Condenser 160' contributes largely to the decrease of loss through the inner feedback path shown as occurring in the neighborhood of and above the upper transfer frequency region.

The magnitude and phase of the resultant fed-back voltage at the cross-over frequencies will depend upon the relative phases as well as the relative magnitudes of the propagations through the two feedback paths; so, in order to obtain the most advantageous resultant phase shift, the phase shift in each path is controlled and adjusted to have the most advantageous phase shift at the cross-over frequencies. In this connection reference may be had to the dashed curve 421 in Fig. 16B. A curve such as this, for example, might be obtained, instead of the dotted curve 416, if the phase shifts through the two feedback paths differed by 180 degrees at the upper cross-over frequency, since the two equal and opposite transmissions through the two feedback paths would amount to a very high resultant loss for transmission from the last plate generator to the first grid. Such a condition should be avoided, and the transition from curve 401 to curve 404 be smooth, as indicated by curve 406 in Fig. 16A, in order to avoid having an excessive slope in the curve of resultant loop gain versus frequency. For avoiding such condition, in an amplifier such as that of Fig. 16 with two feedback paths, a favorable relation between the phase shifts through the feedback paths at the cross-over frequency is a difference of approximately 120 degrees. That is, with the two feedback paths, at the cross-over frequency it is advantageous to have the phase shift of one of the feedback paths exceed that of the other by approximately 120 degrees.

A phase control network **270** in the feedback path f' comprises series arms consisting of resistances **271** and **272** shunted by condensers **273** and **274**, and comprises shunt resistance arms **275** and **276**. The values of these impedances may be for example as follows: resistances **271** and **272**, 2660 ohms each, capacities **273** and **274**, .00045 microfarad each; resistance **275**, 10 ohms; and resistance **276**, 100 ohms. This network **270** functions in a manner similar to that of network **205** of Fig. 15 and enters into the phase control of β_2 at the upper cross-over frequency.

This amplifier is designed for use not only as a transmitting amplifier but also for other uses, requiring a different gain; and to increase the gain, say 5 decibels, a pad **280** having an insertion loss of 5 decibels can be inserted in the feedback path f', the feedback resistor **163'** being at the same time changed, say from 80 ohms to 44.5 ohms. The function of inserting a different resistance at **163** when pad **280** is increased is to maintain the feedback through the transformers constant. Thus in this amplifier maintaining the feedback constant through the coils tends to maintain the impedance at the input and output constant.

The hybrid coils **5'** and **7'** in Fig. 16 function in the same general manner as the hybrid coils **5** and **7** in Fig. 15. The net **9'** of the hybrid coil **5'** is shown as a resistance **285** in series with parallel connected resistance **286** and capacity **287**. An inductance **291** and a capacity **292** are shown in series with the incoming circuit. By way of example, the resistance **285** may be 130 ohms, the resistance **286** may be 50 ohms, the capacity **287** may be .3 microfarad, the inductance **291** may be .47 millihenry, and the capacity **292** may be .06 microfarad. These values render the input impedance of the amplifier suitable for connection to a circuit whose impedance is a resistance of 600 ohms. Inductances **293** and **294** are shown connected in series with the outgoing line. By way of example, these inductances may each be approximately 50 microhenries. With these values they render the output impedance of the amplifier suitable for connection to a 19 gauge cable.

In appropriately modifying ϕ, the loop phase shift of a feedback amplifier or system, with a view to complying with Nyquist's rule so as to avoid singing, the design of the system is often facilitated by curves such for example as those of Fig. 17 enabling the change θ_{CF} that feedback around an amplifier through a feedback path produces in the phase shift through the amplifier to be found graphically from the loop gain $G_{|\mu\beta|}$ and the loop phase shift ϕ of the feedback loop through the amplifier and the feedback path. The curves of Fig. 17 are obtained as indicated below from the following equation for change in phase due to feedback:

$$\theta_{CF}=\tan^{-1}\frac{|\mu\beta|\sin\phi}{1-|\mu\beta|\cos\phi}$$

Since $G_{|\mu\beta|}=20\log_{10}|\mu\beta|$

$$|\mu\beta|=10^{\frac{G_{|\mu\beta|}}{20}}$$

and, substituting this value of $|\mu\beta|$ in the formula just given for θ_{CF} yields

$$\theta_{CF}=\tan^{-1}\frac{10^{\frac{G_{\mu\beta}}{20}}\sin\phi}{1-10^{\frac{G_{|\mu\beta|}}{20}}\cos\phi}$$

The curves of Fig. 17 are the loci of this latter equation for fixed values of the parameter $G_{|\mu\beta|}$. Positive values of the parameter represent a gain around amplifier and feedback path, negative values a loss. The points of intersection of either of the light dotted lines with the family of contour lines are points of either maxima or minima of the contour lines (and satisfy boundary B of my above-mentioned article).

Fig. 17A gives a similar family of parametric curves, but between ϕ and $G_{|\mu\beta|}$, and with θ_{CF} as the parameter. For positive values of ϕ the sign of θ_{CF} is also positive.

Fig. 18 shows a negative feedback amplifier which is a modification of the amplifier of Fig. 1 suitable, for example, for a three-channel repeater for an open wire carrier telephone system of the general type of the present commercial type C carrier telephone system described by H. A. Affel, C. S. Demarest and C. W. Green in the Bell System Technical Journal, July 1928, pages 564 to 629.

A 24-volt A battery or other suitable direct current source with its positive pole grounded and its negative pole connected to the terminal designated −24V. supplies heating current to the filaments of tubes **1** and **2**; and this source and in series therewith a 130-volt B battery or other suitable direct current source connected from ground to the terminal marked +124V., supply space current for the tubes in the general manner described above in connection with Fig. 15 and Fig. 16. A choke coil **301** and a by-pass condenser **302** serve as a filter for the plate and screen voltage supply circuits of the tubes; and a resistance **303** and by-pass condenser **304** serve as a potentiometer and filter for the voltage supply circuit of tube **1**. Grid biasing voltage for tube **1** is supplied from resistor **305** by-passed by condenser **306**; and grid biasing voltage for tube **2** is supplied by resistor **307** by-passed by condenser **308**.

The interstage coupling circuit comprises a transformer **310**, resistors **311** and **312** and a by-pass condenser or direct current stopping condenser **315** whose capacity may be for example of the order of .1 microfarad. The resistor **311** is in series with the primary winding of transformer **310** in the plate circuit of tube **1** and the resistor **312** is in series with the secondary winding of the transformer **310** in the grid circuit of tube **2**. The transformer may have a turns ratio of 1:1, for example, and the resistances **311** and **312** may be, for instance, 10,000 ohms and 100,000 ohms, respectively. Without the resistances, the transformer would appear at low frequencies as a shunt inductance introducing −90 degree phase shift and at high frequencies as a shunt capacity introducing +90 phase shift. The resistances render the phase shift of the interstage coupling network substantially zero at low frequencies and reduce the phase shift to some 40 degrees or 50 degrees at high frequencies, parasitic shunt capacities tending to limit the amount that these resistances reduce the high frequency phase shift.

Thus, the resistances reduce the phase shift some 50 degrees or 40 degrees at high frequencies at which, without such reduction of phase shift in the interstage coupling circuit, the system might sing; and the resistances also increase the singing margin at low frequencies.

Transformer **320** in the feedback path *f* may be, for example, a transformer of unity ratio or other suitable ratio transformer functioning similarly to the transformers **215** and **215'** in Figs. 15 and 16.

A β circuit network **3** serves to reduce singing tendency or give increased margin against singing at high and low frequencies (frequencies above and below the utilized frequency range) by reducing the rate at which the loop gain decreases with increase or decrease of frequency and consequently reducing the loop phase shift. As indicated above, the faster the loop gain decreases with frequency change at any given frequency, the greater is the loop phase shift at that frequency. The input transformer, the interstage coupling circuit and the output transformer each decreases its transmission at about 6 decibels per octave of frequency at frequencies above and below the utilized frequency band at which the loop gain without network **3** is greater than zero. To reduce the rate of change of loop gain from this objectionably high value of 18 decibels per octave to say 10 decibels per octave, network **3** is inserted. Over the utilized frequency band it is substantially a constant resistance pad. At high frequencies and at low frequencies the attenuation of the network decreases at such a rate (about 8 decibels per octave) that the loop attenuation decreases at only about 10 decibels per octave. This network **3** is shown as comprising resistances **321** to **328** and capacities **329** to **331**. The values of these elements may be, for example, as follows: resistances **321** and **322**, 500 ohms each; resistances **323** and **324**, 600 ohms each; resistance **325**, 550 ohms; resistance **326**, 130 ohms; resistances **327** and **328**, 50 ohms each; capacity **329**, 1 microfarad, and capacities **330** and **331**, .0005 microfarad each.

As indicated above, it is often convenient to distinguish between single loop electrical feedback amplifiers or systems on the basis of the general type of the connection of the feedback path to the amplifier and load circuits and similarly, the type of the connection of the feedback path to the amplifier and the sending circuit. The type of feedback connection at the amplifier input may be, for example, (A) shunt, (B) series, (C) balanced bridge, (D) unbalanced bridge, (E) balanced hybrid, (F) unbalanced hybrid or (G) grid. The terms balanced and unbalanced here refer to whether the connection is biconjugate without feedback. (With feedback the unbalanced bridge and hybrid connections are automatically conjugate in certain respects as indicated above.) Grid feedback refers to feeding back to an auxiliary grid, which can be separate from the sending circuit, for example, as in the British patent of S. T. Meyers and I. G. Wilson, No. 475,400. At the amplifier output the feedback connection may be, for example, any of types (A) to (F). (Ordinarily, transferring energy from an auxiliary plate separate from the output plate to the input is not considered as bringing back to the input a portion of the actual output since the current in the auxiliary plate circuit does not fulfill the requirement that it be a true copy, or true amplified or attenuated copy, of the load current.) Any of the input connections can be used with any of the output connections; and input and output connections may be coalesced, for example, as in Figs. 8, 9 and 10. In accordance with the invention, a hybrid coil type of feedback connection at the amplifier input or output can be used with a variety of general types of feedback connections at the amplifier output or input; and the specific types of hybrid coil feedback connections described above are merely examples of a large variety of species of a general type of biconjugate circuit using transformers, any of which can be used as feedback connections in accordance with the invention. The multiplicity of the species of such transformer networks available is brought out for example in the paper on "Maximum output networks for telephone substation and repeater circuits," by G. A. Campbell and R. M. Foster, Transactions of the American Institute of Electrical Engineers, vol. 39, 1920.

Fig. 19 shows by way of example a feedback circuit with a modified form of the hybrid coil type of feedback connections of Fig. 1. This circuit (which is dominated by E. H. Perkins application Serial No. 114,393, entitled Electric wave amplifying systems, filed of even date herewith), may be for instance a negative feedback amplifier circuit of the general type described in connection with Fig. 1, with hybrid-hybrid feedback around an amplifier **350** through feedback path *f*. The sending circuit or incoming circuit **6** connects to the bridge points of input hybrid coil H_1 whose balancing network is shown at **351**. The receiving circuit or outgoing circuit **8** is connected across the bridge points of output hybrid coil H_2 whose balancing network is shown at **352**.

Fig. 19 is only one of a number of biconjugate network arrangements all of which possess the same fundamental advantages. This particular example would be useful for connecting to unbalanced circuits as, for example, coaxial pipes and possess the advantages of precision matched impedances even in the face of high and non-uniform voltage step-up transformers. It has the further advantage of a feedback path *f* that is electrically insulated from the remainder of the circuit and hence there is the greatest possible flexibility from the standpoint of impedance value or type of network that might be inserted in the feedback path to accurately control the transmission characteristics of the amplifier. This circuit has all of the advantages of feedback through the input and output transformers.

What is claimed is:

1. A wave translating system comprising an amplifying path, a second path for connection thereto, and means for feeding back in said amplifying path a portion of its output waves in gain-reducing phase, comprising a feedback path and a bridge transformer network connecting said amplifying path to said feedback path and to said second path, said network comprising a transformer having three windings, with two of said windings in two of said paths, respectively.

2. An amplifier for connection to a wave transmission path, said amplifier having an amplifying path, a feedback path, a fourth path and means for stabilizing the impedance presented to said wave transmission path by said amplifier against variations in the amount of feedback, comprising a network for interconnecting said four paths, said network including a transformer having three windings in three of said paths, respectively.

3. A wave amplifying circuit and a connecting

circuit therefor, said amplifying circuit having means for producing negative feedback in amount dependent upon transmission level in the amplifying circuit, and means for preventing variations that the level changes produce in the amount of feedback from varying the impedance of the amplifier facing said connecting circuit.

4. An amplifier having an amplifying path and a feedback path therefor for feeding back waves in gain-reducing phase, a balancing circuit for said amplifying path, a wave transmission circuit, and a transformer having a two-part winding, one part in said feedback path and the other part in said transmission circuit, and a second winding connected in said amplifying path, said balancing circuit being bridged across the division points of the two-part winding.

5. An amplifier comprising an amplifying path and a feedback path therefor feeding waves back in gain-reducing phase, two circuits, and means to associate the amplifier with said two circuits, comprising a transformer having a two-part winding, one part in one of said paths and the other part in one of said two circuits, and a second winding in the other of said paths, the other of said two circuits being bridged across the division points of the divided winding.

6. A wave translating system comprising an amplifying path and a feedback path therefor, a circuit for association with said amplifying path, an impedance, and a hybrid coil with effectively three transformer windings connected one between said circuit and said impedance, a second between said feedback path and said impedance, and a third in said amplifying path.

7. A wave translating system comprising a wave amplifying path and a feedback path therefor for feeding waves back in gain-reducing phase, a wave transmission circuit for association with said amplifying path, an impedance device, and a bridge transformer with four pairs of terminals conjugate in pairs and connected one pair to the amplifying path, a conjugate pair to the impedance device, a third pair to the feedback path, and the fourth pair to the transmission circuit.

8. A wave translating system comprising a wave amplifying path and a feedback path therefor feeding waves back in gain-reducing phase, a wave transmission circuit for association with said amplifying path, and an interconnecting circuit for associating said amplifying path with said wave transmission circuit and said feedback path, said interconnecting circuit comprising a bridge transformer and a balancing network for balancing the impedance of said amplifying path.

9. A wave translating system comprising an amplifying path and a feedback path therefor, a balancing circuit for said amplifying path, a wave transmission circuit for association with said amplifying path, a circuit connecting said transmission circuit and said feedback path in series with each other, and a transformer having a divided winding in series in said circuit and an inductively related winding in said amplifying path, said balancing circuit terminating at the division points of said divided winding.

10. An automatic volume control circuit comprising an amplifier with a feedback path for producing negative feedback of waves to be amplified having in series therein an element of silver sulphide for limiting the amplifier output volume by decreasing in resistance as a result of self-heating caused by increase of current from the amplifier output through the silver sulphide in response to increase in the output volume and thereby increasing the negative feedback.

11. A wave amplifying system producing signal waves and non-signal waves originating in the system, said amplifying system having a feedback path for feeding back said non-signal waves in such phase as to oppose their production, and a self-heating thermo-responsive transmission control element of large negative temperature coefficient of resistance in said path for varying the transmission efficiency of said path in response to change in temperature of said element directly due to current in said element.

12. A wave translating system comprising an amplifying path and a feedback path therefor including a self-heating thermo-responsive impedance element of high temperature coefficient of resistance for controlling transmission through said feedback path in response to change in temperature of said element directly due to current flow in said element, a circuit for connection to said amplifying path, and means comprising a bridge-transformer network for connecting said amplifying path to said circuit and to said feedback path with said circuit and said path in conjugate relation to each other.

13. The method of obtaining a prescribed value of impedance across a given pair of terminals of a circuit including a negative feedback amplifier having a bridge transformer network with four pairs of terminals connected one pair to the amplifying path, a second pair to the feedback path, a third pair to said given terminals, and the fourth pair to an impedance device, which comprises giving the impedance device an impedance value that bears a predetermined fixed ratio to said prescribed impedance value.

14. The method of obtaining an impedance having a resistance component free from resistance noise across a given pair of terminals of a circuit including a negative feedback amplifier having a bridge transformer network with four pairs of terminals connected one pair to the amplifying path, a second pair to the feedback path, a third pair to said given terminals, and the fourth pair to an impedance device, which comprises including a resistance in the impedance device.

15. A wave amplifying system comprising an amplifying path with a gain-reducing feedback path therefor, an impedance for association with said amplifying path, and means for interconnecting said paths and said impedance, comprising a three-winding transformer and a second impedance, said transformer having one winding connecting said impedances, a second winding connecting said feedback path and said second impedance and the third winding connected in said amplifying path, and the ratio of the values of said first and second impedances being substantially unity plus the ratio of the turns of said first and second windings.

16. A feedback system comprising a feedback loop having a feedback path producing negative feedback around the loop over a useful frequency range, the reactive loop phase shift having objectionably high value at frequencies outside of said range, and transmission control means in the feedback path having the slope of its frequency-characteristic of transmission efficiency greater at said frequencies of objectionably high phase shift than in said useful range and opposite in sign at said frequencies to the slope of the frequency-characteristic of transmission efficiency of the remainder of the loop, for reducing the

slope of the frequency-characteristic of the loop gain at said frequencies to reduce the reactive loop phase shift at said frequencies and thereby reduce tendency of the system to sing around said loop.

17. A feedback system comprising an amplifying path, an inner feedback path forming with said amplifying path an inner loop for producing negative feedback, an outer feedback path forming with said amplifying path an outer feedback loop for producing negative feedback, a circuit and an impedance, said inner loop comprising a bridge network for rendering said paths conjugate, and said outer feedback loop comprising a three-winding transformer, said transformer having one winding connecting said circuit and said impedance, a second winding connecting said outer feedback path and said impedance, and the third winding interconnecting the transformer and the bridge network.

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