

March 30, 1943.

H. W. BODE

2,315,040

ELECTRIC WAVE AMPLIFICATION

Filed Dec. 28, 1940

3 Sheets-Sheet 1

FIG. 1

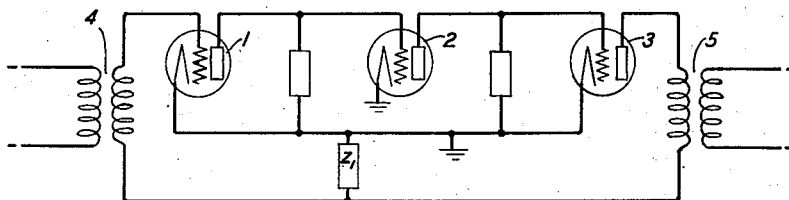


FIG. 2

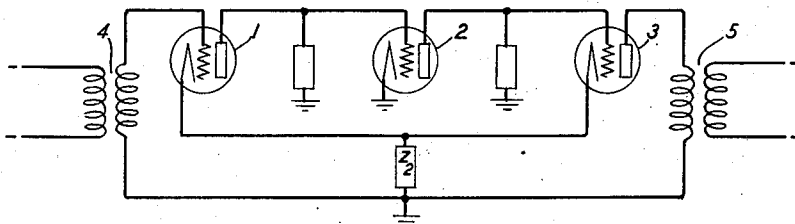
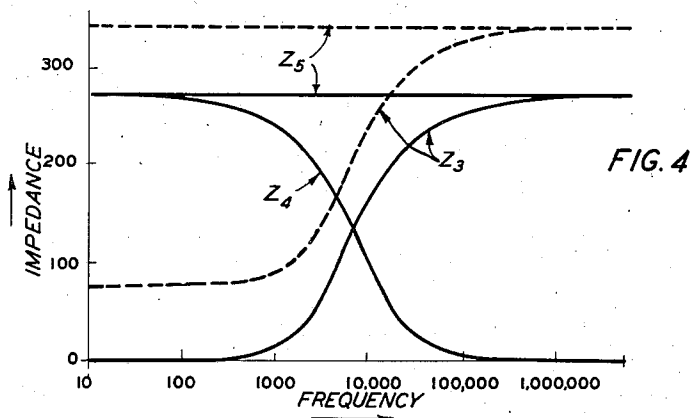
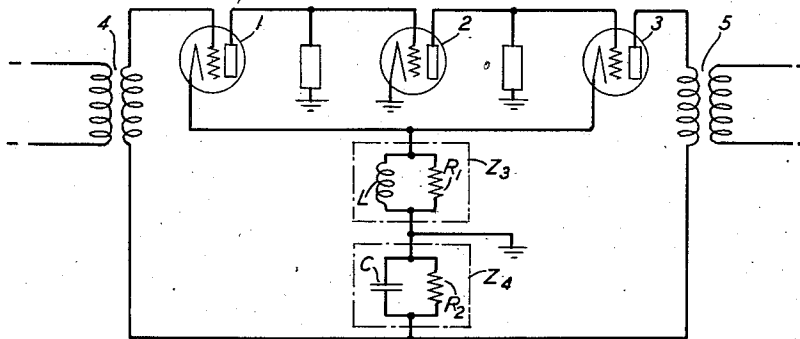


FIG. 3



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FIG. 5

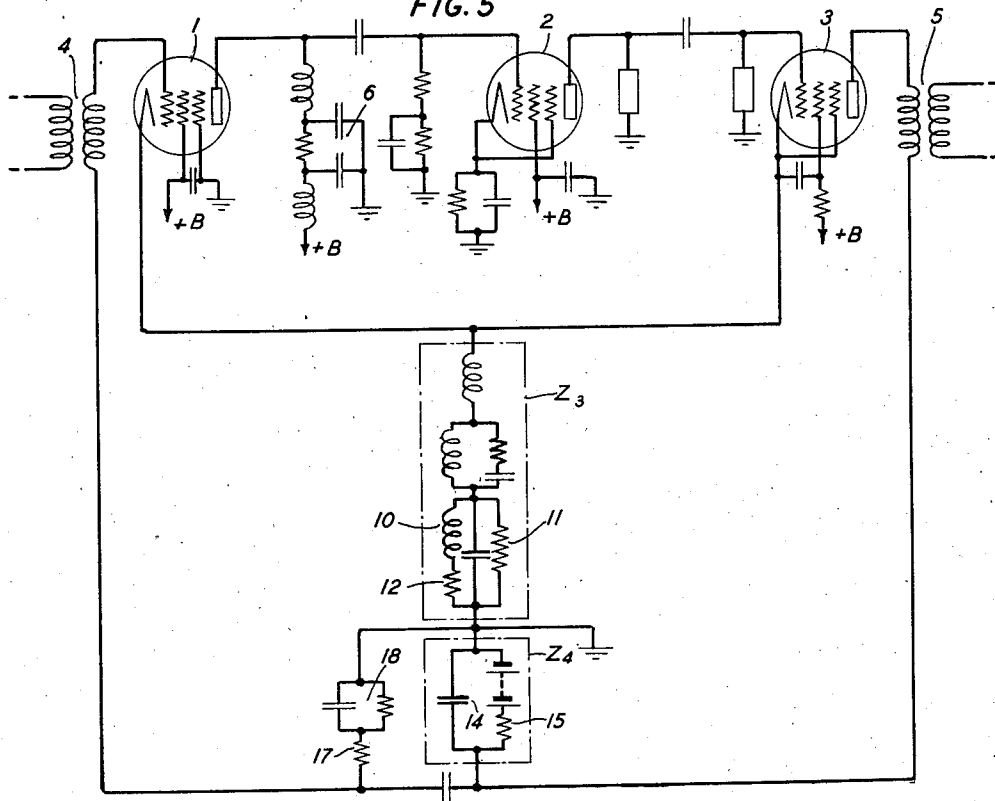


FIG. 6

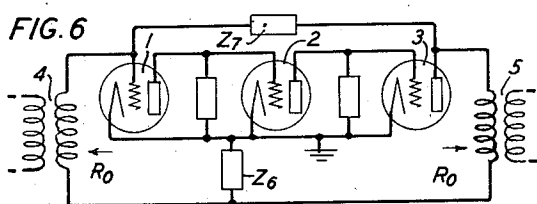


FIG. 7

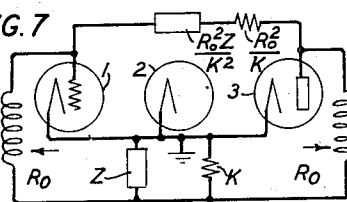
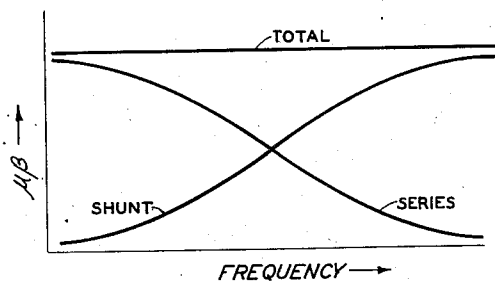


FIG. 8



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FIG. 9

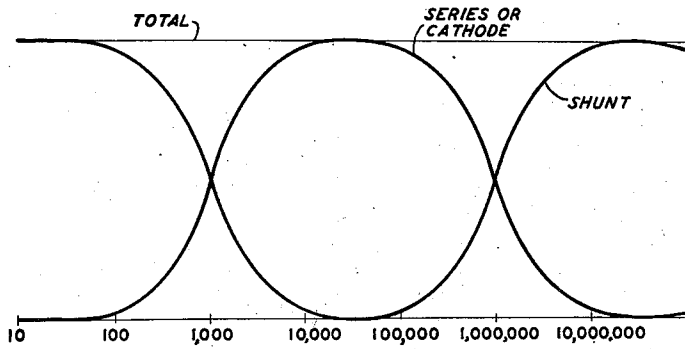


FIG. 10

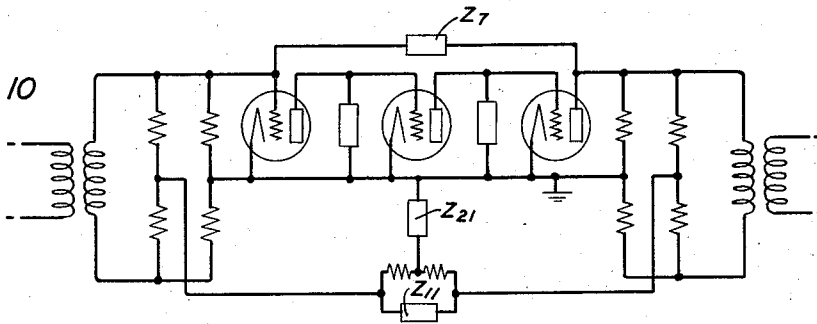


FIG. 11

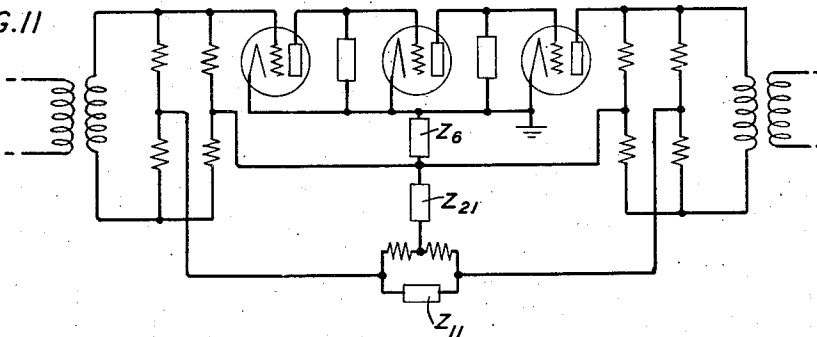
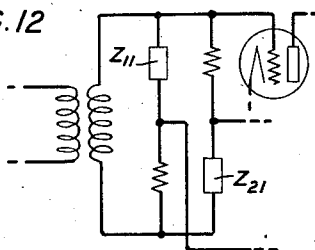


FIG. 12



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2,315,040

ELECTRIC WAVE AMPLIFICATION

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Application December 28, 1940, Serial No. 372,073

14 Claims. (Cl. 179—171)

This invention relates to electric wave amplification and more particularly to stabilized feedback amplifiers adapted for the amplification of signals occupying a wide frequency range.

A principal object of the invention is to improve the performance and simplify the construction of stabilized negative feedback amplifiers.

Another object is to increase the amount of gain stabilizing feedback that may be obtained in an amplifier without sacrifice of other performance characteristics.

Another object is to reduce the effects in a feedback amplifier of spurious or parasitic impedances such as stray capacitances associated with amplifier input and output transformers or like coupling devices.

A further object is to facilitate control of the terminal impedances of an amplifier as modified by feedback.

Still other objects are to combine in one amplifier the respective principal advantages of different types of feedback, to effect a smooth transition over the frequency range from one type of feedback to another and to maintain the vectorial product $\mu\beta$ constant over the signaling frequency range.

The foregoing objects and various others that will appear hereinafter are achieved in stabilized negative feedback amplifiers employing concurrent feedbacks of respectively different types, such as series, shunt, cathode and bridge types. The nature of the present invention and its various features, objects and advantages will appear more fully from a consideration of the following description of the typical embodiments illustrated in the accompanying drawings.

In the drawings:

Figs. 1 and 2 are schematic showings of feedback amplifiers of the normal series and cathode types, respectively;

Fig. 3 illustrates an embodiment of the invention combining series and cathode feedback;

Fig. 4 is a curve diagram to which reference will be made in the description of Fig. 3;

Fig. 5 is a more detailed showing of an amplifier in accordance with Fig. 3;

Figs. 6 and 7 show schematically an amplifier combining series and shunt feedback;

Figs. 8 and 9 comprise curve diagrams to which reference will be made in the description of Figs. 6 and 7;

Fig. 10 shows schematically an amplifier combining bridge and shunt feedback;

Fig. 11 illustrates schematically an amplifier combining bridge and series feedback; and

Fig. 12 illustrates a modification of Figs. 10 and 11.

Inasmuch as preferred embodiments of the invention chosen for purposes of exposition herein involve cathode feedback, or normal series feedback or both, it may be helpful to consider briefly the nature and characteristics of these two types of feedback. Fig. 1 illustrates schematically an amplifier having feedback of the normal series type comprising three amplifying stages 1, 2, 3 with input and output transformers 4 and 5, respectively. Although there may optionally be provided feedback local to the several amplifying stages, the principal feedback in this type of amplifier is of the voltage-voltage type and the forward or mu circuit is maintained at substantially ground potential, at least in the input and output stages. Thus the anode circuit for the third stage is completed to the grounded cathode thereof through a coupling impedance Z_1 , and the resultant voltage drop through the latter is applied in series in the input circuit of the first stage.

A desirable feature of the normal series feedback amplifier as illustrated in Fig. 1 is that the cathodes are maintained at least approximately at ground potential and the feedback is over-all feedback from the last stage to the first. A disadvantage is that the input and output transformers may have associated with them a very large stray capacitance to ground which therefore effectively appears in shunt across the coupling impedance Z_1 . The shunting capacitances tend to be especially large if the amplifier is adapted for television signals (which may range in frequency from 45 cycles per second to several megacycles, for example), for the transformers may take the form of the rather bulky coupling networks disclosed for example in my application for Letters Patent, Serial No. 352,422, filed August 13, 1940. Although the shunting capacitances may have high reactance at frequencies within the signal range, nevertheless they impose a limitation on the maximum achievable performance of the amplifier in view of their effect on the high frequency asymptotic characteristic, as disclosed fully in my U. S. Patent 2,123,178, July 12, 1938.

Negative feedback of the cathode type is exemplified in Fig. 2 which may be considered as showing an amplifier essentially the same as that of Fig. 1 excepting for the change in the principal feedback circuit. The latter in this case is also of the voltage-voltage type with the important difference that the respective low

potential ends of the secondary winding of input transformer 4 and the primary winding of output transformer 5 are maintained at ground potential. In this case, as in Fig. 1, last stage anode current flowing through the coupling impedance Z_2 produces a voltage drop across the latter which is applied in series with the first stage input circuit. The cathodes of the first and third stages, it will be noted, are maintained off ground potential by the signal voltage drop in the coupling impedance Z_2 . The virtues of the cathode type of feedback appear largely in the asymptotic frequency range where the fact that the transformer capacitances to ground do not appear across the feedback circuit removes the restriction present in the normal series feedback amplifier of Fig. 1.

Another principal advantage of the cathode type of feedback results from the redistribution of input and output tube capacities which it permits. Of the total capacity associated with the control grid in the input stage a portion goes to the cathode of the input tube and the remainder to ground. Similarly, the total capacity associated with the anode of the output tube is distributed between the output cathode and ground. With the normal series type of feedback connection, the cathode and ground capacities are indistinguishable, since the cathodes are at ground potential. In the cathode feedback circuit, on the other hand, the capacity between the input grid and ground and the capacity between the output anode and ground are removed from the grid-cathode and anode-cathode paths and fall instead across the high impedance or facing windings of the input and output transformers. Now it is shown in my pending application for patent, Serial No. 297,069, filed September 29, 1939 (Patent No. 2,242,878, May 20, 1941), that in a series or cathode feedback system an appropriate ratio must be maintained between the grid-cathode or anode-cathode capacities and the capacities across the high impedance windings of the corresponding input or output transformer if a satisfactory asymptotic feedback characteristic is to be obtained. In the normal series feedback connection where all of the tube capacities appear in the grid-cathode or anode-cathode paths, this generally means that large supplementary capacities must be added across the transformer. The redistribution of tube capacities in the cathode type feedback circuit on the other hand permits the desired ratio to be met automatically with the addition of at most very small supplementary capacities. The reduction of the total capacity associated with the high impedance windings of the transformers permits a corresponding increase in the impedance ratio which may be obtained with the transformers. In practical cases the transformer impedance ratios obtainable with the cathode type feedback connection may be as much as two or three times those permissible with a normal series feedback structure and there is a directly proportional improvement both in the level of the input signal delivered to the amplifier and in the level of the output power delivered by the amplifier.

The principal difficulties encountered in the cathode type feedback circuit arise in supplying power to the tubes. The power supply circuits themselves are at ground potential. Since the cathodes of the first and third tubes are off ground potential this means that the screen grid and anode currents of these tubes must return

to cathode by flowing through the beta circuit. If the beta circuit includes a direct current resistance the flow of these currents produces a direct current voltage drop across the beta circuit which appears as a grid biasing voltage on the first and third tubes. Inasmuch as the beta circuit impedance Z_1 should ordinarily appear as a pure resistance over the low frequency range, the anode current associated with the output stage tends to produce in the first stage an excessive grid biasing voltage drop across the impedance Z_1 . To reduce the effective resistance of the coupling impedance Z_1 for direct current but not for higher frequencies one could conceivably shunt the coupling impedance with a large inductance or choke coil of low resistance. However, the magnitude of the inductance theoretically required for this purpose, in view of the fact that as suggested the lower end of the signal range may be 45 cycles, produces further complications and is not considered desirable. In a television amplifier especially the beta circuit phase angle is of such importance that the required inductance is very great. The inductance, moreover, must have negligible distributed capacity to avoid difficulties at high frequencies.

Still another problem is presented in the cathode type of feedback in view of the fact that the screen grid currents include, in addition to a direct current component, an alternating component of signal frequency plus distortion products. In typical tubes, the signal and distortion components in the screen grid currents are roughly similar to those in the corresponding anode currents but their level is about 14 decibels lower. If these screen grid current components produce corresponding voltage drops in the feedback impedance Z_1 the practical benefits of negative feedback with respect to gain stabilization and distortion reduction can be realized only to the extent of approximately 14 decibels. The usual remedy of biasing the screen grid through a resistance-condenser filter fails when the lower edge of the signal band is so low as to require a bulky filter condenser having large distributed capacitance to ground.

In the embodiment of the invention illustrated in Fig. 3, normal series feedback and cathode feedback are intimately and effectively combined. The cathodes of the first and third stages are joined together and connected through a common impedance network Z_3 to ground. The low potential ends of the secondary and primary windings respectively of the transformers 4 and 5 are joined together and connected to ground through a common impedance network Z_4 . The two impedance networks Z_3 and Z_4 may be considered as constituting together a single impedance network Z_5 with an internal connection to ground. Preferably network Z_3 is so proportioned that it has substantially constant resistance over one frequency range and is effective in providing negative feedback in that range, and low or substantially zero impedance in a lower frequency range. Conversely, impedance network Z_4 is preferably so proportioned as to have constant resistance over a low frequency range and low or substantially zero impedance over a higher frequency range.

For high frequencies, such as those of interest with respect to the asymptotic characteristic of the amplifier, impedance Z_4 may be considered

zero and the transformers are therefore effectively maintained at ground potential. It will be seen therefore that in the high frequency range the amplifier is essentially of the cathode feedback type and all of the advantages incident to that type of feedback are retained in the critical high frequency range. At low frequencies such that Z_3 may be disregarded, it will be seen that the several cathodes are reduced to substantially ground potential and the amplifier becomes essentially of the normal series type with its attendant advantages at low frequency in respect to the supply of power and biasing voltages. It may be noted that in so far as the Fig. 3 amplifier is of the series feedback type, the stray capacitances to ground associated with the anode and grid electrodes of the amplifying tubes do not enter into or affect the external gain of the amplifier inasmuch as they are in the mu circuit. It may be noted too that there are no high frequency asymptotic requirements on impedance Z_4 . Still another significant point that may be mentioned is that the transformer capacitances to ground and the interelectrode capacitances to the several cathodes are separated, and therefore the possibility is opened for separate treatment and compensation for the capacitances.

Examining now the internal construction of impedance networks Z_3 and Z_4 , an arrangement is shown whereby a smooth transition from series to cathode feedback is effected. More particularly, impedance network Z_3 is represented as comprising an inductance L and a shunting resistance R_1 , and Z_4 is represented as comprising a capacitance C shunted by a resistor R_2 . In actual practice the two impedance networks may be much more complex than here indicated, hence the simple networks shown should be considered as representing the equivalent circuits of the actual networks. The equivalent circuits then are preferably so proportioned that L/C is equal to $R_1 R_2$. If more particularly R_1 and R_2 are equal, then it can be shown that the impedance Z_5 of the combination is equivalent to a resistance R at all frequencies. In other words the feedback coupling impedance is independent of frequency despite the transition from one type of feedback to another. The solid lines in Fig. 4 show in curve diagram form the manner of variation of the three impedances with respect to frequency. Since Fig. 3 has been indicated as showing simplified equivalent circuits for Z_3 and Z_4 , it may be well to point out that the equivalency need not hold at frequencies far removed from the transition frequency range. That is, at very low frequencies it is not essential that inductance L retain its character as such or that R_1 remain a resistance or of unchanged magnitude so long as the over-all impedance Z_3 is negligible. Similarly for Z_4 , it is not essential that its components maintain the character indicated at frequencies where Z_4 is substantially zero. The dotted lines in Fig. 4 apply to a case where a certain amount of cathode feedback is retained at low frequencies, as for example where a resistance effective at low frequencies is interposed in series with the inductance L in the Z_3 structure of Fig. 3. Again it will be observed that the transition is smooth and that the total feedback is constant throughout the signal frequency range.

With regard to the cross-over frequency, that is, the frequency represented by the intersection

of curves Z_3 and Z_4 in Fig. 4, it is desirable in the case of the television amplifier assumed that it be a comparatively low frequency of the order of a few kilocycles. On the one hand the cross-over frequency should be high enough that alternating current components in the screen grid lead are effectively by-passed to the cathode. An upper limit on the cross-over frequency is fixed by the transformer capacitances to ground, for these establish a minimum value for the capacitance element C in Fig. 3, understanding that these capacitances are a part of the capacitance C there represented. Whereas this last consideration might fix an upper limit of a few hundred kilocycles for the cross-over frequency a lower frequency may be found desirable, particularly if a non-uniform gain characteristic, such as one increasing with frequency from say 50 kilocycles upward, is to be incorporated in the amplifier.

In this discussion it has been supposed that the total beta circuit impedance, and therefore also the total amplifier gain, should be flat with frequency. The provision of a variable gain characteristic, however, is also feasible. Thus, a shaping impedance network dominant at high frequencies may be interposed in series with one of the terminal leads of Z_3 and another dominant at low frequencies may be similarly associated with Z_4 , or a single network may be shunted across the series combination of Z_3 and Z_4 . It should be understood too that the provision of a beta circuit impedance that is not flat with frequency is not inconsistent with $\mu\beta$ being constant throughout the signaling frequency range, for as β varies, μ may be made to vary in inverse relation by shaping the interstage networks in well-known ways.

Fig. 5 shows in greater detail an amplifier substantially conforming with Fig. 3 and specifically adapted for amplification of television signals occupying the frequency range from 45 cycles to 3 megacycles, and further adapted for compensating the non-uniform attenuation of a transmission line repeater section. The amplifier tubes 1, 2 and 3 are pentodes and are connected in tandem by suitable coupling impedances. In the first stage the screen grid is provided with a condenser by-pass to ground and the suppressor grid is directly grounded. In the second stage a resistance and shunting condenser are provided in the cathode lead to provide control grid bias and local feedback. The screen grid is provided with a condenser by-pass to ground and the suppressor grid is tied directly to the cathode as it is in tube 3 also. In the third stage the screen grid is provided with a condenser by-pass to the cathode and it is connected to the biasing battery through a resistance.

The beta circuit network Z_3 comprises an inductance 10 of 10 millihenries and a resistor 12 of 70 ohms connected in series across a resistor 11 of 333 ohms. The value of resistor 12 is such as to provide an IR drop suitable for biasing the first and third stage control grids and it incidentally provides a certain amount of cathode feedback even at low frequencies, as described with reference to the dotted line curves of Fig. 4. If desired, resistor 12 may be eliminated and its grid biasing function performed by cathode circuit resistors individual to the first and third stages, in which case Fig. 3 is accurately equivalent. In the beta circuit network Z_4 the capacitance C is obtained by a condenser 14 of 0.086 microfarad where the parallel disposed

transformer-to-ground capacitance is 0.0014 microfarad. Condenser 14 is shunted by a resistor 15 in series with the anode and screen grid voltage source, the latter being represented by the battery symbol and assumed to have negligible internal impedance. Where resistor 12 is not employed resistor 15 may be 333 ohms, the same value as resistor 11. If a resistor 12 is employed it is compensated by shunting resistor 15 with the inverse of resistor 12, which in this case is 1560 ohms, or the resistor 15 may be used alone and assigned an equivalent value, namely, 275 ohms.

A grid leak connection for the first stage grid is provided by a resistor 17 of 0.16 megohm, for example, and if desired a resistance-condenser combination 18 may be interposed in series therewith as shown and the several values so proportioned as additionally to provide low frequency $\mu\beta$ control.

The circuit elements of networks Z_3 and Z_4 that are shown in Fig. 5 but not described above are employed for shaping the beta circuit transmission characteristic at frequencies well above the cross-over frequency to compensate for the rising attenuation frequency characteristic of a preceding repeater section of a coaxial conductor transmission system. These elements have no substantial effect at frequencies in the vicinity of the cross-over region but only at frequencies in the range above say 50 kilocycles.

The cross-over frequency in the Fig. 5 circuit as above described is approximately 5 kilocycles or about one octave below the mean frequency of the signal band. This frequency is high enough that excessive feedback of the output screen grid current components is avoided.

In another embodiment of the present invention series feedback and shunt feedback are advantageously combined, in a manner to be described with reference to Fig. 6. The amplifier represented in this figure is the same as the series feedback amplifier illustrated in Fig. 1, (the series feedback coupling impedance Z_1 being now designated Z_6 .) excepting for the addition of a shunt feedback connection which may extend, for example, as shown from the last stage anode to the first stage control grid and which includes an impedance Z_7 . Associated with this phase of applicant's invention is the special object of increasing the total feedback of an amplifier above the limiting value of feedback, all other conditions being the same, that could be achieved with a single type of feedback. Another special object is to facilitate control of the input and output impedances of an amplifier as they are made to appear by feedback action. Still another object relates to compensation of spurious impedances associated with the coupling network of a series feedback type of amplifier.

Bearing in mind that the voltage fed back to the input grid in a series or shunt feedback amplifier is ordinarily a small fraction of the total output voltage, it will be understood that the series feedback impedance Z_6 is small compared with R_0 , the impedance presented by the high impedance windings of the input and output transformers. Likewise the shunt feedback impedance Z_7 is large compared with R_0 . In a typical case Z_6 may be $R_0/100$ and Z_7 may be $100R_0$. Accordingly it may be supposed to a fair degree of approximation that these ratios are so extreme that R_0 in series with Z_6 , or R_0 in parallel with Z_7 , is substantially equivalent to R_0 alone.

The anode current I in the last stage produces

a voltage drop IR_0 across the output transformer and a voltage drop IZ_6 across impedance Z_6 . The second of these voltage drops is the series feedback voltage. The shunt feedback path comprises a potentiometer made up of impedance Z_7 and the R_0 of the input transformer. With the assumption that Z_7 is much larger than R_0 , the potentiometer causes a fraction R_0/Z_7 of the voltage across the output transformer to be fed back by this path. The total feedback voltage E is therefore

$$E = IZ_6 + I \frac{R_0^2}{Z_7} \quad (1)$$

$$\frac{E}{I} = Z_t = Z_6 + \frac{R_0^2}{Z_7} = Z_6 + R_0^2 Y_7 \quad (2)$$

where Y_7 is the reciprocal of Z_7 and Z_t is the transfer impedance from the output stage anode to the input stage control grid. Ordinarily the ratio of Z_t to $2R_0$ can be equated to the feedback loss β , which is equal to the final amplifier gain.

The foregoing equations can be employed in cases in accordance with the invention where greater feedback is desired than can be obtained by one feedback path alone. Thus if the total desired feedback is measured by the transfer impedance Z_t , and the impedance Z_6 which can actually be realized is known, the supplementary, shunt feedback can be determined from the relation

$$Y_7 = \frac{Z_t - Z_6}{R_0^2} \quad (3)$$

Reciprocally, if the maximum obtainable shunt feedback is known the required series feedback impedance Z_6 to bring the total feedback to a desired greater value can be ascertained.

If, as indicated schematically in Fig. 7, the series feedback coupling impedance comprises a resistance K in parallel with another impedance Z , particularly simple relations are obtained for the case of constant total $\mu\beta$ or feedback. In this case,

$$Y_7 = \frac{1}{R_0^2} \left(K - \frac{KZ}{K+Z} \right) = \frac{1}{R_0^2} \frac{K^2}{K+Z} \quad (4)$$

The necessary shunt feedback impedance Z_7 is then

$$Z_7 = \frac{R_0^2}{K} + \frac{R_0^2}{K^2} Z \quad (5)$$

In other words, the required shunt feedback impedance is a resistance of a certain magnitude in series with an impedance that is a multiple, integral or non-integral, of impedance Z .

The impedance Z shunting the resistance K in Fig. 7 may actually correspond, for example, to an unavoidable parasitic or spurious impedance such, for example, as the capacitance to ground of the input and output transformers. The effect of such capacitance in a series feedback amplifier has been considered hereinbefore and it has been shown that the stray capacitance may reduce the feedback practically to zero before the top of the signal band is reached. To compensate for the spurious capacitance a shunt feedback circuit is provided as in Fig. 7 in which the feedback impedance comprises a capacitance and resistance in series with each other and of such respective magnitude as to satisfy Equation 5. The division of feedback between the two feedback paths, shunt and series, is shown diagrammatically in Fig. 8 for a typical case.

If the parallel impedance Z is more complicated, a more complex division of the frequency

spectrum between the two types of feedback is obtained. For example, if the impedance Z in Fig. 7 comprises in addition to the transformer capacitance a parallel connected inductance element or branch, the feedback division might appear as in Fig. 9, the total feedback remaining constant over the frequency spectrum. The inductive branch assumed may correspond, for example, to an anode current source in series with a choke coil.

Although in the examples described with reference to Fig. 7 the impedance Z has been assumed to represent a parasitic element, it may also be regarded as an impedance branch added deliberately to the circuit to secure some desirable result. For example, it is well known that the apparent or active input and/or output impedance of a feedback amplifier is not necessarily the same as its passive impedance but rather a function of the feedback. In a series feedback amplifier the active impedance Z_a is much larger than the passive impedance Z_p while in a shunt feedback amplifier the converse is true. More specifically, for series feedback,

$$Z_a = Z_p(1 - \mu\beta) \quad (6)$$

and for shunt feedback,

$$Z_a = Z_p / (1 - \mu\beta) \quad (7)$$

In Fig. 7, however, the apparent or active impedance is a function of the relative amounts of series and shunt feedback and the impedance may therefore be adjusted at will by design or adjustment of impedance Z . Thus, by choosing a value for Z which is of the proper order of magnitude and which varies in the desired manner with frequency an active amplifier impedance can be established which follows any prescribed course with respect to frequency at any magnitude between the very high impedance produced by pure series feedback and the very low impedance produced by pure shunt feedback.

Since the series feedback is large and the shunt feedback small when impedance Z is large and conversely for small values of Z , it is evident that the active impedance Z_a must be large when Z is large and small when Z is small. The exact relation for large values of $\mu\beta$ is quite simple and can be shown to be

$$Z_a = \frac{Z_0 Z_t}{R_0} = \frac{R_0}{K} Z \quad (8)$$

In other words the active impedance is merely a multiple of whatever impedance Z is inserted.

Whereas Equation 8 assumes constant $\mu\beta$, any desired active impedance Z_a can be obtained for any given feedback characteristic, represented by Z_t , by designing the shunt and series feedback impedances in accordance with the following relations:

$$Z_0 = \frac{Z_t}{1 + \frac{R_0}{Z_a}} \quad (9)$$

$$Y_t = \frac{1}{R_0} \frac{Z_t}{R_0 + Z_a} \quad (10)$$

The provision of shunt feedback for controlling the input impedance and output impedance of a feedback amplifier or to compensate for parasitic impedance elements associated with a coupling impedance in the series feedback path may be applied equally well to the cathode type of feedback amplifier, as will be apparent from a comparison of Figs. 1 and 2.

In further illustration of the principles under-

lying the present invention two examples involving bridge type feedback will be described with reference to Figs. 10 to 12. In the Fig. 10 amplifier the input and output circuits comprise resistance bridges providing bridge type feedback through a beta circuit path that includes a conventional constant resistance equalizer containing the disposable impedance branches Z_{11} and Z_{21} . A shunt feedback path comprising the impedance Z_7 is provided as in Fig. 6. The equalizer in the bridge feedback circuit is introduced to permit the feedback through the bridge path to be controlled as a function of frequency. The same control can be had by transferring the impedance elements Z_{21} and Z_{11} to the input circuit bridge as indicated in Fig. 12. For a constant resistance equalizer the voltage E_1 fed back to the input control grid through the bridge feedback circuit is

$$E_1 = \frac{K_1}{1 + \frac{Z_{11}}{R_0}} \quad (11)$$

where the denominator represents the frequency variation introduced by the equalizer and K_1 represents the constant losses of the two bridges.

For extreme impedance levels, that is, for very large values of Z_7 , the feedback through the shunt path is

$$E_7 = K_2 Y_7 \quad (12)$$

where K_2 is a constant determined by the bridge impedances. If a constant total feedback K_1 is desired, then

$$Z_7 = \frac{K_2}{K_1} + \frac{K_2}{R_0 K_1} Z_{21} \quad (13)$$

that is, a resistance in series with a multiple of Z_{21} . The active impedance presented by the Fig. 10 amplifier resembles a constant resistance, corresponding to the impedance resulting from bridge feedback, in parallel with an impedance Z_{21} . Over a frequency range where Z_{21} is large compared with the parallel resistance noted, the active impedance is therefore a constant resistance, but for frequencies such that Z_{21} is small compared with the parallel resistance the active amplifier impedance will approximately correspond to Z_{21} .

In the amplifier shown schematically in Fig. 11 bridge feedback as described with reference to Figs. 10 and 12 is combined with normal series feedback introduced by a coupling impedance Z_6 . For constant total feedback the series feedback impedance Z_6 should vary with frequency in proportion to the following relation:

$$Z_6 = \frac{K_1}{R_0 K_2} \frac{R_0 Z_{11}}{R_0 + Z_{11}} \quad (14)$$

that is, as a multiple of Z_{11} in parallel with R_0 . In this case the amplifier impedance will be inverse to that obtained in the case of Fig. 10 and it will vary approximately as the series combination of Z_{11} and the constant resistance.

With respect to the combinations illustrated in Figs. 10 to 12, it is noted that the asymptotic characteristics of a bridge type feedback circuit are generally poor because of high losses in the input and output bridges which make it difficult to obtain large values of feedback in the signal band. The combination disclosed permits the bridge feedback to be maintained over at least part of the band but still permits the eventual feedback to occur in an asymptotically superior path. These combinations may also be found useful in obtaining amplifier impedances depart-

ing slightly from a pure resistance, as would be desirable in the case of an amplifier intended to match a cable circuit used for voice frequencies or low carrier frequencies.

What is claimed is:

1. An electric wave amplifier comprising a plurality of amplifying stages and a plurality of feedback circuits embracing a plurality of said stages, said feedback circuits providing stabilizing feedback of respectively different types and each being so proportioned as to be principally effective in a respective portion of the frequency spectrum.

2. An electric wave amplifier adapted to amplify signal currents occupying a frequency range of at least several octaves, said amplifier comprising a plurality of amplifying stages and a plurality of gain stabilizing feedback circuits embracing said stages, said feedback circuits providing feedback of respectively different types, and the feedback frequency characteristics of said circuits being so proportioned that the several types of feedback are preponderant in respectively different portions of the frequency spectrum and the total feedback provided by said feedback circuits is substantially constant over the frequency range occupied by said signals.

3. An electric wave amplifier comprising an odd number of amplifying stages, each of said stages comprising a space discharge amplifying device having a plurality of electrodes including a cathode, a feedback circuit of the cathode feedback type embracing said stages and comprising feedback coupling means providing cathode-to-ground wave impedance that is substantially greater in the first and last of said stages than in an intermediate stage, and a feedback circuit of the normal series feedback type embracing said stages.

4. An amplifier comprising a plurality of amplifying stages, a feedback circuit of the cathode feedback type embracing said stages and a feedback circuit of the normal series feedback type embracing said stages, said cathode feedback predominating at high frequencies and said series feedback predominating at lower frequencies.

5. A multistage amplifier having amplifier input and output transformers, a feedback circuit of the normal series type comprising said transformers, and a feedback circuit of the cathode type comprising said transformers, said feedback circuit of the normal series type comprising a feedback coupling impedance that has negligible impedance in the asymptotic frequency range of said amplifier.

6. A multistage amplifier comprising amplifier input and output transformers, a voltage-voltage feedback circuit comprising said transformers and a feedback coupling impedance across said feedback circuit, said coupling impedance being grounded at a point electrically intermediate its terminals.

7. An electric wave amplifier, a negative feedback circuit of one type for said amplifier comprising a feedback coupling impedance, said coupling impedance comprising two series-connected sections having different impedance-frequency characteristics, a second negative feedback circuit of another type for said amplifier comprising at least one of said sections, said sections being so

proportioned that the total feedback provided by said feedback circuits is substantially constant over a wide frequency range.

8. A broad band amplifier comprising a plurality of amplifying stages and input and output transformers, a stabilizing feedback circuit of the voltage-voltage type embracing the proximate windings of said transformers and comprising a feedback coupling impedance, said coupling impedance comprising at least two series-connected sections grounded at their junction, and said sections having such respective impedance-frequency characteristics that the feedback is of the normal series type in one frequency range and of the cathode type in another frequency range.

9. An electric wave amplifier comprising at least three amplifying stages, a feedback circuit embracing said stages, said feedback circuit comprising a feedback coupling impedance comprising two series-connected sections shunted across said circuit, said sections being grounded at their junction, said sections having respective impedance-frequency characteristics such that in a transition frequency range the impedances of said sections vary in opposite senses with respect to frequency, the series impedance of said two sections being constant throughout said transition frequency range.

10. An amplifier adapted for the direct amplification of television signals or the like comprising space discharge amplifying devices in a plurality of stages, a cathode feedback circuit and a series feedback circuit embracing said stages, a feedback coupling impedance common to said circuits and grounded at a point electrically between its terminals, and a space discharge current source connected in said series feedback circuit.

11. In an electric wave amplifier, the method of controlling the apparent impedance of said amplifier which comprises controllably feeding back waves amplified therein to the input of said amplifier in such relation as to tend to increase the apparent impedance of said amplifier, concurrently feeding back waves amplified therein to the said input in such relation as to tend to decrease the apparent impedance of said amplifier, and so adjusting the relative preponderance of the two feedbacks in different portions of the frequency range that the said apparent input impedance varies in a predetermined manner over the frequency range.

12. An electric wave amplifier comprising a plurality of amplifying stages and feedback circuits of the series and shunt types respectively embracing said stages, said feedback circuits being so proportioned as to be principally effective in respectively different portions of the frequency spectrum.

13. An amplifier in accordance with claim 12 in which the total $\mu\beta$ of said feedback circuits combined is constant over the frequency range of the waves being amplified.

14. An electric wave amplifier comprising a plurality of amplifying stages and feedback circuits of the bridge and shunt types respectively embracing said stages.

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