March 30, 1943. J. W. McRAE 2,315,442

NEGATIVE FEEDBACK DETECTORS

Filed Feb. 11, 1942 3 Sheets-Sheet 1

FIG. 1

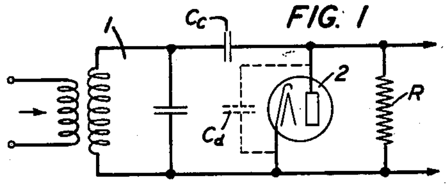


FIG. 1A

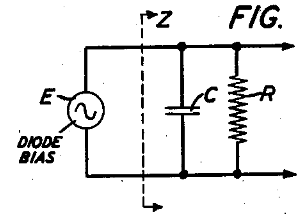


FIG. 2

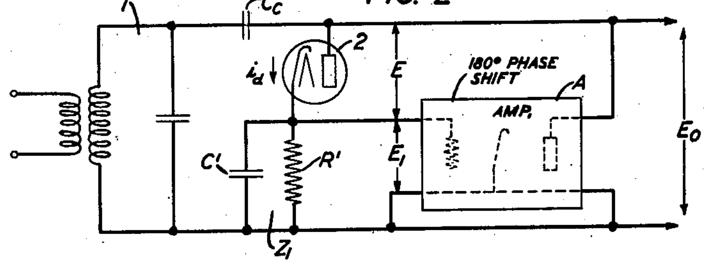


FIG. 3

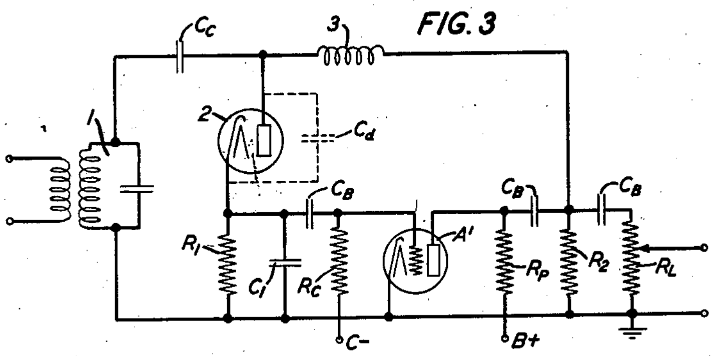
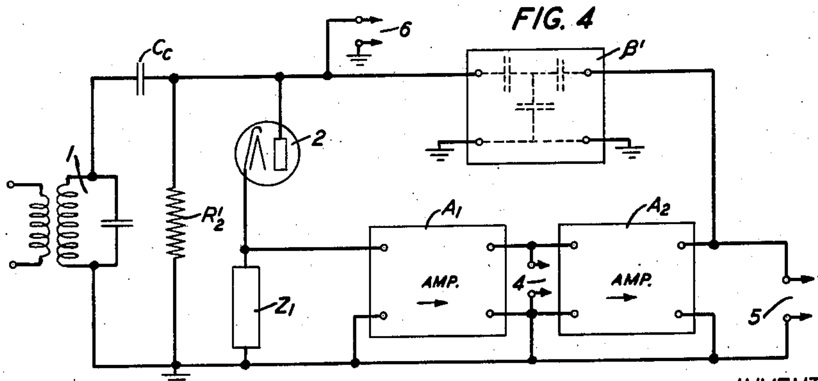


FIG. 4



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NEGATIVE FEEDBACK DETECTORS

Filed Feb. 11, 1942 3 Sheets-Sheet 2

FIG. 5

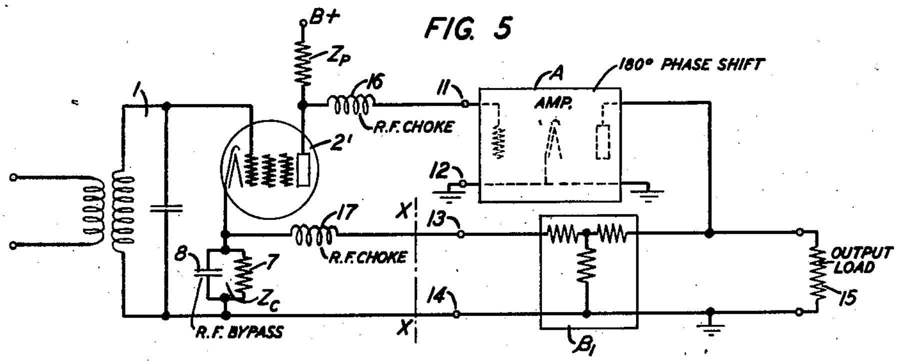


FIG. 5A

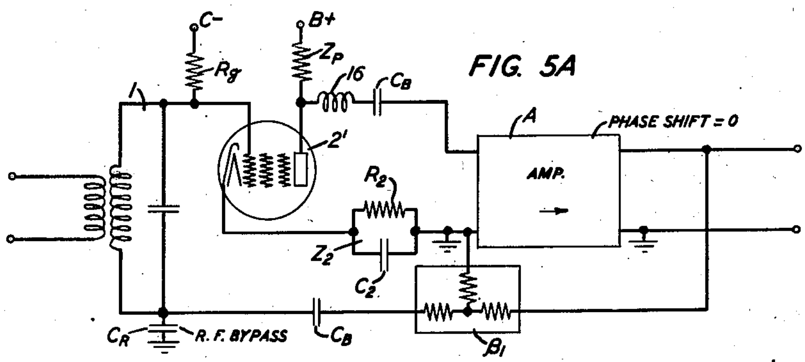


FIG. 5C

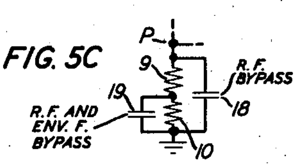
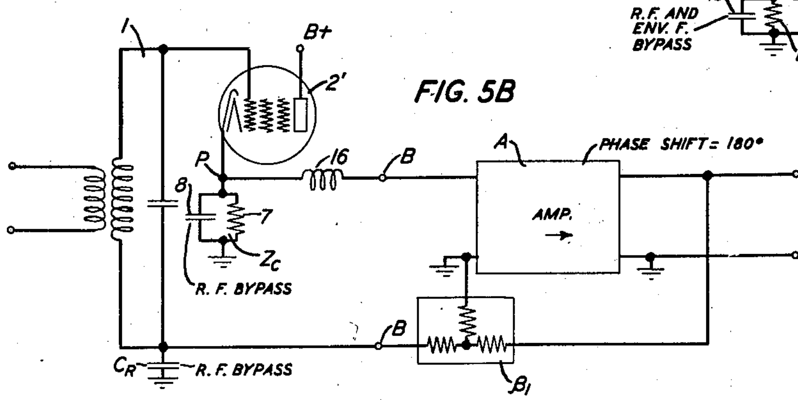


FIG. 5B



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NEGATIVE FEEDBACK DETECTORS

Filed Feb. 11, 1942 3 Sheets-Sheet 3

FIG. 6

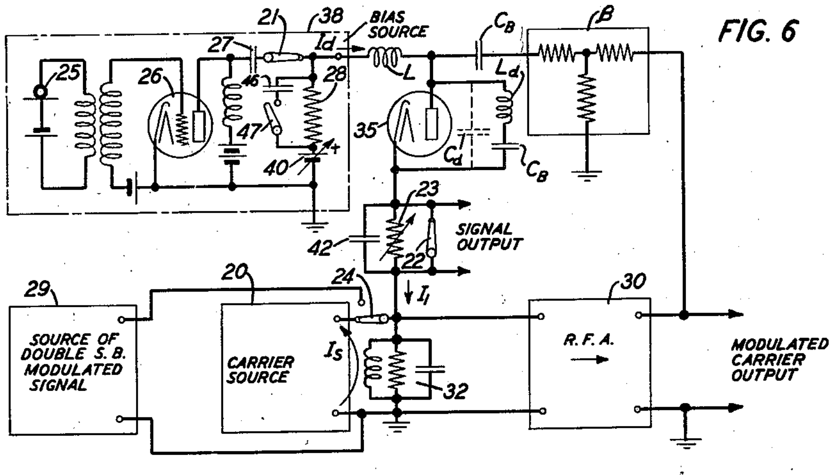
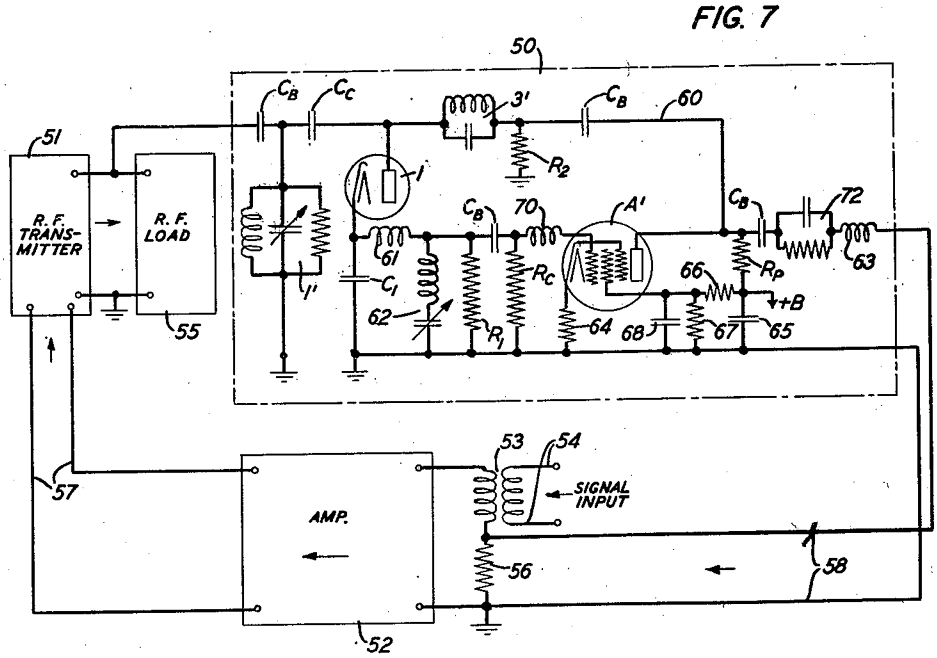


FIG. 7



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Patented Mar. 30, 1943

2,315,442

UNITED STATES PATENT OFFICE

2,315,442

NEGATIVE FEEDBACK DETECTOR

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Application February 11, 1942, Serial No. 430,383

8 Claims. (Cl. 250—27)

This invention relates to modulation, especially detection.

An object of the invention is to increase fidelity of detection or demodulation.

In one specific aspect of the invention an amplifier operating at envelope frequencies and having a large amount of negative feedback is so associated with a self-biasing detector that the feedback not only reduces distortion and phase shift between the detector bias voltage and the amplifier output but so modifies the apparent detector load impedance as to increase fidelity of detection.

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

Fig. 1 is a detector circuit diagram for facilitating explanation of the invention;

Fig. 1A is the envelope frequency equivalent circuit corresponding to Fig. 1;

Fig. 2 shows a detector circuit embodying a form of the invention;

Fig. 3 shows a modified form of circuit with means for feeding back envelope frequency currents to the exclusion of direct current;

Fig. 4 shows a generalized form of detector circuit from which circuits such as those of Figs. 2 and 3 can be derived;

Figs. 5, 5A and 5B show three modified forms of detector circuits employing detector tubes having in addition to the anode and cathode at least one control element;

Fig. 5C shows a modification of a portion of the circuit of Fig. 5B;

Fig. 6 shows a circuit adapted for modulation of a carrier wave or demodulation of a modulated wave; and

Fig. 7 shows an envelope frequency feedback system with the feedback path including a detector circuit generally similar to the detector circuit of Fig. 3.

A fundamental circuit for a self-biasing diode detector is shown in Fig. 1, wherein 1 is a tuned radio frequency input circuit or circuit to which the signal modulated carrier waves to be demodulated are applied, 2 is a diode having inherent plate-cathode capacity C_d, R is a load resistance across which the detected signal voltage is desired, and C_c is a blocking condenser for the detected signal but a low or negligible impedance for the input waves. The signal to distortion ratio of such detectors is limited in two ways. First, there is the non-linearity inherent in the diode characteristic and, second, there is the effect of the frequency characteristic of the load impedance Z, which is shown in its simplest form in Fig. 1A wherein E is the diode bias voltage and $C=C_c+C_d$. In order to reduce the effect of diode non-linearity it is desirable to increase the absolute value of Z. However, because of the presence of capacities C_c and C_d, such an increase in absolute value of Z tends to result in a decrease in what may be called the critical frequency of Z, i. e. the frequency at which Z has its shunt reactance equal to its shunt resistance. Where deep modulation must be handled at relatively high envelope frequencies, excessive reduction of this critical frequency tends to result in increased distortion. For the detection of broadcast signals, where the modulation at high audio frequencies is very small, no difficulty is ordinarily encountered in securing sufficiently high values of Z. In other applications it may be found that Z cannot be made sufficiently high to overcome diode non-linearity without making its critical frequency too low. An example of this kind of application is the beta-circuit demodulator (i. e. the demodulator in the envelope feedback circuit) in multi-plex radio transmitter circuits employing envelope feedback, (for instance circuits of the general type of the circuit of Fig. 7 described hereinafter). Additional examples may be found in some forms of measuring equipment and also possibly in television receivers.

This difficulty can be overcome in many cases with the help of a negative feedback amplifier responding to frequencies in the envelope of the incoming signal, as for example, amplifier A in Fig. 2, wherein the impedance Z_1 across the grid and cathode of the amplifier comprises resistance R′ and radio frequency by-pass capacity C′. In discussing operation of this circuit, only the direct and envelope frequency components of the diode current, and the voltages set up by these currents, need be considered. The diode current i_d (constituted by these current components) flowing through the impedance Z_1 across the grid and cathode of the amplifier A, sets up across Z_1 a voltage $E_1=i_dZ_1$ and this voltage sets up a voltage E_0 across the output terminals of the amplifier, tending to bias the diode to cutoff. The resultant diode bias voltage is $E=E_1+E_0$. Designating the amplification of the amplifier,

$$\frac{E_0}{E_1}$$

as A,

$$E_0=E\frac{A}{A+1} \quad (1)$$

and the apparent diode load impedance, which is defined as the ratio of the diode bias voltage to diode current (averaged over a number of radio frequency cycles), is

$$Z=\frac{E}{i_d}=Z_1(A+1) \qquad (2)$$

If A is large the apparent diode load impedance is therefore approximately equal to the impedance Z_1 multiplied by the amplification of the amplifier and the output voltage E_0 is almost exactly in phase with and slightly less than the diode bias voltage. If, further, A is real and constant over the required band of envelope frequencies, Z will have the same frequency characteristics as Z_1, and since Z_1 may be made much smaller than Z, it may be made to have much more desirable characteristics than a physical impedance of the magnitude of Z. By taking advantage of this fact, distortion in the diode may be kept small, even for signals containing a wide band of envelope frequencies.

The negative bias on the diode, tending to cut it off, is obtained from the output of the amplifier rather than directly from the flow of current through a physical impedance of magnitude Z. Thus the diode current, which is only required to control the input voltage of the amplifier, may be considerably reduced from the same diode bias voltage, and since the apparent diode load impedance is the ratio of diode bias to diode current this reduction is equivalent to an increase in the apparent diode load impedance Z. More important, the fact that this increase in impedance is obtained by the action of a feedback amplifier means that the variation of Z with frequency may be made more satisfactory than in the case of a purely passive physical impedance. In this way, distortion generated in the diode in many cases may be made considerably smaller than in the simple diode circuits (i. e., the diode circuits without the feedback amplifier). The circuit configuration results in a large amount of feedback around the amplifier. In addition to its effect upon the apparent diode load impedance, this feedback tends to reduce distortion generated in the amplifier itself. Thus, a relatively large detected output, with relatively low distortion, may be obtained with only a small input. Besides the resultant high detector input impedance, this means that sufficiently low distortion may often be obtained with smaller diodes than would normally be required. At ultra-high carrier frequencies this is sometimes an important advantage.

Because of the presence of feedback, the envelope frequency circuits should be designed to pass a frequency band somewhat wider than the required envelope band, in order to facilitate proper shaping of the mu-beta loop transmission characteristic as disclosed in H. W. Bode Patent 2,123,178, July 12, 1938. Therefore, to avoid difficulties in keeping radio frequencies out of the amplifier, it is desirable to have the highest envelope frequency low compared to the carrier frequency.

The amplifier may have any desired number of stages, provided the phase of the feedback is maintained correct for the desired negative feedback action. In the envelope frequency band the phase-shift through the amplifier is 180 degrees. Although the amplifier may be capable of responding to direct as well as alternating currents and voltages, still more desirable operation, (in particular, even more desirable impedance characteristics), may be obtained by a modified circuit that makes use of an amplifier which does not transmit direct current but is responsive only to alternating voltage, as for example, the circuit of Fig. 3.

In this circuit of Fig. 3, condensers C_B are blocking condensers for direct current. A radio-frequency choke 3 isolates the plate of the diode 2 from the envelope-frequency amplifier A' at radio frequencies, to prevent overloading the amplifier. R_c is a grid leak resistor, R_p a plate resistor, R_2 a coupling resistor and R_L, a load resistor for the amplifier. There is a phase reversal in transmission at envelope frequencies through the amplifier. C_1 is a radio-frequency by-pass condenser. The average diode bias corresponding to the carrier is set up across the resistors R_1 and R_2 in series, the amplifier being isolated from this voltage by the blocking condensers C_B. On the other hand, the apparent impedance presented to the alternating components of envelope frequency in the diode current is still given by Equation 2. The effect of the blocking condensers must of course be considered in computing the amplification A of the amplifier A' and, except at the lowest envelope frequencies, Z_1 now includes the effect of the grid resistor R_c at the grid of the amplifier tube, (or at the grid of the first amplifier tube where the amplifier is a multistage amplifier).

This arrangement makes it possible to select the direct and alternating components of the apparent diode load impedance independently. In this way the occurrence of the most serious type of distortion in the diode can be avoided. This type of distortion is often known as "non-tracking" distortion since it occurs when the envelope of the signal is decreasing at too great a rate for the bias voltage to follow or "track." When the diode load impedance consists of a simple resistance-capacitance parallel combination, the criterion for the occurrence of this kind of distortion may be simply related to the time constant of the load impedance. However, in more complicated cases, such as those under consideration here, another approach seems desirable.

Non-tracking distortion occurs when the alternating component of diode current becomes greater than the direct, or carrier component. In that case, if there were to be no distortion, the diode current would have to become negative during a portion of the envelope cycle. Since this is impossible, the result follows that the negative peak of the diode alternating current is "cut off" and during this interval the load is isolated from the incoming signal. Thus the output voltage no longer follows the envelope but simply decreases in transient fashion.

To avoid this kind of distortion, the peak magnitude of the ratio between envelope frequency components and the detected carrier component of diode current must be less than unity. For single tone modulation this criterion may be written

$$\frac{kR}{|Z|}\leq 1 \qquad (3)$$

where k is the modulation factor for the incoming signal voltage, R is the resistance presented to the direct (i. e., carrier) component of diode current, and Z is the apparent diode load impedance at the frequency of the envelope of the signal. Now Z usually falls off with increasing frequency. Thus a usually sufficient condition for the avoidance of non-tracking distortion, even for complex modulation, is obtained by setting up the above criterion, using the maximum total modulation factor and the maximum frequency in the envelope band. In many cases, this condition is much too severe and may be relaxed considerably. However it may be taken as the basis for the following discussion of the design procedure.

The first step in this procedure is to set the value for R (equal to R_1+R_2 in Fig. 3) sufficiently high to give the required low distortion due to diode non-linearities. Then the impedance Z_1 is decided upon, the capacitance being large enough to give the required radio frequency by-passing and the resistance then being adjusted to give a reasonably flat characteristic over the required band of envelope frequencies. The amplifier gain necessary to raise the apparent diode load impedance to a value approximating R may then be found. Since the amplification will, in general, not be flat over the entire band, it will be necessary to use Relation 2 in order to find the characteristics of Z. At the highest frequency in the band, Criterion 3 will give a minimum permissible value of Z, in terms of R and the characteristics of the expected signal. The aim of the design should be to make Z exceed this minimum as little as posible, consistent with necessary allowances for loss of gain in the amplifier as the tubes age. In addition, Z should be as flat as posible, so as not to exceed the value of R greatly at the low frequency end of the required band of envelope frequencies. Such excessive values of Z would introduce a type of distortion not as yet mentioned. However, experience indicates that even with ratios of Z to R as great as 3 this type of distortion is not important.

If more than a single stage is required in the amplifier, precautions must be taken to avoid singing around the feedback loop. The usual design methods may of course be employed for this purpose, and it will usually be found satisfactory to consider the diode as a zero impedance generator in making such calculations.

Means should also be provided to keep excessive radio frequency voltages out of the amplifier, for example filtering circuits such as shown in Fig. 7 described hereinafter. This is particularly true if the amplifier tubes are operating nearly to capacity under the given envelope frequency conditions. Then a relatively small radio frequency voltage in the amplifier may considerably increase the distortion. The effect of the necessary filtering circuits on the amplifier feedback loop must, of course, be considered in the stability calculations.

A specific design example of a detector of the type shown in Fig. 3 was one for use in a 12-channel multiplex radio telephone transmitter employing envelope feedback and operating on a carrier frequency of 141 megacycles. A single 6L6 tube was used as the amplifier. This detector was to be used (in the fashion indicated in Fig. 7 described hereinafter) to obtain a sample of the output for feeding back through the beta circuit to the input of the signal amplifier, and therefore it was important that it should introduce negligible distortion throughout the envelope frequency band and that its overall phase shift should be negligibly small up to frequencies of the order of a megacycle. The design was made to permit operation over an envelope frequency band from 12 to 108 kilocycles.

The value of R was set at 550,000 ohms, and values computed for curves (not shown) of A,

$$\frac{A}{A+1}$$

Z_1, and Z. At 108 kilocycles, the absolute value of Z was about 4 decibels higher than R (that is, about 870,000 ohms). There was consequently a large safety factor against non-tracking, even for 100 per cent modulation at 108 kilocycles. In fact, this safety factor was excessive in this application and could have been reduced by increasing R to 870,000 ohms or more. At the lower frequencies, Z increased to a maximum value of about 1.5 megohms. On the other hand, at very high frequencies, Z fell off rather rapidly, the slope being nearly 12 decibels per octave over a portion of the range. In this region the real part of Z was negative. Nevertheless, because of the low internal impedance of the diode, the circuit was found to be stable. In addition, despite the somewhat low value of R, the distortion was found to be sufficiently low.

Fig. 4 shows a general form of detector circuit of which the circuits of Figs. 2 and 3 can be considered specific types. In Fig. 4 the amplifier is shown as having a plurality of sections A_1 and A_2 in tandem, (each of any desired number of stages), with a pair of output terminals **4** between two of the sections. An attenuating or transmission control network β' may be provided between the amplifier output terminals **5** and the diode. Equalization (or control of the attenuation versus frequency and phase shift versus frequency characteristics of the feedback path) to prevent singing may be provided, at least partially, in this attenuating network, and this network may include condensers (indicated in dotted lines) preventing feedback of direct current. The presence of this network provides a choice as to the output (i. e., load) connection for the system. For example, the output of the system may be taken off at the diode (from terminals **6**, across resistor R_2', or at the amplifier output (terminals **5**), or at some intermediate point on the attenuating network β', or at terminals **4**. For each of these conditions analysis of the operation of the system would take the same general form as that presented in connection with the preceding figures of the drawings.

Figs. 5, 5A and 5B show examples of ways in which multielement tubes, that is, tubes having in addition to an anode and cathode at least one discharge control element, may be used as the detector elements cooperating with the amplifier devices. As in the case of the amplifier elements of Figs. 2 to 4, the amplifiers A of Figs. 5, 5A and 5B may be linear devices, the detector elements being the only necessarily non-linear elements of the systems. The multielement detector tubes are shown at **2'** and may be, for example, screen grid tubes or suppressor grid pentodes. The amplifiers A may be assisted to some extent in the amplification function by the gain (to envelope frequency voltages) existing in the detector tube. However since this tube is operated around its cut-off, its gain will probably be very small or may even be negative. In addition the cut-off in multielement tubes ordinarily is not as sharp as in a diode, and because of this greater non-linearity it may be necessary to present a higher equivalent load impedance to the detector than in the case of a diode.

The operation of these circuits is similar in general to those already described. In each case, the principal function of the feedback at envelope frequencies is to facilitate the maintenance of a high load impedance for the detector and to reduce the possibility of non-tracking distortion. In addition, much greater output power may be obtained from the amplifier-detector combination for a given amount of distortion than could be obtained from a simple detector. Finally, the feedback tends to reduce distortion in the amplifier in addition to its effect on the apparent detector load impedance.

One feature of these circuits which is not characteristic of the circuits using diodes is the existence of what may be an appreciable zero signal current in the detector tube. This component of current could be set up in the diode of any of the preceding circuits by the application of a positive direct current bias to the diode anode. Such a bias is known to reduce the possibility of non-tracking in a simple diode detector but it has the disadvantage of reducing the diode input impedance for small signals. The circuits of Figs. 5, 5A and 5B do not have this disadvantage and the input impedance presented to the signal will be considerably higher than that for any type of detector using a diode. This is a result of the use of a multielement tube as the detector, and of the fact that the detector may be operated with no grid current. At the same time, as will be shown later, the no-signal current in the plate-cathode circuit of the detector tube is similar to a positive bias on a diode in reducing the probability of non-tracking distortion.

Consider first the circuit of Fig. 5, which is drawn for the case in which amplifier A is a direct-current amplifier. With no signal, an equilibrium condition is established between the plate-cathode current in the detector tube 2′ and the direct current bias voltage developed across cathode impedance Z_c, which may be, for example, a resistance 7 and a carrier or radio frequency by-pass condenser 8 connected parallel and having its impedance value Z_c and its resistance value R_c.

The detector plate current flowing through plate circuit impedance Z_p, which may be a resistor or other suitable impedance of impedance value Z_p and resistance value R_p, sets up a voltage between amplifier input terminals 11 and 12 which, after amplification, appears across terminals 13 and 14 and thus across cathode impedance Z_c. The sense of this amplified voltage is such as to drive tube 2′ toward cut-off and thus only a small plate current is required to set up the no-signal bias. Let impedance Z_c' be defined as the ratio E_c/I_c, where E_c is the bias developed across Z_c and I_c is the detector tube plate-cathode current. Then this apparent impedance will be greater than the actual value of Z_c and its magnitude will be a measure of the effectiveness of the feedback in reducing the necessary detector current. The magnitude of Z_c' may be calculated in terms of Z_p, Z_c, A, β_1 and an additional impedance Z_0. Impedance Z_0 is the impedance measured between terminals 13 and 14, with the connections to Z_c broken at XX. During this measurement the output load 15 must be left connected. The apparent impedance Z_c' is then:

$$Z_c' = \left(\frac{Z_p A\beta_1 + Z_0}{Z_c + Z_0}\right) Z_c$$

where $A\beta_1$ is the voltage gain of the amplifier-attenuator combination measured between input terminals 11 and 12 and output terminals 13 and 14, with the load connected and with the leads to Z_c broken at XX. The factor in parentheses evidently multiplies the value of Z_c up to its apparent value Z_c'. This expression is evidently valid at any frequency and reduces to the form

$$R_c' = \left(\frac{R_p A\beta_1 + R_0}{R_c + R_0}\right) R_c$$

at zero frequency.

Having established the value of apparent impedance in the cathode lead of the detector tube, the analysis of the system becomes relatively simple. When a steady carrier is applied through input circuit 1 an additional direct current component of current flows through tube 2′, of sufficient magnitude to develop across Z_c a voltage E very nearly equal to the radio frequency voltage between grid and cathode of tube 2′. (Z_c is arranged to include a path, as for example the path through condenser 8, of low impedance at the carrier frequency so that practically the entire applied voltage appears between the grid and cathode of the tube.) Thus there are now two components of current through the detector tube, the no-signal component and the component due to the signal carrier. These two components may be written as

$$I_0 + \frac{E}{R_c'}$$

or

$$\frac{E_b}{R_c'} + \frac{E}{R_c'}$$

where E_b is the voltage developed across Z_c, or between grid and cathode of tube 2′, in the absence of signal.

When the carrier is modulated sinusoidally with a modulation factor k, the voltage across Z_c must vary proportionally and a third component of current therefore must flow in tube 2′. This component is of envelope frequency and has an amplitude

$$\frac{kE}{|Z_c'|}$$

where Z_c' is now evaluated at the envelope frequency in question. Non-tracking distortion cannot occur if the amplitude of this envelope frequency component of detector current is less than the sum of the two direct current components. For no non-tracking, therefore,

$$\frac{kE}{|Z_c'|} \leq \frac{E_b}{R_c'} + \frac{E}{R_c'}$$

or

$$\frac{kR_c'}{|Z_c'|} \cdot \frac{E}{E_b + E} \leq 1$$

comparing this criterion with that for the diode, it will be observed that the existence of the factor E/E_b+E tends to reduce the difficulty of preventing non-tracking distortion.

The circuit shown could be modified in several ways without changing its method of operation. For example, attenuator β_1 might be omitted, or the output obtained from terminals 13 and 14 instead of as shown.

More important modifications would be required in order to make the amplifier responsive only to alternating voltages. In that case, blocking condensers (not shown) could be inserted at the points marked 11 and 13. Then the direct current path through Z_c should be made to have a high resistance R_c, comparable in magnitude to Z_c', as computed for envelope frequencies in the required range. This computation should be made from the same formulas as for the direct current amplifier. The non-tracking criterion for this case is

$$\frac{kR_c}{|Z_c'|}\cdot\frac{E}{E_b+E}\leq 1$$

From this expression, it will be evident that R_c should be about equal to $|Z_c'|$. It should not exceed $|Z_c'|$ by more than, say, a factor of 2.

Radio frequency chokes 16 and 17 prevent the incoming radio frequency waves from being fed back or from reaching the amplifier A.

In Fig. 5A, the amplifier is effective in reducing the current through tube 2′ in much the same way as in Fig. 5. The most important difference between the two circuits is that the envelope frequency bias voltage set up by the signal appears on the grid in Fig. 5A and on the cathode in Fig. 5. Since Fig. 5A is drawn for the case in which the amplifier is isolated from direct voltages, the direct current biases due to the no-signal and carrier components of current are developed on the cathode, across resistance R_2.

Blocking condensers C_B isolate the amplifier from direct currents, and condenser C_2 by-passes resistance R_2. These three condensers all have low reactance for envelope frequencies and C_2 must also have a low reactance at radio frequencies. Condenser C_R must have a high reactance for envelope frequencies and a low reactance for radio frequencies. The resistance R_g is a high resistance grid leak used to maintain the grid of tube 2′ at a definite average voltage C—. The zero-signal current may be adjusted by varying the voltage C—. Since resistance R_2 is large, the necessary zero signal current tube 2′ is small.

When a steady carrier is applied to tube 2′ through tuned circuit 1, the current through R_2 increases an amount sufficient to give an increase in voltage across R_2 nearly equal to the magnitude E of the carrier voltage. The two direct current components of detector current are therefore

$$\frac{E_b}{R^2}+\frac{E}{R^2}$$

where E_b is the zero signal bias across R_2.

The envelope frequency component of detector current must be of such magnitude as to develop a voltage of magnitude almost equal to kE on the grid. This requires a current I_p through Z_p and tube 2′ of magnitude

$$I_p=\frac{kE}{Z_pA\beta_1}$$

From this expression, it is evident that an apparent transfer impedance Z' may be defined such that $Z'=kE/I_p$ where kE is the magnitude of the input modulation envelope and I_p is the envelope frequency component of detector current. In that case $Z'=Z_pA\beta_1$ and the criterion for no non-tracking is

$$\frac{kR_2}{|Z'|}\cdot\frac{E}{E_b+E}\leq 1$$

As before, R_2 should be approximately equal to Z'.

This circuit may be modified to permit the use of a direct current amplifier. In that case, the blocking condensers C_B, and impedance Z_2 should be omitted, (i. e., short-circuited). Grid leak resistor R_g and bias supply C— may also be omitted. The zero signal equilibrium current through tube 2′ will then depend on the constants of the amplifier and tube 2′. The apparent transfer impedance Z′ will have the same value as before, but the resistance R_2 in the above expressions must be replaced by the value to which Z′ reduces at zero frequency. Let this value be R′. The equivalent bias voltage E_b must be defined as the product of the zero signal current through tube 2′ and the resistance R′. With these definitions made, the criterion for no non-tracking becomes

$$\frac{kR'}{|Z'|}\cdot\frac{E}{E_b+E}\leq 1$$

The circuit of Fig. 5B differs from both the preceding circuits in that the amplifier input voltage is obtained from an impedance in the cathode lead, rather than the plate lead, of tube 2′. Otherwise, it most nearly resembles Fig. 5A. However, the amplifier in Fig. 5B must have essentially 180 degree phase shift throughout the envelope frequency band, as opposed to zero degrees in Fig. 5A.

In the form shown, the circuit is adapted for a direct current amplifier. In that case, the zero-signal current through tube 2′ depends on the characteristics of the tube 2′ and of amplifier A and attenuator β_1. However, the entire detector tube current flows through impedance Z_c and the bias voltage E_b which will appear in the no non-tracking criterion is the voltage appearing across Z_c in the absence of signal.

The apparent impedance Z_c' which limits the flow of detector tube plate-cathode current is

$$Z_c'=Z_cA\beta_1$$

This reduces to R_c' at zero frequency. Following the same type of reasoning as outlined above, the criterion for no non-tracking is

$$\frac{kR_c'}{|Z_c'|}\cdot\frac{E}{E_b+E}\leq 1$$

where all the symbols have already been defined.

The circuit of Fig. 5B may be converted for the use of an amplifier isolated from direct voltages by following the general procedure already described. Blocking condensers, as shown at C_B in Fig. 5A should be inserted at the points marked B in Fig. 5B and a grid leak resistor and C-supply arranged to hold the grid of tube 2′ in Fig. 5B at the desired steady bias voltage as resistor R_g and the C-supply in Fig. 5A bias the grid of tube 2′ in Fig. 5A. The direct current path in Z_c should be made a high resistance of magnitude approximately equal to Z_c'. Apparent impedance Z_c' should be calculated in the same way as before, and the criterion for no non-tracking becomes

$$\frac{kR_c}{|Z_c'|}\cdot\frac{E}{E_b+E}\leq 1$$

where R_c is the resistance of the direct current path in Z_c.

Fig. 5C shows a form of cathode-lead network suitable as a substitute for that shown between point P and ground in Fig. 5B when the amplifier is to be an alternating current amplifier (i. e., when the amplifier is to be isolated for direct current). In Fig. 5C the cathode lead impedance comprises resistances 9 and 10 in series, with a radio-frequency by-pass condenser 18 (of high reactance at the envelope frequencies) connected across both 9 and 10, and with resistance 19 shunted by a by-pass condenser 19 which has a low reactance for the envelope frequencies as well as for the radio or carrier frequency.

As mentioned under Fig. 5, the output load in any of these three circuits may be connected on the output side of attenuator β_1 or the attenuator may be omitted. The amplifier A and attenuator β_1 must be designed to have a suitable characteristic to obtain stability against singing around the feedback loop. The zero-signal equilibrium point of the direct current amplifiers used in any of these three circuits should be adjusted to give a suitable value of E_b to minimize non-tracking. In cases not employing direct current amplifiers, this voltage E_b may be adjusted by changes in the plate supply voltage or the C-voltage supply. In many practical cases it will also be necessary to include suitable radio-frequency choke coils and by-pass condensers in order to keep spurious radio-frequency voltages out of the amplifier.

As indicated above, the amplifier A of Figs. 5, 5A and 5B may be a linear amplifier, the only necessarily non-linear device in the system being the detector tube that feeds the amplifier.

Fig. 6 shows a system adapted for modulation of a carrier wave when the wave to be modulated, for example, a radio-frequency carrier wave, is supplied by source 20, through switch 24, and adapted for demodulation of a modulated wave when the modulated wave, for example, a signal modulated radio-frequency carrier wave, is supplied by source 29, through the switch 26. Either or both of the switches 21 and 22 may be omitted, if desired. However, if desired, switch 22 may be in the closed condition, during modulation, to short-circuit impedance 23; and, if desired, switch 21 may be open, during demodulation, to remove from circuit the modulating source. The modulating source may be provided, for example, by telephone transmitter 25 and amplifier 26 adapted to supply speech waves through blocking condenser 27 to resistance 28. During demodulation, the terminals of resistance 23 are the output terminals of the system, across which the signal voltage appears; and during modulation the output terminals of radio-frequency amplifier 30, which may be a linear amplifier with a constant gain of several times 10 decibels, for example, are the output terminals of the system, across which the modulated carrier voltage appears. The input of amplifier 30 is connected across impedance 32, which is an anti-resonant circuit at the carrier frequency of the waves delivered thereto by source 20 and may have zero impedance for direct current and for the signal frequency.

In modulating a carrier wave with a signal, the system is a double sideband modulator. Radio-frequency amplifier 30 is capable of amplifying a band of frequencies centering on the carrier frequency. Its output voltage, modified if desired by a beta-circuit network β shown as a network of generalized impedances, is fed back to its input through a biased diode 35 in series in the feedback circuit, the bias voltage for the diode being a voltage proportional to the desired signal envelope. This bias voltage is obtained from bias source 38 as the sum of the biasing voltage of a battery or direct current source 40 and the voltage developed across resistance 28 by the diode current, and is applied through the circuit extending from ground, through battery 40, resistance 28, plate-cathode path in diode 35, switch 22 and circuit 32 back to ground. Even though the bias from source 40 is positive, for reasons to be described later, the bias developed by the diode current in resistance 23 is large enough to make the net bias on the diode negative and nearly equal to the instantaneous magnitude of the radio-frequency voltage across the diode. Whenever (a peak of) the radio-frequency output voltage of amplifier 20 exceeds the bias voltage on the diode, a pulse of current passes through the diode. The phases in the amplifier are adjusted so that the pulse is applied to the input of the amplifier in such a phase as to tend to reduce the input voltage (across impedance 32) and consequently the amplifier output voltage. Thus the (radio-frequency peak) output voltage is held very close to the applied bias on the diode, provided the amplifier 20 has large gain. This results in very small envelope distortion. It should be noted that the diode capacitance C_d is anti-resonated by inductance L_d in order that the total available radio frequency voltage may be applied between the diode anode and cathode. Blocking condensers C_B and by-pass condenser 42 have low reactance at the signal carrier frequency but high reactance at the envelope or signal frequency. Choke coil L is inserted between the diode anode and the bias source 38 in order that radio-frequency voltages may be kept out of the bias source.

In order to obtain the relation necessary for the design of a circuit of this type, an analysis similar to those already made must be performed. As before the diode current may be divided into three components, a direct current component due to the bias battery 40, another direct current component due to the carrier of the radio frequency voltage and a third component of envelope or signal frequency. Each of these currents flows through the diode and through impedance 32 in a series of short pulses, occurring at a rate corresponding to the carrier frequency. Thus each current may be analyzed into its average value and a series of Fourier components. The average value, at least in the case of the envelope frequency component, must be average over only a few radio frequency periods. The lowest frequency in the Fourier series will be the carrier frequency, the higher frequencies all being multiples of this fundamental frequency. Only the average value and the carrier frequency component are of interest here and, since the duration of each pulse is very small, there will be a constant ratio between the average value and the carrier frequency component. Let this ratio be M or, in other words, assume that the average value of each component of diode current is M times the carrier frequency component of that same current.

Before proceeding with the analysis, it will be necessary to define several further constants. First, let it be assumed that the radio frequency source 20 supplies a constant radio frequency current I_s which flows through impedance 32 thus producing a radio frequency voltage which is amplified by the radio frequency amplifier. When the diode cathode is not heated, a voltage E_0 consequently appears between its anode and cathode. When the diode cathode is heated and diode current flows through impedance 32, this voltage is reduced to a new value E_c, corresponding to the unmodulated carrier condition. Voltage E_c must evidently be less than half voltage E_0 in order that the carrier may be completely modulated.

The magnitude of the carrier frequency voltage E_0 is given by

$$E_0 = I_s Z A\beta = I_s Z' \quad (1)$$

where Z is the impedance of network **32** as the carrier frequency, A is the gain of the RF amplifier and β is the loss in the attenuator. The quantity $ZA\beta$ is equivalent to an apparent transfer impedance Z' where the prime refers to the fact that the impedance is to be evaluated at the carrier frequency.

In the same way, the voltage E_c is given by

$$E_c = (I_s - I_1')Z' \quad (2)$$

where I_1' is the carrier frequency component of the diode current with switch **22** closed and switch **21** open. Thus,

$$(E_0 - E_c) = I_1'Z' \quad (3)$$

and under these conditions the average value of diode current will be

$$I_d' = MI_1' + \frac{E_b}{R} = \frac{E_c}{R} + \frac{E_b}{R} \quad (4)$$

where E_b is the voltage of bias source **40** and R is the resistance of resistance **28**, the factor M having already been defined. This relation expresses the fact that there are two direct current components of diode current, the first being that sufficient to build up a negative bias on the diode approximately equal to the magnitude of voltage E_c, while the second is the component of current due to the bias source **40**. From these four equations, inserting the requirement that E_0 must be at least twice E_c, the following criterion for the adjustment of resistance R and impedance Z' may be established:

$$\frac{MR}{|Z'|} \leq 1 \quad (5)$$

Observance of this criterion will insure that proper operating conditions are obtained for the case of an unmodulated carrier.

In order to determine the conditions necessary for operation with switch **21** closed and a sinusoidal modulating voltage supplied at the terminals of the bias source, it will be necessary to consider the magnitude of the envelope frequency component of current through the diode. Using double primes to represent quantities evaluated at envelope or side band frequencies, this current becomes

$$I_d'' = MI_1'' = \frac{MkE_c}{|Z''|} \quad (6)$$

where k is the modulation factor for the resultant modulated frequency voltage appearing across the diode. The condition necessary for no non-tracking distortion may therefore be written

$$\frac{I_d'}{I_d''} = \frac{|Z''|}{MkR} + \frac{E_b|Z''|}{E_c MkR} \geq 1 \quad (7)$$

which reduces to

$$\frac{MkR}{|Z''|} \cdot \frac{E_c}{E_b + E_c} \leq 1 \quad (8)$$

It will be observed that this criterion is very similar to those developed for the preceding circuits. The factor M will be approximately one-half and its actual value may be determined by measurement of the fundamental component of current to the diode compared with the resultant average current through resistor **28**. In making this measurement, it will be advisable to reduce the voltage E_b from the bias source **40** to zero.

If switch **22** is open while the system is operating in accordance with the above description, the signal frequency voltage appearing across resistor **23** may be used to monitor the operation of the system. Resistor **23** should then have a magnitude as small as is consistent with the desired sensitivity of the monitor.

With switch **21** open, and switch **24** operated to replace source **20** by the double sideband modulation generator **29** or source of double sideband modulated signal, the system may be operated as a detector. In that case a condenser **46** having a low reactance at signal frequencies should be connected in parallel with resistance **28** as by switch **47**. A steady bias will then appear across this condenser and resistance **28**, approximately equal to the carrier value of the voltage appearing across the diode. When this carrier is modulated at signal frequency, an additional current, varying in magnitude at signal frequency, will flow through the diode and impedance **32**. This current, consisting of a series of pulses at carrier frequency, will tend to hold the diode radio-frequency voltage constant through the combined action of impedance **32**, amplifier A and attenuator β. Thus the current through the diode will contain a component whose magnitude is proportional to the modulation on the input signal. If switch **22** is opened, a voltage corresponding to the signal will therefore be developed across resistance **23**.

Under these conditions the device will also operate as a limiter. In other words, the radio-frequency voltage delivered at the output of the radio-frequency amplifier will be held nearly constant, despite variations in the voltage delivered from the source.

Fig. 7 shows a detector circuit **50**, generally similar to that of Fig. 3, in an envelope frequency negative feedback system comprising radio transmitter **51** (including carrier oscillation generating and modulating means) and signal input amplifier **52** for amplifying signals supplied to its input transformer **53** by signal input circuit **54** and transmitting the amplified signals through connection **57** to the modulator in the radio transmitter to modulate the carrier oscillations generated in the radio transmitter. The modulated carrier wave generated by the combination of the carrier oscillations and the signals in the modulator is transmitted from the radio transmitter to a radio-frequency load circuit **55**, for example, an antenna circuit.

The detector circuit **50** is shown with its input connected across the circuit **55** and its output connected across resistor **56** which is in series with the secondary winding of the signal input transformer **53** in the input circuit of the signal amplifier **52**. Modulated waves, from the modulator in the transmitter **51**, are demodulated in detector **50** and the resulting signal waves are fed back through connection **58** to the input of amplifier **52** in phase opposition to signal waves supplied to the amplifier from circuit **54**. Thus, there is negative feedback around the feedback loop including the modulator in the radio transmitter **51**, the detector **50**, and the amplifier **52**. In other words there is feedback around this loop in the proper phase and amplitude to obtain the improvements in distortion, noise, etc., which accrue from the application of negative feedback.

The amount of this negative feedback around this envelope feedback loop may be large, as for example several times 10 decibels. However, in envelope-frequency feedback operation, where a frequency change occurs at the modulator in the mu-circuit of the feedback loop, a limitation on the amount of improvement obtainable (in envelope distortion, etc.) by large amounts of feedback around the loop is encountered due to the presence of a detector or demodulator; because inasmuch as the demodulator is in the beta-circuit of this feedback loop, any distortion generated in the demodulator is present in the transmitter output. Thus the problem of reducing non-linear distortion at the transmitter is not merely a matter of obtaining sufficient envelope feedback around this loop, but involves obtaining a demodulator whose distortion is sufficiently low to realize the improvement of transmitter distortion possible with the increased amount of feedback. Moreover, if a usual type of high level linear rectifier were used, it would need to have a large filament emission with small transmit time and work into a high impedance load, and these requirements would not be compatible with those for the maximum amount of feedback, which are low tube capacitance and low output impedance. However, all these requirements are satisfied by the detector 50 in the system of Fig. 7. This detector is a circuit which includes a linear rectifier, the diode 1, and effectively applies local negative feedback to the diode through linear amplifier A′ in the general fashion described above in connection with Fig. 3, for example. This circuit allows the use of a small rectifier tube with small transit time, low capacitance and only moderate filament emission working into a low output impedance. This feedback makes the rectifier operate as if it had a high output impedance (load impedance) so that large filament emission is not required for small distortion. Though without this feedback a very high physical resistance might be used, it would so increase the time constant that non-tracking in the diode circuit might result. The feedback in the detector may be, for example, several times 10 decibels, and reduces the time constant by a large factor. Thus the fidelity of the detector is improved by local feedback in the demodulator which so modifies the impedance presented to the diode as to maintain its fidelity even when the transmitter is deeply modulated at high envelope frequencies.

In Fig. 7, condensers C_B are blocking condensers. C_C is a blocking condenser for envelope frequencies. Circuit 1′ comprising parallel connected inductance, capacity and resistance is tuned to the radio carrier frequency, f_0, which may be for example of the order of 162 megacycles, the signal being for example a group of 12 carrier telephone messages occupying the frequency band extending from 60 kilocycles to 108 kilocycles. This circuit 1′ acts as a shunt to the signal frequency that is present along with the modulated radio wave on the modulator plate. It helps to insure that the output will be fed back by means of modulation and demodulation and not directly at the signal frequency, its inductance presenting a very low impedance at the envelope frequency so as to minimize any distortion pick-up of envelope voltage from the modulating amplifier. Also, it increases the input impedance of the detector by anti-resonating the tube and the stray capacities.

R_1 and C_1 are the resistance and condenser across which the envelope voltage is developed that is to be amplified by A′. The phase of the envelope voltage developed in the plate circuit of A′ is such as to produce the desired local feedback for the diode, the plate of A′ being coupled to the plate of the diode by connection 60 which includes a blocking condenser C_B and a circuit 3′ tuned to f_0. The circuit 3′ isolates the plate of the diode from the envelope frequency amplifier at radio frequencies. It prevents the radio frequency from getting on the plate of amplifier A′ and helps prevent it from getting through to the grid of the first tube of the signal-frequency amplifier 52. Radio-frequency choke 61 and series-tuned circuit 62 which is tuned to f_0 serve as a filter for the grid of amplifier A′, minimizing the radio-frequency carrier voltage which otherwise might appear on the grid of A′ and cause overloading. Radio-frequency choke 70 further reduces the radio-frequency voltage on the grid of A′. Choke 63, anti-resonant to f_0, further suppresses the radio-frequency voltage to a value which will not cause distortion on the first signal-frequency amplifier grid. Resistor 64 is a grid bias and local negative feedback resistor for A′. Condenser 65 cooperates with resistance R_p to filter the plate current for tube A′. Resistances 66 and 67 form a potential divider for the screen grid circuit of the tube. Elements 65, 66 and 68 filter the screen grid current. Resistance R_2 corresponds to the resistance R_2 in Fig. 3. Network 72 is part of the equalizing network needed in the main feedback loop. It has no significant effect on the operation of the demodulator circuit.

In the case of the detector 50 as well as others of the detectors described above, an advantageous feature is that it is not necessary that the radio-frequency feeding circuit have low impedance for the modulation frequencies.

What is claimed is:

1. A wave translating system comprising a source of signal modulated carrier waves, a detector for demodulating said waves, an amplifier responsive to signal energy from said detector, means for transmitting signal energy from said detector to said amplifier, means connecting said source across a portion of the output circuit of said amplifier, and means for reducing the signal current in said detector comprising means for transmitting signal energy from the output circuit of said amplifier to said detector.

2. A demodulating system comprising a source of signal modulated carrier waves, a detector-amplifier circuit comprising a detector and a signal amplifier, said detector-amplifier circuit having an input circuit and an amplifier output circuit, means for connecting said source across said input circuit, and means for applying signal waves from a portion of said output circuit across said input circuit, the phase shift through said amplifier having such value as to produce negative feedback of signal waves from said output circuit across said input circuit.

3. A demodulating system comprising a source of signal modulated carrier waves, a detector, a signal amplifier having an input circuit and an output circuit, means for connecting said source across said detector and said input circuit in series, and signal transmitting means for connecting a portion of said output circuit across said detector and said input circuit in series, the phase shift through said amplifier having such value as to produce negative feedback of signal waves from its output circuit to its input circuit through said detector.

4. A demodulating system comprising a source of signal modulated carrier waves, a detector, a detector load circuit including an amplifier for

signal waves having an input circuit and an output circuit, a circuit for transmitting signal waves including said detector, said input circuit and a portion of said output circuit in series, and means connecting said source across said portion, said amplifier having its phase shift such that it produces negative feedback of signal waves from its output circuit to its input circuit through said detector and negative biasing voltage for the detector varying in accordance with the signal and tending to reduce signal current flow through the detector resulting from detection of modulated waves supplied to said detector from said source.

5. A wave translating system comprising a source of waves, a rectifier, an amplifier, means transmitting energy from said rectifier to said amplifier, negative feedback means for said amplifier connecting in serial relation the amplifier input circuit, a path through said rectifier and a path forming a portion of the amplifier output circuit, and means connecting said source across one of said paths.

6. A demodulating circuit comprising a source of signal modulated carrier waves, a diode, an amplifier responsive to signal waves, and means for feeding said diode from said source and said amplifier from said diode and producing negative feedback of signal waves in said amplifier, comprising means connecting in serial relation said diode, the input circuit of said amplifier and a branched circuit including in one branch said source and in another branch a portion of the output circuit of said amplifier.

7. A wave translating system comprising a source of signal modulated carrier waves to be demodulated, a diode detector, an amplifier, means for transmitting signal waves from said detector to the input circuit of said amplifier, negative feedback means for said amplifier connecting in serial relation the input circuit of said amplifier, a path through said rectifier and a portion of the output circuit of said amplifier, and means connecting said source across said portion.

8. A wave translating system comprising a source of signal modulated carrier waves, a detector comprising an anode, a cathode and a control electrode connected to said source for demodulating said waves, an amplifier responsive to signal energy from said detector, means transmitting signal energy from said detector to said amplifier, and means for producing negative feedback of signal energy from the output circuit of said amplifier to said control electrode and cathode of said detector.

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