

Aug. 15, 1944.

S. T. MEYERS

2,355,991

STABILIZED NON-LINEAR FEEDBACK SYSTEM

Filed May 16, 1942

5 Sheets-Sheet 1

FIG. 1

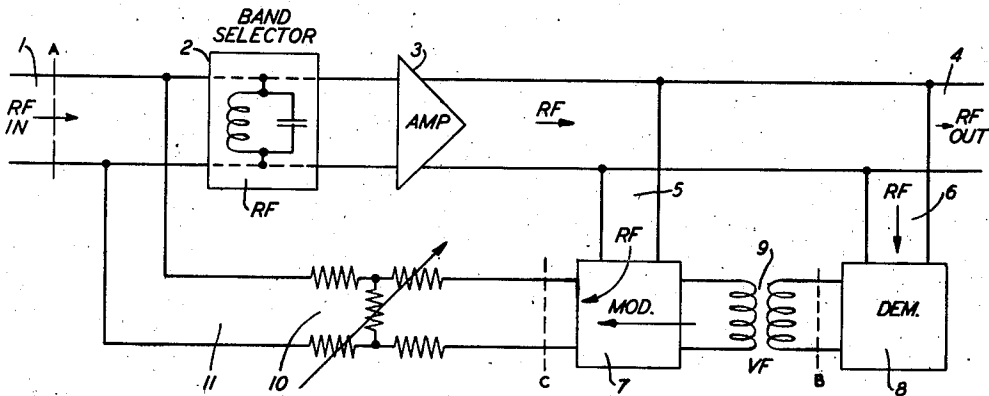
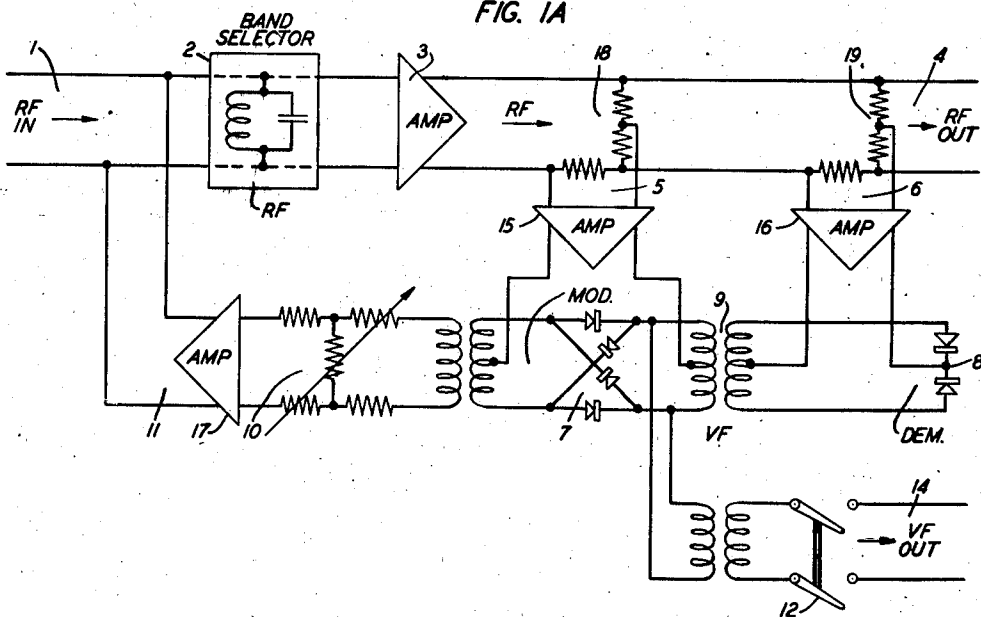


FIG. 1A



INVENTOR
S. T. MEYERS
BY
B. C. Seger
ATTORNEY

Aug. 15, 1944.

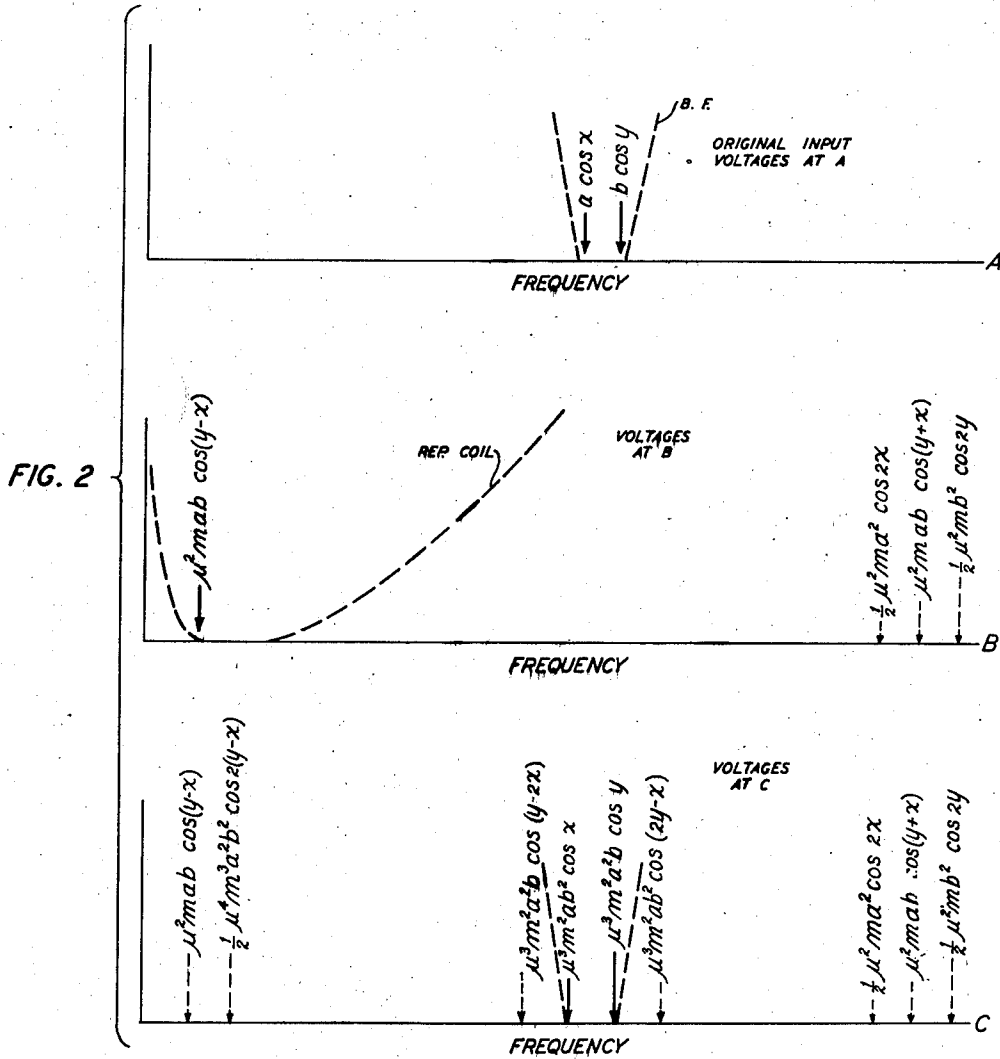
S. T. MEYERS

2,355,991

STABILIZED NON-LINEAR FEEDBACK SYSTEM

Filed May 16, 1942

5 Sheets-Sheet 2



INVENTOR
S. T. MEYERS

BY

B. L. Nege
ATTORNEY

Aug. 15, 1944.

S. T. MEYERS

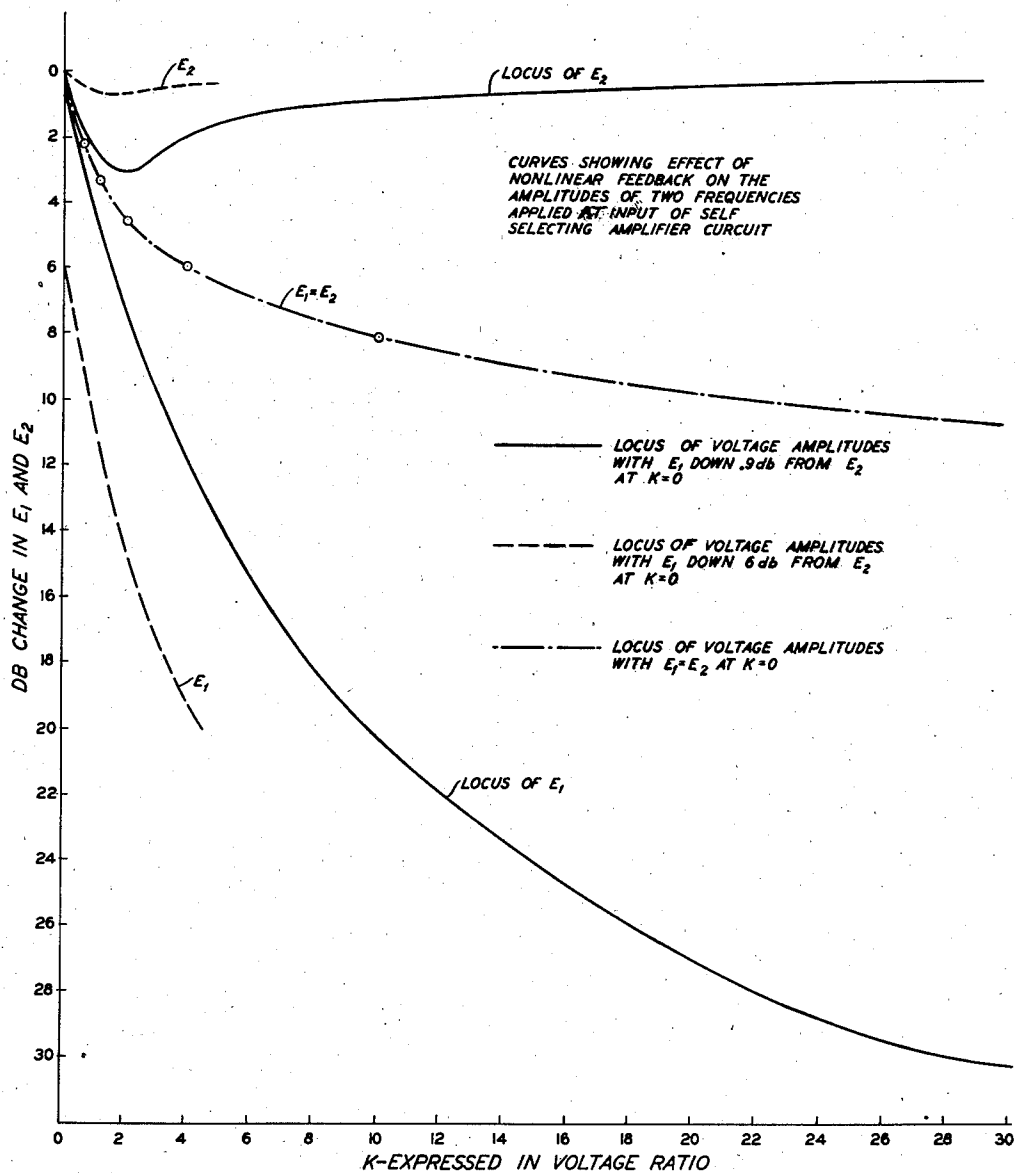
2,355,991

STABILIZED NON-LINEAR FEEDBACK SYSTEM

Filed May 16, 1942

5 Sheets-Sheet 3

FIG. 3



INVENTOR
S. T. MEYERS

BY

B. L. Fager
ATTORNEY

Aug. 15, 1944.

S. T. MEYERS

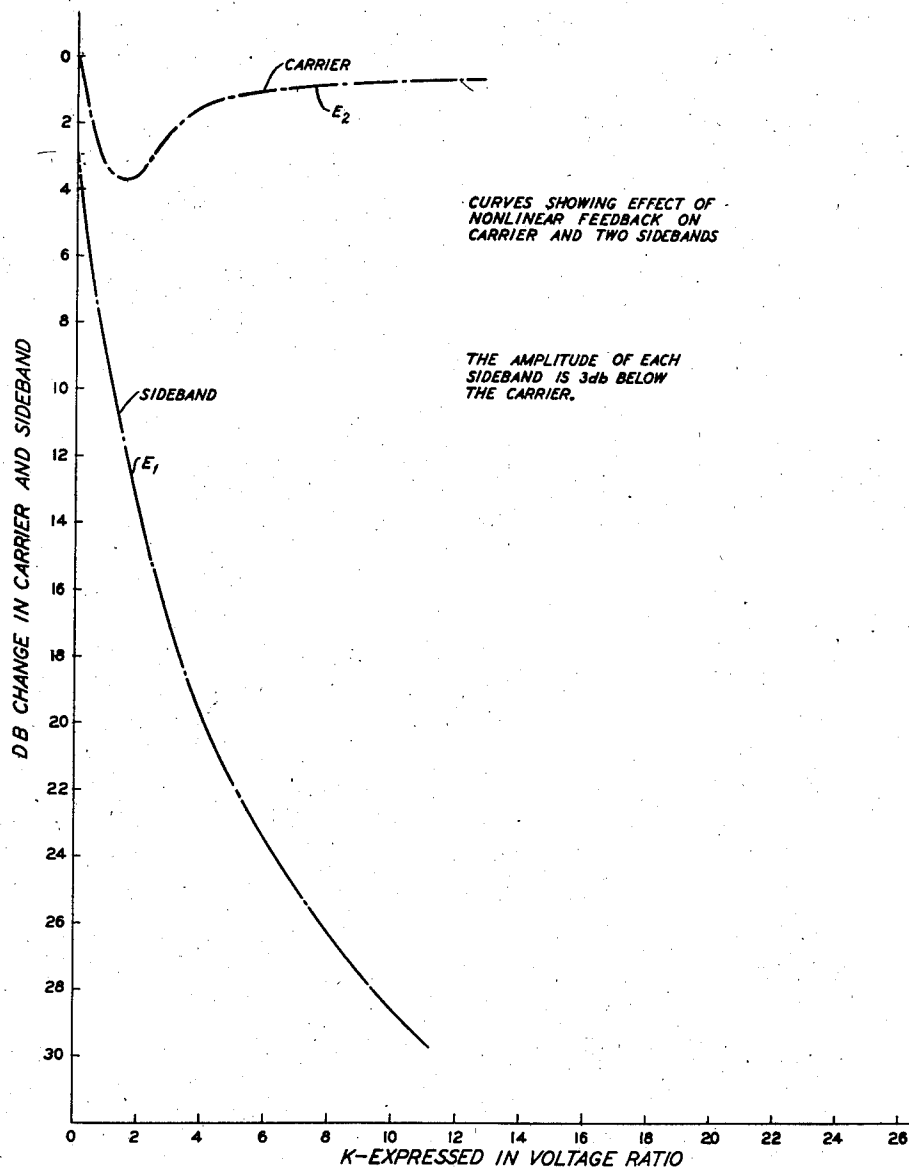
2,355,991

STABILIZED NON-LINEAR FEEDBACK SYSTEM

Filed May 16, 1942

5 Sheets-Sheet 4

FIG. 4



INVENTOR
S. T. MEYERS

BY

B. C. Eger
ATTORNEY

Aug. 15, 1944.

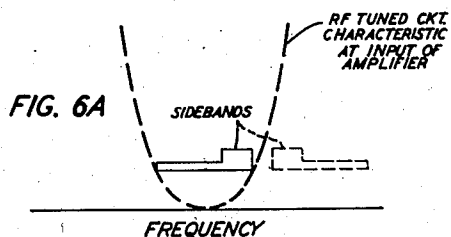
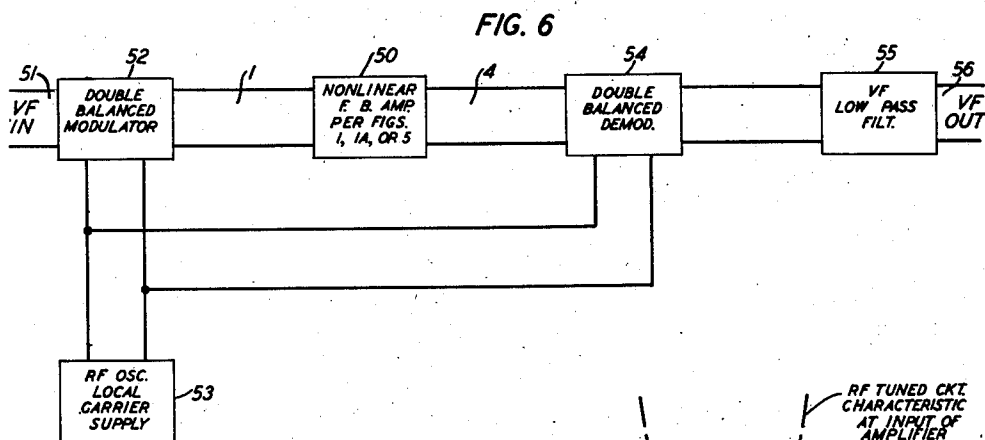
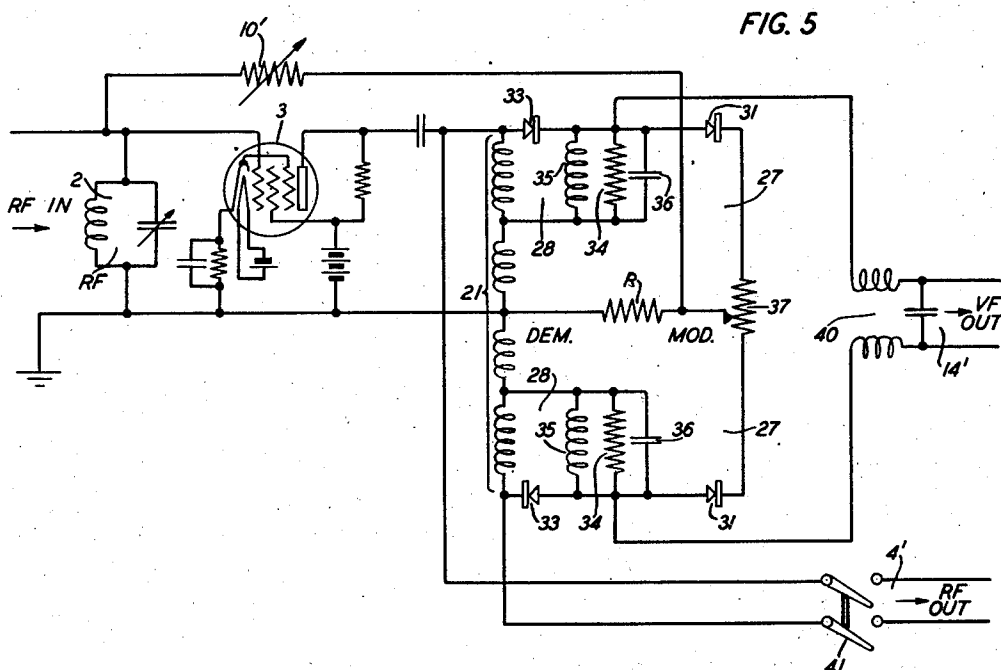
S. T. MEYERS

2,355,991

STABILIZED NON-LINEAR FEEDBACK SYSTEM

Filed May 16, 1942

5 Sheets-Sheet 5



INVENTOR
S. T. MEYERS
BY
B. G. Eger
ATTORNEY

UNITED STATES PATENT OFFICE

2,355,991

STABILIZED NONLINEAR FEEDBACK SYSTEM

Stanley T. Meyers, East Orange, N. J., assignor to
Bell Telephone Laboratories, Incorporated, New
York, N. Y., a corporation of New York

Application May 16, 1942, Serial No. 443,227

5 Claims. (Cl. 178-44)

This invention relates to non-linear feedback systems.

Objects of the invention are amplitude discrimination and automatic tuning, for example, to eliminate disturbing frequencies in detector-amplifier circuits of measuring equipment, or select and reinforce the unmodulated carrier component of a received modulated wave in demodulator circuits.

In one specific aspect the invention may be an amplifier, for example, a voice frequency or carrier frequency amplifier, with non-linear negative feedback reducing the band width of the amplifier. The feedback can narrow the pass band of the amplifier to even a fraction of a cycle, if desired, and can cause the amplifier to select the strongest frequency of a group of frequencies applied at the input (though the frequency of the strongest component vary with time) and amplify it to the exclusion of all the others, thus making the amplifier self-tuning. Such a self-tuning voice frequency amplifier, for instance, is advantageous for example as a narrow band selector in measuring circuits, to eliminate noise interference (which does not completely override the desired signal) and modulation which disturb the accuracy of the measurements. A carrier frequency amplifier, for instance, with the non-linear feedback giving it this property of automatic tuning, is advantageous for example, in radio and carrier systems where at the receiving terminals it is desired to reinforce the carrier by selecting it out, amplifying it and reintroducing it into the circuit, or to increase the carrier amplitude relatively to the side-band amplitude. With sufficient feedback the amplifier can select out the carrier and amplify it without altering the phase of the carrier. In double side-band systems the carrier energy can be, for example, even as small as that of the side-bands and still be selected out to the exclusion of the side-bands, giving distortion-free carrier, i. e., pure carrier or carrier without amplitude modulation.

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

Fig. 1 is a block schematic showing of a self-tuning radio-frequency amplifier embodying a form of the invention;

Fig. 1A is a diagram of a circuit embodying an amplifier of the type of Fig. 1;

Fig. 2 shows the frequency spectrum at various points in the feedback loop of Fig. 1;

Figs. 3 and 4 are curves facilitating explanation of the invention;

Fig. 5 is a diagram of a circuit embodying a radio frequency amplifier in which carrier reinforcement is obtained by non-linear feedback;

Fig. 6 shows a self-tuning or self-selecting amplifier for use in voice frequency circuits; and

Fig. 6A indicates diagrammatically the frequency spectrum at a frequency selective circuit in the amplifier of Fig. 6.

The self-tuning amplifier of Fig. 1 can select the strongest component of a group of frequencies applied to the input and amplify it to the exclusion of the others. For example, when the input contains an unmodulated carrier component and noise or signal side-band energy of different frequency which has no component as great as the carrier component, the amplifier can amplify the carrier to the exclusion of the noise or signal side-band, or can discriminate in favor of the carrier to any desired extent, even though the frequency of the carrier may vary with time. This figure shows incoming line or circuit 1, band selector or frequency selective circuit 2 which may be, for example, a tuned radio frequency circuit, amplifier 3, and outgoing line or circuit 4. Paths 5 and 6 connect the amplifier output circuit to a non-linear modulator 7, and non-linear demodulator 8, each of the square law type, for example. The demodulated frequencies may be voice frequencies for instance, with repeating coil 9 connecting the demodulator to the modulator a voice frequency repeating coil which has poor transmission to direct current and radio frequencies. The combination of the amplifier output and the demodulator output applied to the modulator produce frequencies of various sorts which are returned to the input through path 11 which may include a pad 10. Examples of forms the modulator and the demodulator may have are indicated for instance in Fig. 1A described hereinafter, wherein the demodulator 8 is shown, for instance, as of the balanced type and the modulator 7 is shown, for instance, as of the general type (double-balanced type) disclosed in F. A. Cowan Patent 2,025,158, December 24, 1935, or in Fig. 2C of R. S. Caruthers' paper on "Copper oxide modulators in carrier telephone systems," Bell System Technical Journal, April 1939. The rectifying elements of the modulator and demodulator may be, for example, copper oxide elements or diode or grid controlled vacuum tubes, the desired square law or non-linear characteristics of the modulator and the demodulator being obtained by suitable choice of the copper oxide elements or the vacuum tubes or by association with the rectifying elements of suitable resistance

networks, biasing sources, etc. (not shown). The modulator prevents any voltages applied to it (notably those applied by path 5) from being directly or linearly transmitted to pad 10 and frequency selective circuit 2. However, as brought out below, some of the voltages produced by modulation in the modulator have frequencies identical with those of the original impressed voltages (i. e., the voltages received from line 1), and these modulation products (along with others) are transmitted from the modulator to pad 10 and circuit 2.

Assuming, to simplify picturing operation of the system, that the tuned RF circuit holds such relationship to the frequency of the original impressed voltages that only these frequencies may pass and no others of the frequencies received from the modulator, any spurious frequencies produced in the modulator can be neglected or in other words consideration of the feedback components can be limited to those of the same frequency as the original impressed voltages. Then, applying to the system at point A in Fig. 1 the input

$$a \cos x + b \cos y \quad (1)$$

as shown at A in Fig. 2, the amplifier output without feedback is

$$\mu(a \cos x + b \cos y) \quad (2)$$

where μ is the amplification of the amplifier 3 expressed as a voltage ratio. Without feedback, at point B in Fig. 1 the demodulator output, with square law, is

$$m\mu^2(a \cos x + b \cos y)^2 \quad (3)$$

where m is a factor of proportionality. Of all the modulation products produced by this process those derived from the product term are the only ones that will pass through the VF repeating coil 9. The other products are direct current and second harmonics of x and y and the sum term of $(x+y)$ which will be suppressed, as indicated at B in Fig. 2. The only VF product is

$$\mu^2 m a b \cos (y-x) \quad (4)$$

If now a portion of this wave be added to a portion of the amplifier output wave and both be impressed on the modulator, the resulting modulator output at point C in Fig. 1 (without feedback) is

$$m[\mu(a \cos x + b \cos y) + m\mu^2 ab \cos (y-x)]^2 \quad (5)$$

assuming the same factor of proportionality m for the modulator. By squaring the following terms are produced as shown at C, Fig. 2.

$$\frac{1}{2}\mu^2 m a^2 \cos 2x \quad \mu^3 m^2 a^2 b \cos y \quad (6)$$

$$\frac{1}{2}\mu^2 m b^2 \cos 2y \quad \mu^3 m^2 a b^2 \cos x$$

$$\mu^2 m a b \cos (y+x) \quad \mu^3 m^2 a^2 b \cos (y-2x)$$

$$\mu^2 m a b \cos (y-x) \quad \mu^3 m^2 a b^2 \cos (2y-x)$$

$$\frac{1}{2}\mu^4 m^3 a^2 b^2 \cos 2(y-x)$$

These frequencies will be reduced in amplitude by a ratio n in passing through the pad 10 on the way back to the input. With the band filter at the input having its transmission characteristic as assumed above, all these frequencies but those of x and y are suppressed by it. Then (for this transmission once through the feedback loop)

the coefficients of the fed back components x and y including the ratio n are given by

$$\mu^3 m^2 n a b^2 \cos x + \mu^3 m^2 n a^2 b \cos y \quad (7)$$

Let $K = \mu^3 m^2 n$ for simplicity. Then the coefficients are given by

$$K a b^2 \cos x + K a^2 b \cos y \quad (8)$$

Here the factor by which the original coefficients of x and y must be multiplied to obtain the coefficients of the components that are fed back (in a single trip through the feedback loop) a factor that in linear feedback circuits has been called $\mu\beta$, are

$$K b^2 \text{ for the } \cos x \text{ term} \quad (9)$$

and

$$K a^2 \text{ for the } \cos y \text{ term}$$

Thus, if b is a larger number than a , more feedback may be expected for the $\cos x$ term than for the $\cos y$ term. Considering now, steady state propagation with the feedback loop closed, relationships governing the final coefficients of x and y (i. e., the coefficients of x and y with feedback), may be set up in much the same manner as is done in linear feedback circuits. Designating final coefficients of $\cos x$ and $\cos y$ as E_1 and E_2 , respectively, in one trip around the feedback loop these coefficients, as in the case of a and b above, become

$$E_1 (K E_2^2) \text{ for the } \cos x \text{ term} \quad (10)$$

and

$$E_2 (K E_1^2) \text{ for the } \cos y \text{ term}$$

These are the final fed back voltages (with the feedback loop closed). Assuming negative feedback the final voltages at the input are equal to the initial impressed voltages minus the final fed back components; thus,

$$E_1 = a - E_1 (K E_2^2) \text{ and } E_2 = b - E_2 (K E_1^2) \quad (11)$$

$$= \frac{a}{1 + K E_2^2} = \frac{b}{1 + K E_1^2}$$

where a and b are the initial coefficients of the $\cos x$ and $\cos y$ respectively. Substituting the values of E_1 and E_2 from (11) in the Equations 11 for E_2 and E_1 , respectively, we have

$$E_1 = \frac{a}{1 + K \left[\frac{b}{1 + K E_1^2} \right]^2} \text{ and } E_2 = \frac{b}{1 + K \left[\frac{a}{1 + K E_2^2} \right]^2} \quad (12)$$

whence E_1 is given by

$$K^2 E_1^5 - a K^2 E_1^4 + 2 K E_1^3 - 2 a K E_1^2 + E_1 (1 + K b^2) - a = 0 \quad (13)$$

and E_2 is given by

$$K E_2^5 - b K^2 E_2^4 + 2 K E_2^3 - 2 b K E_2^2 + E_2 (1 + K a^2) - b = 0 \quad (14)$$

Recourse to general solution of these 5th order equations is unnecessary. A few samples in the substitution of numerical values to obtain particular solutions by the cut and try method readily serves to show the trends of E_1 and E_2 as a function of a and b and the factor K , which may be referred to as the feedback factor but should not be confused with the β nor the $\mu\beta$ of a linear feedback system whose loop propagation is $\mu\beta$. The solid lines in Fig. 3 show the effect of the feedback on E_1 and E_2 when the initial voltage a is down .9 decibel from b . As K proceeds from zero both E_1 and E_2 go down alike at first. Then on reaching a minimum E_2 reverses its trend and gradually returns to

its initial value from below. E_1 at the same time becomes smaller and smaller as K approaches infinity. The above equations, then, become

$$E_1 \rightarrow \frac{a}{Kb^2} \text{ and } E_2 \rightarrow b \text{ as } K \rightarrow \infty$$

Though these equations have other roots the roots plotted are the only real roots that form a smooth characteristic from $K=0$ to $K=\infty$. The dotted line of Fig. 3 shows what happens when a is 6 decibels below b . It will be noted that for small values of feedback the influence of E_1 on E_2 is much smaller than in the previous case shown by the solid line. If a is closer to the value of b than that shown by the solid line, the dip in the characteristic of E_2 can be expected to be greater and the minimum point can be expected to move toward the larger values of feedback. When a is equal to b (which is the limit for beyond this E_1 and E_2 will merely switch places) E_1 and E_2 will be reduced equally by any feedback that is applied to the system. This is shown by the dot-dash curve in Fig. 3.

Effects upon operation of the system obtained by allowing feedback of modulation products whose frequencies differ from those of the voltages initially impressed on the system, can be pictured in a simple way by considering operation of the system with feedback allowed of not only modulation products having the same frequencies as the impressed waves but simultaneously another modulation product, for example, one of the nearby products of frequency $(y-2x)$ or $(2y-x)$ shown at C in Fig. 2.

For instance, if the pass band of the band filter or frequency selective circuit 2 at the input is widened by moving its lower cut-off down to include $(y-2x)$ then three final voltages E_1 , E_2 and E_3 must be considered. The newly added voltage E_3 is the final coefficient of the $(y-2x)$ term. Amplifying and demodulating as previously indicated and considering only voice frequencies passing through the repeating coil at the demodulator output, and modulating as previously indicated and considering only the frequencies that pass through the band filter, the following equations can be derived for E_1 , E_2 and E_3 :

$$E_1 = \frac{a}{1 + K(E_2 + E_3)^2} \quad (15)$$

$$E_2 = \frac{b - KE_1^2 E_3}{1 + K(E_1^2 + E_3^2)}$$

$$E_3 = -\frac{KE_1^2 E_2}{1 + K(E_1^2 + E_2^2)}$$

On inspection these equations tell very little. Each term depends on coefficients of the other two terms. However, not knowing what the final values of E_1 , E_2 and E_3 are it is permissible to assume values and see if their substitution in the equations produce identities. If identities result the assumptions have been correct. Assume that

$$E_1 \rightarrow \frac{a}{Kb^2} \text{ and } E_2 \rightarrow b \text{ as } K \rightarrow \infty \quad (15A)$$

This is equivalent to assuming that the properties of the circuit are the same as with the narrower band. Substituting these values of E_1 and E_2 in the equation for E_3 we have

$$E_3 = \frac{a^2}{b} \frac{1}{K^2} \quad (16)$$

This shows E_3 goes down as the square of K which is faster than the rate of reduction of E_1 which only goes down directly as K . In the equations of E_1 and E_2 , therefore, where E_3 is added to E_1 and E_2 , E_3 will have second order effect and can be neglected. Applying this to the equations for E_1 and E_2 we obtain

$$E_1 = \frac{a}{1 + KE_2^2} \text{ and } E_2 = \frac{b}{1 + KE_1^2} \quad (17)$$

The factor $KE_1^2 E_3$ which appeared in E_2 is neglected because it approaches zero. From these two equations are obtained the values of E_1 and E_2 as $K \rightarrow \infty$ that were assumed above. The identities have been established proving the correctness of the initial assumptions. These are not the only roots but it is reasonable, using Equations 12 as a guide along with the curves of Fig. 3 which show only one positive real root for E_1 and E_2 at large values of K , to consider the above roots likewise the only positive real roots for E_1 and E_2 . Thus it appears that for large values of feedback the automatic selecting properties of this circuit are unchanged by the inclusion of the $(y-2x)$ term in the feedback.

Suppose now that the band filter at the input instead of being set to include $(y-2x)$ is shifted upward a little to exclude $(y-2x)$ but include $(2y-x)$. Going through the demodulating and modulating processes as before but calling E_3 the coefficient of the $(2y-x)$ term a similar set of equations can be derived which are:

$$E_1 = \frac{a - KE_2^2 E_3}{1 + K(E_2^2 + E_3^2)} \quad (18)$$

$$E_2 = \frac{b}{1 + K(E_1 + E_3)^2}$$

$$E_3 = \frac{-KE_1 E_2^2}{1 + K(E_1^2 + E_2^2)}$$

This time, however, identities are not established by making the same guesses as previously. To satisfy all these equations

$$-E_3 \rightarrow E_1 \rightarrow \left(\frac{a}{2K}\right)^{1/3} \quad (18A)$$

and

$$E_2 \rightarrow b, \text{ all as } K \rightarrow \infty$$

As in the previous case the other roots appear not of interest. Notwithstanding the fact that reduction in E_1 , (which is equal to $-E_3$ when K is large), may be retarded by the presence of the $(2y-x)$ term, with sufficient non-linear feedback the unwanted frequencies can be reduced without affecting the final amplitude of the strongest frequency.

However, the pass band of circuit 2 preferably is sufficiently narrow to suppress second harmonics of the frequencies initially impressed on the system at the input.

An important property of this circuit is the effect of its non-linear feedback on carrier and two side-bands. Suppose at the input of Fig. 1 a frequency, y , and upper and lower side-bands $(y+v)$ and $(y-v)$ are impressed, the pass band of the filter 2 extending from $(y-v)$ to $(y+v)$. The amplitudes of the side-bands are identical. This input is:

$$E_1 \cos(y-v) + E_2 \cos y + E_1 \cos(y+v)$$

Going through the demodulating and modulating processes as previously using E_2 as the final value of $\cos y$ and E_1 as the final value of $\cos(y-v)$ and of $\cos(y+v)$, the following final

expression for the amplitudes due to feedback can be derived:

$$E_1 = a - KE_1(E_1^2 + 2E_2^2) \quad (19)$$

$$= \frac{a}{1 + K(E_1^2 + 2E_2^2)}$$

and

$$E_2 = \frac{b}{1 + 4KE_1^2}$$

By substituting specific values in these equations the trends of E_1 and E_2 can be readily plotted. Fig. 4 shows how E_1 and E_2 the side-band and carrier amplitudes, respectively, vary with feedback. The curve shows the limiting case where the carrier is 3 decibels above one side-band. This represents more than 100 per cent modulation. It is seen that, as feedback increases, carrier and side-bands go down alike for small values of feedback. Then as the feedback gets steadily larger the carrier reduction strikes a minimum and then reverses its trend and gradually returns to its initial value. The side-bands, however, keep on decreasing, the reduction ratio approaching $1/K$ for the larger values of feedback. Thus, by the proper amount of non-linear negative feedback any excess of carrier over side-bands may be produced. Distortion products of the side-bands are small in this feedback process. This may be inferred from the equations involving the third order distortion $(y-2x)$ and $(2y-x)$ discussed above. If, in these equations, x be considered a lower side-band and y the carrier, it will be noticed that $(y-2x)$ is a distortion product of (x) and $(2y-x)$ is an upper side-band, not a distortion product. Since the $(y-2x)$ term goes down by the factor K^2 while the $(2y-x)$ term goes down by the factor $K^{1/3}$, the distortion is reduced faster than the side-band. A similar situation is presented where x is an upper side-band instead of a lower, because it produces a similar set of frequencies symmetrical about the carrier. When, however, we have an x (or in other words a side-band) equal distances above and below the carrier, one set of modulation products, i. e., the $(y-2x)$ product and its counterpart which is symmetrically located above the carrier, stands out as distortion and is reduced according to K^2 , while the other set, i. e., the $(2y-x)$ product and its counterpart which is symmetrically located below the carrier, overlaps the side-bands and is counted as side-bands. So, since the reduction factor for the side-bands as shown by Equation 19 is K , the distortion with respect to side-band goes down by the factor $K^2/K=K$.

Thus, in double side-band carrier systems the carrier can be reinforced, i. e., increased in amplitude relatively to the side-bands, by applying non-linear negative feedback to the carrier or radio frequency amplifier, for example as in Fig. 1, so as to reduce the side-band amplitude but only to the point where the desired excess of carrier is reached. The amplifier output can then go to a detector and with a sufficient amount of non-linear negative feedback in the amplifier the demodulated wave is substantially distortionless, because of the high amplitude of carrier relative to side-band and also because the (non-linear) negative feedback of the side-bands flattens the frequency response of the amplifier for the side-bands (much as negative feedback flattens frequency response of an amplifier in which the negative feedback is linear).

As to the frequency characteristic of an amplifier of the type shown in Fig. 1, the pass band of the amplifier, or band over which the gain to the weak frequencies is not reduced, or is least reduced, by the non-linear negative feedback, extends symmetrically about the selected frequency, the selected frequency being as indicated above, the strongest frequency of a group applied at the input. Referring to Fig. 1 it may be seen that the pass band of the amplifier covers those frequencies at which there is little or no feedback (as regards the non-linear negative feedback through 10). In other words, within the free transmission range of circuit 2, the amplifier gain at any instant varies with frequency inversely as the amount of the non-linear negative feedback. The amount of the feedback at the radio frequencies, and so the extent of the pass band of the amplifier, is controlled by the VF branch of the feedback loop. No feedback, and so no gain reduction by feedback, occurs in the bands between the selected frequency and the sum or difference of the selected frequency and the lowest frequency to get through the VF branch. Therefore, the lower the bottom of the pass band of the VF branch the narrower the pass band of the amplifier to weak frequencies, i. e., the lower the low end cut-off in the VF branch the narrower the pass band of the amplifier. Conversely the higher the low end cut-off of the VF branch the wider the pass band of the amplifier.

When the input to Fig. 1 is a carrier with a side-band or with both upper and lower side-bands, the amplified output, in case the amount of the non-linear negative feedback has been made sufficiently great to substantially completely suppress the side-band energy, is an unmodulated wave of the same frequency (from instant to instant) as the unmodulated carrier component of the incoming modulated wave, and may be used for any desired purpose for which such a wave is suitable. For example, it can be fed to a demodulator (not shown) for detecting waves from the circuit 1, in order to reinforce the carrier at the detector so as to reduce the distortion produced in the detection of those waves. In case the amount of the non-linear negative feedback in Fig. 1 has been limited sufficiently to preserve in the amplified output an amount of side-band energy adequate for signal reception, this amplified output may be fed to a detector (not shown) in which the distortion produced in the detection will be reduced by the amplitude increase of unmodulated carrier component, relative to side-band, obtained by the circuit of Fig. 1.

Fig. 1A shows an amplifier of the type shown in Fig. 1 but with switch 12 which can be closed to employ, instead of the detector last mentioned, the detector 8. The voice frequency signal detected by demodulator 8 is then delivered to outgoing voice frequency line or circuit 14. This detection in the demodulator 8 is rendered substantially distortionless by reason of the amplitude increase of unmodulated carrier component, relative to side-band, obtained in the output of amplifier 3 by the non-linear negative feedback which takes place in Fig. 1A in the same general manner as described above in connection with Fig. 1. If desired, amplifiers 15, 16 and 17 or other unilateral devices may be included in paths 5, 6 and 11. These amplifiers can adjust the amplification or transmission of these paths as desired and restrict energy flow

through these paths to the directions indicated by the arrows in Fig. 1. With or without these amplifiers, bridges 18 and 19 may be employed in the output circuit of amplifier 3, if desired. Conjugacy provided by bridge 18 prevents modulator 7 from delivering modulation products directly through path 5 to path 6 or outgoing line 4, and prevents path 6 and line 4 from affecting the modulator; and conjugacy provided by bridge 19 prevents demodulator 8 from delivering modulation products directly through path 6 to outgoing line 4 and prevents line 4 from affecting the demodulator.

Fig. 5 shows schematically a radio frequency amplifier and detector circuit (useful for example for amplifying unmodulated waves with reduction of noise and interference, or removing amplitude modulation from or amplifying and detecting single or double side-band modulated carrier waves) in which automatic tuning and carrier reinforcement can be obtained by non-linear feedback as in Fig. 1A. In receiving modulated carrier waves, for instance, tuned RF circuit 2 at the input selects out the desired carrier and side-bands and the vacuum tube circuit 3 amplifies them. Its output circuit contains a radio frequency autotransformer 21 to which are connected rectifiers to form the square law or non-linear balanced demodulator 28 and modulator 27. The demodulator rectifiers are designated 33. The modulator rectifiers are designated 31. Across the load resistor 34 for each half of the demodulator is bridged a retard coil 35 which will pass voice frequencies and bypass direct current and a condenser 36 which will pass voice frequencies and by-pass radio frequencies. The voice frequencies, then are combined with part of the original carrier wave from the autotransformer and applied to the modulator. The output of the modulator is derived across resistance R and is fed back through a feedback resistance 10' to the input. If the demodulator and modulator are well balanced the output across R will be conjugate to the input applied at the autotransformer and no linear transmission will result. To insure a good balance, potentiometer 37 in the modulator circuit may be adjusted for an optimum balance. In this way only modulation products are fed back, any direct transmission to the applied signal being suppressed. However, if the potentiometer is unbalanced some direct or linear feedback to the applied signal is obtained. Such linear feedback may be made positive or negative depending on the way the potentiometer is unbalanced. A small amount of positive linear feedback (less than unity) can serve to build up a large carrier while the non-linear feedback (feedback to modulation products), being negative feedback (and preferably much greater than unity), serves to reduce the side-bands and flatten out the frequency response much the same as linear negative feedback flattens out the frequency characteristic of an ordinary negative feedback amplifier. In this way the circuit can build up a large carrier excess over side-band amplitudes without impairing the quality of the side-bands, and thus increase selectivity. If negative linear feedback is desired the potentiometer may be unbalanced in the opposite direction. If desired, a filter 40 for suppressing radio frequencies may be included in the outgoing voice frequency circuit 14', which can feed an audio-frequency amplifier, for example (not shown).

Closing switch 41 transmits to outgoing radio

frequency circuit 4' radio frequency output from the amplifier 3. This output may include adequate side-band energy for detection of the signal, and be fed from circuit 4' to a detector (not shown); or, if the radio frequency output of the amplifier, via line 4', is to be substantially free of amplitude modulation, this condition may be obtained for example by adjusting feedback resistance 10' to give sufficiently large non-linear negative feedback, and if desired also adjusting the contact of potentiometer 37 to give positive linear feedback of the unmodulated carrier component.

Fig. 6 shows how an amplifier 50 such as the circuit of Fig. 1, Fig. 1A or Fig. 5 may be employed to make a self-selecting amplifier circuit for voice or audio frequencies. The voice frequency band from incoming circuit 51 is translated to some high frequency (such as a radio frequency) by means of the double-balanced linear modulator 52 at the input. The local carrier supplied to the modulator by oscillator or source 53 is suppressed. The side-bands are so located by the frequency translation that one of them lies within the pass band of the tuned circuit 2 at the input of the amplifier as indicated diagrammatically in Fig. 6A. The carrier being suppressed the amplifier can only select the strongest frequency in the side-bands. In the case of Fig. 6A the amplifier will only select a frequency in the lower side-band because any duplicate frequency in the upper side-band will be smaller in magnitude due to the loss characteristic of the tuned circuit 2. The strongest frequency having been selected and amplified it is demodulated down to voice frequency level by the demodulator 54 at the output, preferably a double-balanced linear demodulator. A voice frequency low-pass filter 55 then serves to clean out or suppress from outgoing circuit 56 high frequency modulation products.

In the claims, the terms "non-linear modulator" and "non-linear demodulator," respectively, refer to a modulator and a demodulator which functions as a non-linear device, that is, a device whose output is non-linearly proportional to its input throughout the useful range of the device. For example, a square law modulator or demodulator is a modulator or demodulator whose output is proportional to the square of its input, as distinguished, for instance, from the modulator and the demodulator of Wise Patent 2,224,580, which function as linear devices, i. e., as devices whose output is directly proportional to their input.

What is claimed is:

1. A wave translating system comprising an amplifier having a negative feedback circuit including a non-linear modulator with means for balancing the modulator to prevent direct transmission through said feedback circuit and a frequency selective input circuit for said amplifier connected to receive modulation products of said modulator, said system further comprising a non-linear demodulator external to said feedback circuit, means for feeding the demodulator with amplified waves from the amplifier, and means for feeding the modulator with demodulation products of the demodulator.

2. A wave translating system comprising amplifying means having a frequency selective input circuit and an output circuit, a source for supplying to said input circuit waves of different frequencies and amplitudes, a utilization circuit for a wave of one of said frequencies coupled to

said output circuit, a square law modulator, a square law demodulator, means for supplying modulation products of said modulator to said frequency selective input circuit, means supplying from said output circuit of said amplifying means to said modulator waves selected by said frequency selective input circuit, said modulator being balanced to prevent direct transmission of said waves through said modulator to said frequency selective input circuit, means supplying from said amplifying means to said demodulator waves selected by said frequency selective input circuit, and frequency selective means supplying demodulated waves from said demodulator to said modulator.

3. A wave translating system comprising amplifying means having a frequency selective input circuit, a source for supplying to said input circuit two waves differing in frequency and in amplitude, a double balanced non-linear modulator, a non-linear demodulator, separate paths respectively feeding the modulator and the demodulator with waves amplified by said amplifying means, means for supplying from the demodulator to the modulator a demodulation product of frequency equal to the difference between the frequencies of said two waves to the exclusion of other demodulation products of said waves of said two frequencies, and means for supplying from the modulator to said frequency selective input circuit of said amplifying means modulation components whose frequencies are respectively equal to said two frequencies, said frequency selective input circuit passing said two frequencies to the substantial exclusion of the frequencies of the other modulation products of the three waves constituted by the waves of said two frequencies and their difference frequency.

4. A self-selecting amplifier circuit for audio frequencies, comprising a carrier frequency wave amplifying system, a double balanced modulator, a demodulator, means for supplying to the signal input branch of the modulator a wave including two audio frequency components of different amplitudes whose frequency separation exceeds an octave, means for supplying a carrier wave of frequency of higher order of magnitude than the frequencies of said components to the carrier input branch of the modulator, means for connecting the side-band output branch of the modula-

tor to the input of the carrier frequency wave amplifying system, a frequency selective circuit of compass less than an octave included in the input circuit of the carrier frequency wave amplifying system for selecting one side-band delivered thereto by the side-band output branch of the modulator to the exclusion of the other side-band, means for connecting the output of the carrier frequency wave amplifying system to the demodulator, and means for supplying a demodulating carrier wave to the demodulator, said carrier frequency wave amplifying system comprising an amplifying device having a negative feedback circuit including a non-linear modulator with means for balancing the non-linear modulator to prevent direct transmission through said feedback circuit, and said carrier frequency wave amplifying system further comprising a non-linear demodulator external to said feedback circuit, means for feeding said non-linear demodulator with amplified waves from said amplifying device, and means for feeding said non-linear modulator with a demodulation product of said non-linear demodulator.

5. A wave translating system comprising a source for supplying a modulated wave including an unmodulated carrier component and upper and lower side-bands thereof, an amplifier connected to said source having a frequency selective input circuit for selecting said modulated wave, a square law duplex modulator having its output circuit conjugate to its input circuit and having means operable for adjusting balance of the modulator, a square law demodulator for demodulating said modulated wave having a frequency selective output circuit for selecting the demodulated side-bands, means connecting the amplifier output circuit to the demodulator input circuit, means connecting the demodulator output circuit and a portion of the amplifier output circuit in series with each other in the modulator input circuit, means including a circuit connecting the modulator output circuit to said frequency selective amplifier input circuit for producing negative feedback of the side-bands, and a load circuit connected to the demodulator output circuit for receiving the demodulated side-bands.

STANLEY T. MEYERS.