Nov. 16, 1948. G. J. C. ANDRESEN 2,453,958

SIGNAL AMPLITUDE LIMITING SYSTEM

Filed July 20, 1946 3 Sheets-Sheet 1

FIG. 1

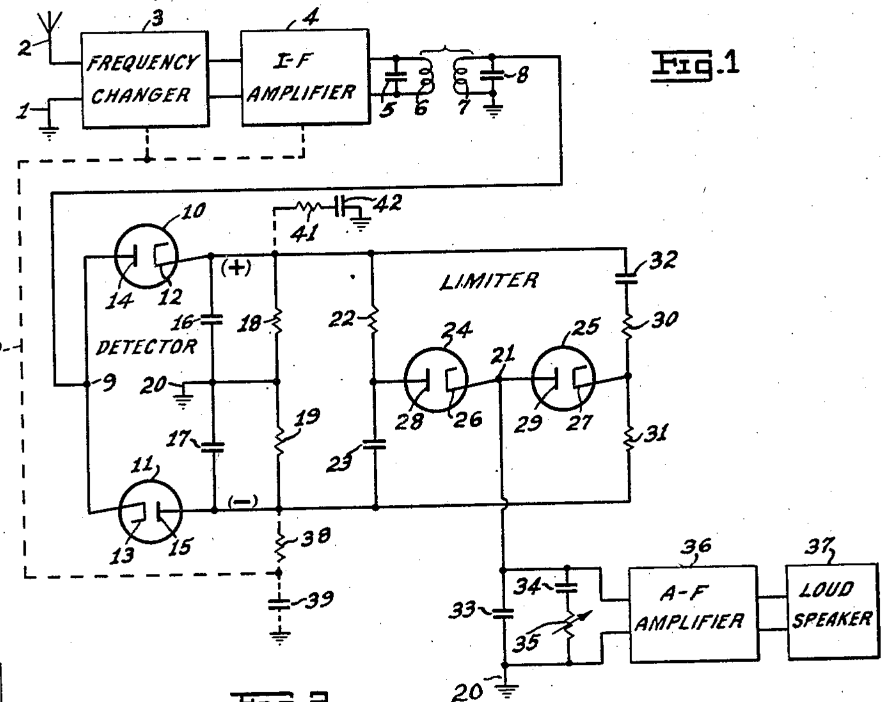

FIG. 2

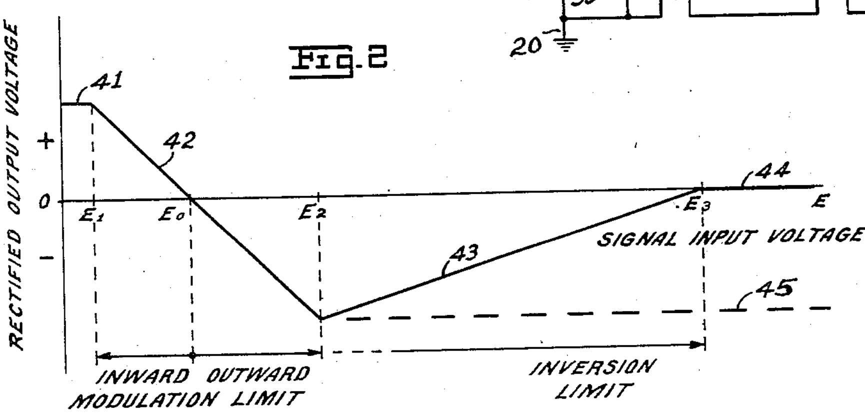

FIG. 5

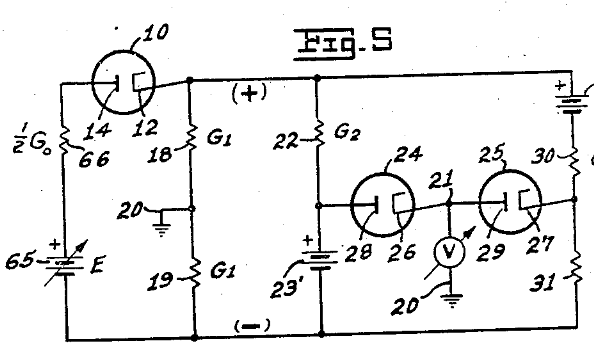

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FIG. 3

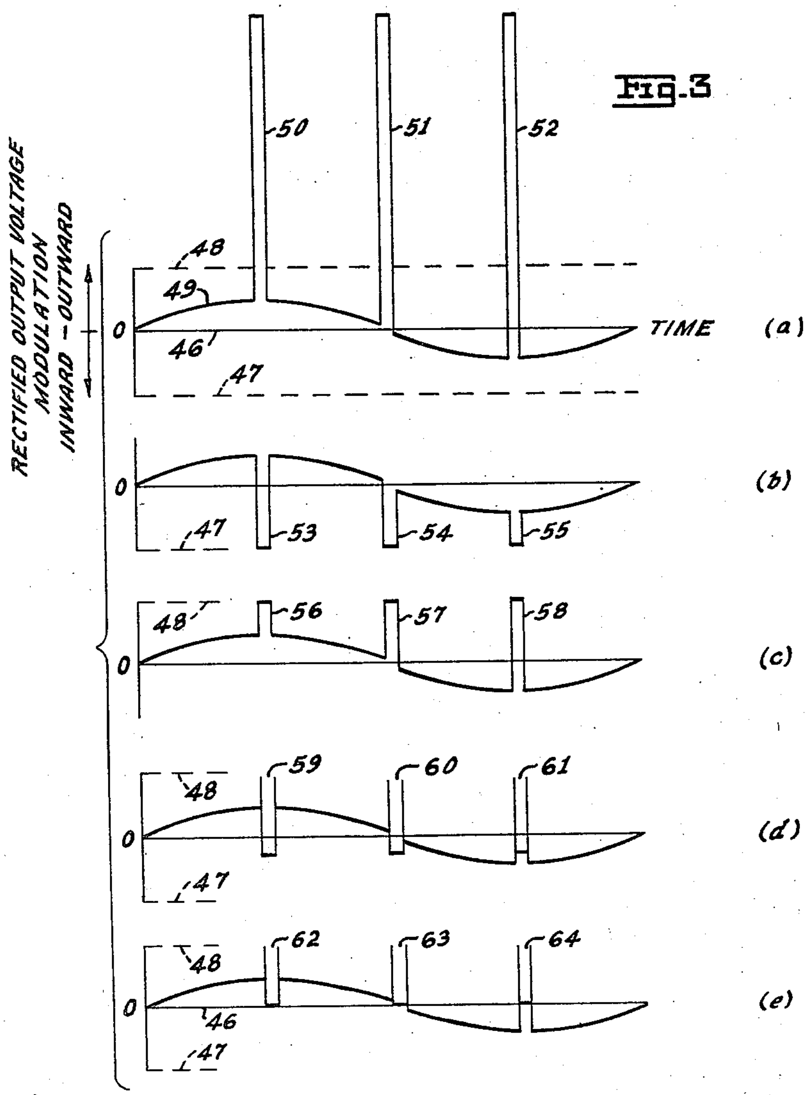

FIG. 4

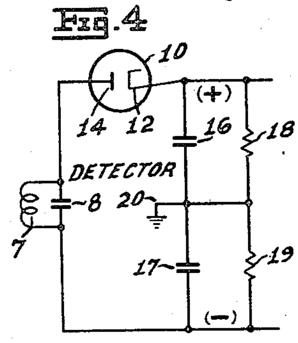

FIG. 6

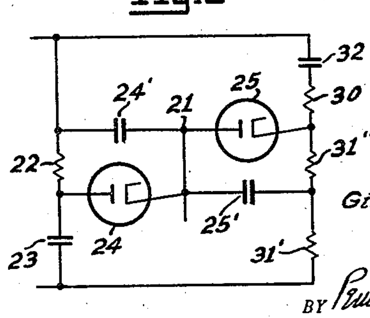

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Fig. 7

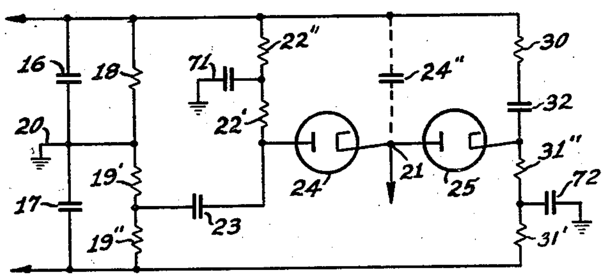


Fig. 8

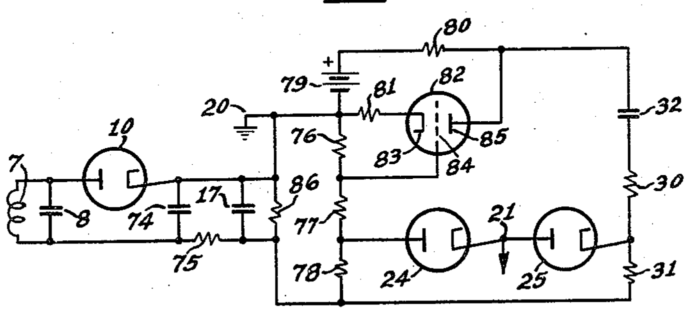
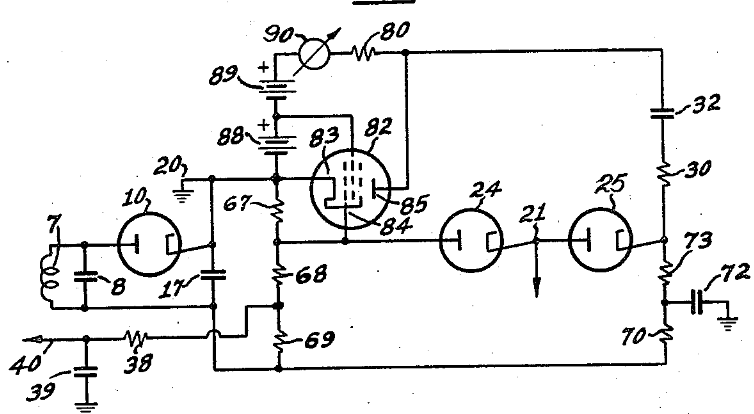


Fig. 9



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UNITED STATES PATENT OFFICE

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SIGNAL AMPLITUDE LIMITING SYSTEM

Gilbert J. C. Andresen, Stow, Ohio

Application July 20, 1946, Serial No. 685,111

9 Claims. (Cl. 178—44)

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This invention is directed to improvements in signal amplitude-limiting systems and especially to such systems adapted for limiting the peaks of noise pulses superimposed on the fluctuations of a desired signal. A particular application is in association with the detector in a radio receiver, in which case the rectified carrier from the detector may be utilized to control automatically the limiting level so it is maintained in a desired relation to the maximum amplitude of modulation of the carrier.

The present invention in certain respects is an improvement of the automatic limiter systems shown in my Patent 2,418,389, issued April 1, 1947, and my pending application Serial No. 738,165, and makes use of features disclosed and claimed therein.

Certain types of noise interference or static appear in a radio receiver as pulses of modulation superimposed on the desired signal modulation. These pulses are characterized by large amplitude and short duration. The amplitude may be many times the carrier amplitude. The duration is comparable with the period of the highest frequency of modulation which can be reproduced by the amplifier and detector and therefore much less than the period of the frequencies which appear in the modulation with greatest amplitude.

Many schemes have been proposed for minimizing the detrimental effect of noise pulses in a receiver. Many of these schemes rely on limiters of various kinds to cut off the peaks of the pulses insofar as they exceed the level of greatest modulation. There are two classes distinguished by whether the pulse is limited at the level of maximum outward modulation or that of maximum inward modulation. A representative circuit in the former class is a diode limiter adapted to follow the detector in a radio receiver, and including a diode rectifier as a limiter whose level is adjusted automatically to twice the carrier level in order to protect the greatest modulation. A representative circuit of the second class cuts off the signal for the duration of the pulse, which effectively sets the peak value of the pulse at the level of maximum inward modulation or zero amplitude. In either of these classes, the peak amplitude of the residual noise pulse, measured from the level of zero modulation, is equal to the maximum amplitude of the modulation to be reproduced. While devices such as shunt capacitors or low-pass filters may be used in some cases for further reduction of the pulse peak amplitude, it still remains greater than the weaker amplitudes of signal modulation.

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The object of the present invention is the reduction of the peak amplitude of noise pulses to a value much less than the maximum amplitude of the signal to be reproduced in such a way that the limiting level automatically adjusts itself in proportion to the strength of signal being received. This is accomplished by an improved circuit arrangement which, like the circuits of my above-mentioned patent and pending application, inverts the peak insofar as it exceeds a limiting level near the maximum amplitude of the signal and may, if desired, include means for cutting off the inverted peak at the average value of the signal, so the disturbing effect of a strong pulse is minimized. In a modulated-carrier signal receiver, the latter level corresponds to the unmodulated condition.

As in the systems of my above-mentioned patent and pending application, the present invention provides both direct and inverse circuits coupling the input circuit to the output circuit and the direct coupling circuit includes a rectifier for disabling this circuit at amplitudes exceeding a certain cutoff level. Also the cutoff level of the limiter is adjusted automatically to correspond with the maximum modulation intended to be reproduced free of distortion; this is accomplished in the present invention by a simple arrangement of properly proportioned capacitors and resistors free of additional tubes or critical adjustments. In the direct coupling circuit and also in the inverse coupling circuit the time constant of the capacitance with its associated resistance is greater than the greatest period of signal fluctuations to be reproduced, so the voltage of the charge on each capacitance is determined by the average signal amplitude. By this arrangement the peak of any disturbing pulse exceeding the cutoff level is inverted and the cut-off level is automatically adjusted to the strength of signal being received. More specific applications and further features will appear in the following description of representative circuits and their operation.

In the drawings, Fig. 1 shows a radio receiver embodying the invention, Fig. 2 shows the limiting characteristics of the invention, Fig. 3 shows the effect on noise pulses achieved by this invention as compared with other limiters, Fig. 4 shows an alternative arrangement of the detector associated with the invention, Fig. 5 shows a direct-current circuit for demonstrating the limiting action of the invention, Fig. 6

shows an arrangement for neutralizing the effects of incidental capacitance in the circuits of this invention, Fig. 7 shows a modification of the limiting circuits of Fig. 1, Fig. 8 shows an alternative arrangement including a reversing repeater, and Fig. 9 shows an arrangement including the features of both Fig. 7 and Fig. 8.

Referring to Fig. 1, the invention is shown as embodied in a superheterodyne radio receiver which is otherwise of conventional design. The received modulated-carrier signal is picked up by the ground connection 1 and antenna connection 2. In the frequency changer 3 the carrier frequency is changed to an intermediate frequency, at which it is amplified in the intermediate-frequency amplifier 4. Whatever tuning and selectivity are required may be provided in these sections 3 and 4. The intermediate-frequency output circuits are shown as a double-tuned inductively coupled transformer comprising the primary capacitor 5 and inductor 6, and the secondary inductor 7 and capacitor 8. The detector and the limiter circuits of this invention are shown between terminals 9 and 21, to be described below. The shunt capacitor 33, shown connected between the limiter output terminal 21 and the common ground terminal 20, is an aid in smoothing out and reducing the peak amplitude of residual very short pulses which may remain after the limiting action. This capacitor preferably has the largest value that is permissible while preserving the reproduction of the signal components at the higher audio frequencies. Since this requirement varies with the types of signals and operating services, an adjustable tone control is also provided in the form of another capacitor 34 and the adjustable resistor 35. The signal is further amplified in the audio-frequency amplifier 36 and is reproduced in the loud speaker 37. An optional circuit for automatic volume control is shown in dotted lines, comprising resistor 38, capacitor 39, and biasing connection 40, whose operation will be described further below. The same applies to the balancing circuit comprising resistor 41 and capacitor 42.

The detector and the limiting circuits of this invention are shown in detail in Fig. 1. The detector circuit includes two rectifying circuits in order to secure rectified voltages of both polarities. One of the rectifying circuits includes the intermediate-frequency circuit 7, 8, the diode rectifier 10 comprising cathode 12 and the anode 14, and the load resistor 18 shunted by the intermediate-frequency bypass capacitor 16. The other rectifying circuit includes the same intermediate-frequency circuit 7, 8, the other diode rectifier 11 comprising cathode 13 and anode 15, and the other load resistor 19 shunted by the other intermediate-frequency bypass capacitor 17. These two circuits deliver positive and negative rectified voltages at the respective positive (+) and negative (−) terminals relative to the common ground terminal 20.

The limiting circuits of this invention, as shown in Fig.1, are connected between the positive (+) and negative (−) input terminals, the common ground terminal 20 (shown in two places), and the output terminal 21. The limiting circuits include two circuits respectively coupling the signal from input circuit to output circuit with opposite polarities. The input circuit comprises essentially the rectifying circuits described above. The output circuit is the audio-frequency circuit beginning at terminals 20, 21, and in this example

including capacitor 33, tone control 34, 35, and the input circuit of the audio-frequency amplifier 36. One of the coupling circuits, here denoted the direct coupling circuit, is identified with the resistors 19 and 22, the capacitor 23 and the diode rectifier 24 having cathode 26 and anode 28, because these elements are essential to this coupling circuit. Likewise the other coupling circuit, here denoted the inverse coupling circuit, is identified with the resistors 18, 30, 31, the capacitor 32, and the diode rectifier 25 having cathode 27 and anode 29.

Although for certain purposes of theory or analysis the resistors 19 and 22 may be said to be in the direct coupling circuit and hence to be essential to that circuit, it is preferred for convenience of definition of the several elements of the system to consider the direct coupling path as including capacitor 23 and diode rectifier 24 and to consider the resistor 19 in the input circuit and the resistor 22 in a separate circuit herein denoted as the direct current path. Similarly with respect to the inverse coupling circuit it is convenient for purposes of definition to consider the capacitor 32, resistor 30, and diode rectifier 25 as being in the inverse coupling circuit and to consider resistor 18 as being in the input circuit and resistor 31 as being in the direct current circuit. As is pointed out hereinafter in connection with Fig. 7, it is desirable in some instances to add filters to the resistors 22 and 31 to eliminate or minimize the flow of any current other than direct current through the path in which these resistors are included, namely the direct current path.

The limiter includes two current paths in opposite directions through the same rectifiers 24, 25 connected in series with one another. One of these is the direct-current path from the positive (+) input terminal through the resistor 22, through the rectifiers 24, 25 from anode to cathode, through the resistor 31 and back to the negative (−) input terminal. Therefore, for the purposes of this path, the rectifiers are poled to conduct rectified current from the detector. The other path is the alternating current path between the positive (+) and negative (−) input terminals and including in series the capacitor 32, the resistor 30, the rectifiers 25, 24 and the capacitor 23. For the purposes of this path, the rectifiers are poled so that a sudden increase in the rectified voltages, corresponding to outward modulation of the carrier, causes current through the rectifiers which opposes the initial direct current and may be sufficient to cancel it and thereby to cut off the conduction of the rectifiers. It is this phenomenon which determines the limiting action in this circuit. The direct-current path makes the rectifiers conductive for signals within certain limits of amplitude, while the alternating-current path makes them non-conductive for signals exceeding these limits.

The capacitors 23, 32 in the alternating-current path are intended to be charged up to certain voltages determined by the direct-current path, and to retain most of this charge during signal fluctuations. Therefore the time constant of each of these capacitors with its associated resistors is preferably made greater than the greatest period of signal fluctuations to be reproduced with maximum modulation. Once each capacitor has been charged in proportion to the carrier amplitude, it behaves as a fixed bias in the limiting circuits.

Fig. 2 shows graphically the limiting characteristics of this invention, as a basis for a more complete description of the circuit relations and operation. The abscissas are proportional to the signal input voltage to the detector, which is taken as the envelope voltage of the modulated carrier that would appear across the intermediate-frequency circuit 7, 8 in Fig. 1, if the detector were disconnected. The ordinates are proportional to the rectified output voltage as it appears between terminals 20 and 21 after the limiting action. This plot is based on a preferred relation that the resistors 22 and 31 have the same resistance ratio as resistors 18 and 19, respectively. This ratio may be unity, as in the representative set of circuit constants given below. This preferred relation causes the output voltage to be zero for an unmodulated carrier.

In Fig. 2, the signal input voltage in general is denoted E, and its critical values are denoted by subscripts as follows:

E_0 is the carrier voltage;
E_1 is the limit of inward modulation;
E_2 is the limit of outward modulation;
E_3 is the limit of pulse inversion.

The limit of inward modulation is imposed by the rectifiers 10, 11 in the detector circuit as a result of the load imposed by the limiter. The limit of outward modulation is imposed by cutoff in the rectifier 24 in the direct coupling circuit. The limit of pulse inversion is imposed by cutoff in the rectifier 25 in the inverse coupling circuit.

From these limits in Fig. 2, the range of normal signal detection is along the line 42 between the inward limit E_1 and the outward limit E_2. This range is approximately centered on the carrier voltage E_0 and the output voltage is proportional to the modulation of the carrier. In this range, all the rectifiers are conducting. Through the direct and inverse coupling circuits, the direct coupling of rectified modulation through capacitor 23 and rectifier 24 is predominant because the inverse coupling through the other capacitor 32 and rectifier 25 is reduced by their series resistor 30. Since the direct coupling is from the negative (—) input terminal, the slope of the line 42 is negative.

The inward limit E_1 is imposed by the failure of conduction in the rectifiers 10, 11, caused by the load conductance toward rectified modulation exceeding the load conductance toward the rectified carrier. In other words, the alternating-current load conductance exceeds the direct-current load conductance, because both of the paths in the limiter carry alternating current while only one carries direct current. This limiting action is shown by the horizontal line 41 and is effective for any inward modulation which reduces the signal voltage below the inward limit E_1.

The other two limits are imposed by the particular action of this limiter circuit. As mentioned above, the alternating-current through the rectifiers 24, 25 during outward modulation, opposes the current through the direct-current path. Outward modulation to the level E_2 is the amount which just reduces to zero the current in the first rectifier 24 and thereby cuts off the coupling in the direct coupling circuit. The second rectifier 25 carries also the slight additional current required by the audio-frequency load between the output terminals 20, 21. Therefore the second rectifier remains conductive as long as the output voltage is negative. The line 43 represents the performance while the first rectifier 24 is cut off but the second rectifier 25 remains conductive. This line shows how extreme peaks of outward modulation exceeding E_3 are inverted. This is a great improvement over the usual limiting action represented by the horizontal line 45, in which the peaks of outward modulation are merely clipped at that level.

While the inversion of the peaks of excessive noise pulses is in itself a substantial advantage, the complete form of the present invention provides a further improvement. For input voltages in excess of E_3 in Fig. 2, the second rectifier 25 also cuts off and imposes a limit on the inverted peaks. Beyond E_3, the effective line of the graph is denoted 44, and is coincident with the horizontal axis of zero output voltage. In other words, a very strong pulse of noise would have its peak inverted and cut off at the level of zero output voltage, so it would cause minimum disturbance in the output of the limiter. This result is the sole function of the second rectifier 25. If this result is not required, the second rectifier 25 may be omitted and replaced by a direct connection. This may be preferable if some limiting action before the detector serves to prevent noise peaks substantially in excess of E_3. If this simplification is employed, the effective line beyond E_3 is a continuation of the straight line 43 instead of the line 44.

Curve 42 corresponds to normal operation with all three rectifiers conducting. The inward cut off line 41 corresponds to the condition when rectifier 10 is non-conducting and the others are conducting. The peak inversion line 43 corresponds to the condition when the rectifier 24 is non-conducting and the others conducting; this curve illustrates an important feature of my invention. The inversion cut off line 44 represents the condition when rectifiers 24 and 25 are both non-conducting.

If any particular form of curve be desired it is possible to approximate it by appropriate choice of circuit constants in my improved limiter circuit.

Fig. 3 shows the benefits of this invention as compared with other schemes for limiting noise pulses. In each graph, the abscissas are proportional to time and the ordinates are, as in Fig. 2, proportional to the rectified output voltage. However, the ordinates are inverted relative to Fig. 2 in order to show the rectified outward modulation in the upward direction. Fig. 3 (*a*) is an idealized presentation of excessive noise pulses of short duration superimposed on a sine wave of rectified modulation, without the benefit of any limiting action. The horizontal axis 46 represents zero output for an unmodulated carrier, such as E_0 in Fig. 2. The lower and upper dotted lines 47 and 48 represent the limits of complete inward and outward modulation. The sine wave 49 represents a signal modulation of about one-half of complete modulation. Three noise pulses 50, 51, 52 are shown superimposed on the sine wave of modulation at different phases so all conditions are exemplified. Each noise pulse is idealized to simplify the presentation, by showing it as rectangular in shape. The flat peak, however, may be approximated by limiting action before the detector.

In Fig. 3 (*b*) is illustrated the action of one type of limiter, which cuts off the intermediate-frequency amplifier for the duration of the noise pulse. This establishes during the pulse a level of zero for the modulated carrier, but this is actually the level of complete inward modulation 47. The response to the three noise pulses of Fig.

3 (*a*) is therefore as shown by the inverted residual pulses **53, 54, 55.**

In Fig. 3 (*c*) is shown the action of another type of limiter, which merely cuts off the peaks at the level of complete outward modulation **48** which is twice the assumed carrier amplitude. The three pulses **56, 57, 58** show what remains of the pulses of Fig. 3 (*a*). While the behavior toward each pulse is different, the total residual disturbance caused by a number of pulses at random is about the same as in Fig. 3 (*b*).

Fig. 3 (*d*) shows the improved result obtained by pulse inversion in the simplified form of the present invention, with the omission of the second rectifier **25** as mentioned above. The three pulses assume the forms **59, 60, 61.** While the edges of each pulse rise to the limit of outward modulation **48**, the peak is depressed to a level which is actually within the useful range of modulation, between **47** and **48**. This makes the net disturbance less than in either Fig. 3 (*b*) or Fig. 3 (*c*).

Fig. 3 (*e*) shows the ultimate improvement obtained by the complete form of the invention. The three pulses are reduced to the shapes **62, 63, 64**, in which the inverted peaks are cut off at the average amplitude of the signal **46** which is the level of zero modulation. On the average, the limiting of the peaks at this level causes the minimum of disturbance to the signal. This is especially true during weak modulation, when any one of the other forms of limiters leaves the peaks stronger than the signal. The residual edges of the pulses are of much shorter duration than the entire pulse so they are more susceptible to smoothing by such means as the shunt capacitor **33** and the tone control **34, 35.**

Various detector circuits may be employed to rectify the modulated-carrier signal and thus provide a source of voltage. In Fig. 1, rectified voltages of both polarities are obtained by two rectifying circuits connected in parallel across the intermediate-frequency output circuit **7, 8**, and having their load resistors **18, 19** connected with opposite polarity in the limiter input circuit. This is a full-wave rectifier in which the half-waves of each polarity are rectified to secure a voltage of the corresponding polarity.

An alternative detector circuit is shown in Fig. 4, requiring only the one rectifier **10** instead of two rectifiers. In this case, the current from a half-wave rectifier is conducted through the two load resistors **18** and **19** in series to secure rectified voltages of opposite polarity relative to the terminal **20** at the junction of these resistors. Substantially the same results are obtained from a proper design of either circuit and the remainder of the limiting arrangement is the same in either case. Since the detector circuit of Fig. 4 is simpler, it will be used as the basis for further analysis.

A paper by H. A. Wheeler entitled "Design Formulas for Diode Detectors" (Proceedings I. R. E., June 1938, pp. 745–780) gives, in Figs. 11 and 13 thereof, a fundamental method for using low frequencies only in demonstrating the behavior of a diode rectifier. Fig. 5 shows the same point of view adapted to the detector and limiter of the present invention. The purpose is to demonstrate its action by means of a direct-current circuit. The resistors **18, 19, 22, 30, 31** are all retained in their same circuit relations. The same applies to the diode rectifiers **10, 24, 25.** The capacitors **23** and **32** are intended to hold a fixed voltage during modulation so they are replaced by batteries **23'** and **32'**. The audio-frequency output circuit is replaced by a zero-center direct-current voltmeter V assumed to have much greater resistance than the associated resistors. The intermediate-frequency signal input circuit is replaced by a source of adjustable direct voltage E in series with a resistor **66**. The voltage E is equal to the envelope voltage of the modulated signal which would appear across the circuit **7, 8** in Fig. 4 if the rectifier **10** were disconnected. The resistor **66** is double the carrier-frequency resistance which would be measured across the circuit **7, 8** at resonance, looking backward into the intermediate frequency amplifier; this relation is taken from the Wheeler paper.

In Fig. 5, each resistor is also given a value of conductance for reference in the formulas to be given. For simplicity in analysis, and also in many cases, in practice, I prefer to make resistors **18** and **19** of Figs. 1 and 4 equal. In the present analysis each is given the conductance value G_1. The resistors **22** and **31** will then also be equal and the conductance of each is denoted G_2. The resistors **30** and **66** are given the conductance values G_3 and one-half G_0 respectively.

From the equality of resistors **22** and **31** it devolves that capacitors **23** and **32** would be charged to the same voltage, and this voltage is assigned to each of the batteries **23'** and **32'** in Fig. 5. The envelope voltage E when unmodulated is equal to the carrier envelope voltage E_0. This voltage is somewhat greater than the total voltage of the two batteries **23'** and **32'**, by the amount of the voltage drop in resistor **66**. The rectifiers **10, 24, 25** are assumed to conduct current perfectly with zero resistance in the conductive direction, and to cut off the current completely with zero conductance in the reverse direction. This and other assumptions inherent in the demonstration circuit of Fig. 5 can be approximated in practice, with substantially the same results in limiting action.

In order to evaluate the action in the direct-current circuit of Fig. 5, the output voltage is observed in voltmeter V in response to variation of the input voltage E for the several conditions of the three rectifiers **10, 24, 25.** Referring to Fig. 2, the normal operating line **42** is computed with all three rectifiers conducting. The inward cutoff line **41** is computed with the rectifier **10** non-conducting and the others conducting. The peak inversion line **43** is computed with the rectifier **24** non-conducting and the others conducting. The inversion cutoff line **44** is the obvious result with rectifiers **24** and **25** both non-conducting. The formulas for these lines determine the critical points of intersection thereof, and these points are of most interest in the present invention. Each of these points corresponds to the cutoff in one of the rectifiers. The following formulas enable the evaluation of these points from given circuit values or of the design values for a circuit to secure given relations among these critical points.

The first set of formulas determines the relations among the critical points in Fig. 2 for given circuit values in the notation of Fig. 5.

$$m_1=\frac{E_0-E_1}{E_0}=1-\frac{G_0}{G_0+G_1+G_2}\,\frac{G_2+2G_3}{G_1+2G_2+2G_3}=$$

$$1-\frac{1}{1+\dfrac{G_1+G_2}{G_0}}\,\frac{1}{1+\dfrac{G_1+G_2}{G_2+2G_3}} \quad (1)$$

$$m_2=\frac{E_2-E_0}{E_0}=\frac{G_0+G_1+2G_2+G_3}{G_0+G_1+G_2}\,\frac{G_2}{G_3} \quad (2)$$

$$m_3=nm_2=\frac{E_3-E_0}{E_0}=\frac{G_0+G_1+2G_2+G_3}{G_0+G_1+G_2}\,\frac{G_2}{G_3-G_2}\qquad(3)$$

$$n=\frac{m_3}{m_2}=\frac{E_3-E_0}{E_2-E_0}=\frac{G^3}{G_3-G_2}>1\qquad(4)$$

in which:

G_0, G_1, G_2, G_3, are the conductance values of the circuit elements so designated in Fig. 5, all expressed in the same units, such as micromhos;

E_0, E_1, E_2, E_3 are the critical values of the input voltage so designated in Fig. 2, all expressed in the same units, such as volts;

m_1 is the modulation factor of the limit of inward modulation;

m_2 is the modulation factor of the limit of outward modulation;

m_3 is the modulation factor of the pulse peak which just reaches the limit of inversion;

n is the ratio of the two preceding modulation factors as defined in Equation 4.

It is noted that m_1 is always less than unity, m_2 may be less or greater than unity, and m_3 is greater than m_2 so n is always greater than unity. With reference to Fig. 4, G_0 is equal to the intermediate-frequency conductance presented by the circuit 7, 8 in that arrangement. This situation is altered by the parallel connection of the rectifiers 10, 11 in Fig. 1, so that in other arrangement $4G_0$ is the intermediate-frequency conductance presented by the circuit 7, 8.

A special case is noted for the condition

$$G_0=G_1=G_2$$

and $G_3=2G_2$. From the above formulas,

$$m_1=16/21=0.76$$

$m_2=1$; $m_3=2$; $n=2$. This would give very good performance.

On the other hand, if the critical points in Fig. 2 are given, the following second set of formulas determines the ratios of conductance required in the circuit design. They are expressed relative to G_2 because this plan gives the simplest formulas. They are expressed in terms of the modulation factors defined above, which can be computed from the critical voltages by the first set of formulas.

$$\frac{G_3}{G_2}=\frac{m_3}{m_3-m_2}=\frac{n}{n-1}>1\qquad(5)$$

$$\frac{G_0+G_1}{G_2}=\frac{2(n-1)+n(1-m_2)}{1-n(1-m_2)}\qquad(6)$$

$$\frac{G_0}{G_2}=(1-m_1)\,\frac{\dfrac{2n-1}{1-n(1-m_1)}+\dfrac{3n-1}{n-1}}{\dfrac{1-n(1-m_1)}{2n-1}\,\dfrac{3n-1}{n-1}+(1-m_1)}\qquad(7)$$

$$\frac{G_1}{G_2}=\frac{G_0+G_1}{G_2}-\frac{G_0}{G_2}\qquad(8)$$

All symbols have the same meanings as defined above. It is noted that Equation 6 involves only m_2 and n, while 7 involves only m_1 and n.

The above formulas can be satisfied only if m_2 is within the following limits:

$$\frac{n-1}{n}<m_2<\frac{3n-2}{n}\qquad(9)$$

This is always satisfied if $m_2=1$, which is an ideal value if complete outward modulation is to be preserved. If m_2 is less than one, the same condition requires that

$$n<\frac{1}{1-m_2}\qquad(10)$$

From Equation 5 it appears that G_3 approaches G_2 if n is very large. This approximates the condition of level limiting according to the line 45 in Fig. 2 and the resulting pulse behavior as in Fig. 3 (*c*). The benefits of this invention in respect to the inversion of pulse peaks are secured only if G_3 is substantially greater than G_2. On the other hand, other requirements may restrict G_3 within a range just a little greater than G_2, but this difference is substantial for the purposes of this invention. In other words, under the assumptions of Fig. 5, the resistor 30 is likely to have less resistance but not very much less than each of the resistors 22 and 31.

In general, a lesser value of either of the following conductance ratios permits further inward modulation, according to Equation 1;

$$\frac{G_0}{G_1+G_2}\text{ and }\frac{G_2+2G_3}{G_1+G_2}\qquad(11)$$

Therefore this objective is promoted by reducing G_0 and G_3, the latter requiring also reducing G_2. In other words this objective is furthered by increasing G_1 or reducing any of the other conductances.

From the foregoing it will be seen that certain relations in the coupling circuit are required to secure inversion of the peaks of noise pulses in accordance with this invention. On the assumption of equal resistors 18 and 19, it is found that equality of resistors 30 and 31 gives merely a level cutoff of the peaks as illustrated in Fig. 3 (*c*) according to the line 45 in Fig. 2. To secure inversion of the peaks in this case, it is necessary that the value of resistor 30 be less than that of 31. In general, it must be substantially less than some specific value determined by the other resistors in the circuit, but need not be very much less, to obtain the benefits of inversion.

There may be some capacitive induction of pulse currents through the rectifiers 24 and 25 even when they are cut off against conduction. Fig. 6 shows some simple expedients for neutralizing the coupling caused by such induction. It is a slight modification of the limiter circuit from Fig. 1, by the addition of two small neutralizing capacitors 24′ and 25′, each having about the same capacitance as the rectifier of the same reference number. The capacitor 24′ is adjusted to neutralize the capacitive coupling through the first rectifier 24, and the capacitor 25′ is likewise adjusted with respect to the second rectifier 25. Assuming equal values of resistors 18 and 19 in Fig. 1, as well as resistors 22 and 31, the latter is divided in two parts for neutralizing purposes as follows. One part 31′ is made equal to 30 while the other part 31″ includes the remainder of 31. Each rectifier may be neutralized separately by disconnecting the other rectifier and neutralizing capacitor, then balancing for minimum output in response to noise pulses. The neutralizing capacitors may be made of dummy diodes of the same structure as used for the diode rectifiers 24 and 25, to obtain approximate neutralization without adjustment.

An automatic volume control may be added in Fig. 1 as shown in dotted lines. The negative rectified voltage developed at the negative (−) terminal is freed of audio-frequency fluctuations by the series resistor **38** and the shunt capacitor **39** in the usual manner. The resulting negative voltage is then applied through the connection **40** to reduce the gain in one or more stages in sections **3** and **4**. To retain the balance in the limiter, it may be desirable to add a dummy circuit on the positive (+) terminal, comprising resistor **41** equal to **38** and capacitor **42** equal to **39**. Some of the above formulas would be modified by the addition of the control circuit, particularly the formulas involving the limit on inward modulation.

The intermediate-frequency amplifier **4** inherently limits the peak amplitude of noise pulses at a level which is usually substantially above the maximum amplitude of the signal during outward modulation. In utilizing the present invention to greatest advantage, it is desirable that the limiting level in this amplifier be above the level of inversion limiting, denoted E_3 in Fig. 2. If automatic volume control is provided to hold the carrier amplitude E_0 below a certain value, this condition can be assured at all times. As a result, all strong pulses of noise would reach the detector with sufficient peak amplitude to be inverted down to the level of zero modulation, as illustrated in Fig. 3 (*e*).

As a representative circuit design embodying the invention, the following circuit constants are suggested:

Shunt conductance of circuit **7**, **8**: $G_0=4$ micromhos
Resistors **18**, **19**: $G_1=2$ micromhos
Resistors **22**, **31**: $G_2=1$ micromhos
Resistor **30**: $G_3=4/3$ micromhos
Capacitors **16**, **17**: 25 micromicrofarads
Capacitor **23**: 0.04 microfarad
Capacitor **32**: 0.02 microfarad
Capacitor **33**: 0.0005 microfarad
Capacitor **34**: 0.0025 microfarad

Fig. 2 shows approximately the performance obtained by these constants, as represented by the following values of the modulation factors computed from the above formulas:

$$m_1=24/35=0.68$$
$$m_2=1$$
$$m_3=4$$
$$n=4$$

In practice it may be desirable to secure closer limiting of both inward and outward modulation, in which case certain constants in the preceding table may be changed to the following values, the other remaining the same.

Resistors **18**, **19**: $G_1=1$ micromhos
Resistor **30**: $G_3=2$ micromhos

The resulting performance factors will be modified as follows, showing limiting at smaller modulation factors:

$$m_1=11/21=0.52$$
$$m_2=0.75$$
$$m_3=1.5$$
$$n=2$$

Fig. 7 shows certain modifications of Fig. 1 which require additional circuit elements but may in some cases yield advantages to justify the additions. There is redrawn only that part of the circuit which includes the changes. The elements which retain similar functions in the circuit are numbered as in Fig. 1.

In Fig. 7, the principal changes are in the direct-current path. The first resistor, **22** in Fig. 1, is divided in two parts **22'** and **22''**, and an added capacitor **71** is connected between the intermediate point and the common ground terminal. Likewise the second resistor, **31** in Fig. 1, is divided into two parts **31'** and **31''**, and an added capacitor **72** is connected between the intermediate point and the common ground terminal.

The purpose of these changes is to obviate an incidental and undesired action of the resistors **22** and **31** in the direct-current path in Fig. 1. These resistors are needed only to provide direct current through the rectifiers **24** and **25** in order to establish proper cutoff levels. However, the same resistors also carry some alternating current which is in opposition to that carried by the alternating-current path **23**, **30**, **32**. The added capacitors **71** and **72** in Fig. 7 serve to filter out any modulation-frequency current which would otherwise be conducted through the rectifiers **24** and **25** by way of the direct-current path.

Another change in Fig. 7 is the division of another resistor, **19** in Fig. 1, into two parts **19'** and **19''**, so the capacitor **23** can be connected to the intermediate point. The ratio of the two parts is chosen to give the desired amount of coupling in the direct circuit, through capacitor **23** and rectifier **24**.

During the time the diode **24** in Fig. 7 is nonconductive, it still provides undesired capacitive coupling to the output terminal. After the manner of Fig. 6, a very small added capacitor **24''** is connected and adjusted to neutralize this capacitive coupling.

Representative circuit constants in Fig. 7 are as follows:

Resistors **18**, **22'**, **22''**, **31'**, **31''**=2 micromhos
Resistor **19'**=2.85 micromhos
Resistor **19''**=6.66 micromhos
Resistor **30**=0.5 micromho
Capacitors **23**, **71**, **72**=0.05 microfarad
Capacitor **32**=0.01 microfarad

The arrangement of Fig. 8 is a modification of Fig. 1 to operate from a detector circuit which delivers a rectified voltage of only one polarity. The detector circuit corresponds to Fig. 4 insofar as the elements **7**, **8**, **10** are concerned. However, the detector is connected to deliver a rectified voltage of only one polarity (negative) relative to the common ground terminal **20**. The shunt capacitors **74**, **17** and the series resistor **75** are designed to filter out the carrier-frequency components and pass the modulation-frequency components from the rectifier **10**. The resistor **86** and the remainder of the limiter circuit form the direct-current and modulation-frequency load on the detector circuit.

Since the limiter requires both polarities of rectified modulation voltage, a polarity-reversing repeater is added to supplement the one polarity of rectified voltage from the detector. This repeater comprises the vacuum tube **82** and associated circuits, connected as a modulation-frequency amplifier. The tube **82** is a triode with cathode **83**, control grid **84** and anode **85**. The required direct current in the anode circuit is supplied by the battery **79**, or equivalent, through the resistor **80**. The effective transconductance of the tube **82** is reduced and stabilized by the resistor **81** in the cathode lead.

In Fig. 8, the direct coupling circuit comprises the divided resistor **76**, **77**, **78** and the limiting rectifier **24**. The provision of the direct-current path in this case, yet to be described, requires that the rectifier **24** be connected to an intermediate point, as shown between the parts 77 and 78.

The inverse coupling circuit in Fig. 8 includes the polarity-reversing repeater circuit. The grid 84 is connected to a point between the parts 76 and 77, of the same polarity as the direct coupling. The output of the repeater from its anode 85 contains the modulation components in opposite polarity. The remainder of the inverse circuit comprises the capacitor 32, the resistors 30, 31, and the other limiting rectifier 25. The capacitor 32 removes the direct current and passes modulation components. The voltage ratio of the repeater 82 and the ratio of resistors 30 and 31 are so proportioned that the principal coupling through the rectifier 25 comes from the repeater through resistor 30, not from the detector directly through resistor 31. This relation establishes the inverse coupling in this circuit.

The direct-current path identified in Figs. 1 and 7 is between input terminals having rectified voltages of opposite polarity. In Fig. 8, the corresponding direct-current path is between the common terminal and an input terminal of one polarity. It includes in series the resistor parts 76, 77, the rectifiers 24, 25, and the resistor 31.

The alternating-current path in Fig. 8, as in Figs. 1 and 7, is connected between terminals having modulation components of opposite polarities. This path comprises the resistor part 78, the rectifiers 24, 25, the resistor 30 and the capacitor 32 connected in series.

The direct-current and alternating-current paths perform the same function in all figures, as described with reference to Fig. 1. Likewise, the limiting action of the rectifiers 24, 25 is the same in all figures.

The following list of circuit constants is representative of a proper design in accordance with Fig. 8.

Resistors 30, 31=1 micromho
Resistor 76=10 micromhos
Resistor 77=6.66 micromhos
Resistor 78=4 micromhos
Resistor 80=2 micromhos
Resistor 81=100 micromhos
Capacitor 32=0.04 microfarad
Voltage 79=250 volts
Triode 82=6J5 type

The arrangement shown in Fig. 9 includes in one combination a number of features found in the other figures. The detector circuit is like Fig. 8, with the omission of elements 74, 75 which may not be needed. The automatic volume control is similar to that shown in dotted lines in Fig. 1, comprising elements 38, 39, 40. As in Fig. 8, the detector supplies only one polarity of rectified voltage and the other polarity of modulation components is obtained by the use of a polarity-reversing repeater 82. In one part of the direct-current path, a shunt capacitor 72 is provided to obviate alternating-current coupling through that path, as in Fig. 7.

The repeater in Fig. 9 includes a vacuum tube 82 of pentode type instead of triode type, serving the same function. The battery is divided in two parts 88, 89 to furnish an intermediate voltage for the usual screen grid. The same tube is used to actuate the meter 90 to give an indication of relative signal strength.

The divided resistor 67, 68, 69 is connected between parts 67, 68 to the grid 84 of the repeater and to the rectifier 24. The automatic volume control is connected between parts 68, 69.

The direct coupling circuit includes elements 67, 68, 69, 24. The inverse coupling circuit includes the repeater 82, with its accessories, and elements 32, 30, 69, 72, 25.

The direct-current path includes elements 67, 24, 25, 73, 70. The alternating-current path includes elements 69, 68, 24, 25, 30, 32.

Representative circuit constants for Fig. 9 are as follows:

Resistor 30=0.05 micromho
Resistors 67, 68=10.0 micromhos
Resistor 69=20 micromhos
Resistor 70=2 micromhos
Resistor 73=1 micromho
Capacitor 32=0.02 microfarad
Capacitor 72=0.05 microfarad
Voltage 88=100 volts
Voltage 89=180 volts
Pentode 82=6B8 type

For convenience and clarity of description, some terms are used herein which should not be interpreted as unduly limiting the scope of the invention. For example, the "direct" and "inverse" coupling circuits do not necessarily provide coupling respectively of direct and reverse polarity. These terms denote the circuits whose couplings predominate respectively during normal reproduction of the signal and during inversion of the excess pulse. These two couplings do always oppose each other, as indicated by the terms "direct" and "inverse."

The terms "direct-current path" and "direct coupling circuit" should not be confused with one another; "direct-current path" is used to distinguish from "alternating-current path" whereas "direct coupling circuit" is used to distinguish from "inverse coupling circuit." The direct current path and the alternating current path are divided between the two coupling circuits.

Each rectifier may be a diode vacuum tube or any other form of rectifier whose conduction in one direction far exceeds that in the other direction.

The "greatest period of signal fluctuations to be reproduced" may be interpreted as follows in the case of sine-wave modulation at low audio frequencies. The "period of fluctuation" is taken as a part of a complete cycle, such as the "radian period" which is about one-sixth of one cycle.

While there have been described what are considered to be preferred embodiments of the invention, various modifications may be made while adhering to the invention. It is therefore intended in the appended claims to cover all such modifications as fall within the true spirit and scope of the invention.

1. A signal coupling and amplitude limiting system comprising an input circuit including an element across which rectified voltage of a modulated-carrier signal appears, an output circuit, means for deriving from said input circuit two opposing modulation voltages, a direct coupling circuit for coupling one of said opposing modulation voltages from said input to said output circuit, a rectifier connected in said direct coupling circuit, a direct current circuit connected through said rectifier and across at least part of the source of rectified modulated-carrier causing said rectifier to conduct modulation the amplitude of which does not exceed a cutoff level determined by the potential of the rectified carrier,

and an inverse coupling circuit for coupling the other of said modulation voltages from said input circuit to said output circuit for at least amplitudes above the cutoff level of the direct coupling circuit, the time constants of said direct and inverse coupling circuits being greater than the greatest period of single fluctuations to be reproduced and said direct and inverse coupling circuits being so proportioned that the direct coupling circuit provides the greater coupling for amplitudes below the cutoff level of the direct circuit, whereby the peak in any pulse exceeding the cutoff level in the direct coupling circuit is inverted in the output circuit.

2. A signal coupling and amplitude limiting system in accordance with claim 1 in which the direct coupling circuit transmits one polarity of the modulation component of the rectified modulated-carrier signal and the means for securing an opposing polarity of the modulation component comprises a reversing repeater connected in the inverse coupling circuit.

3. A signal coupling and amplitude limiting system in accordance with claim 2 in which a rectifier is included in said inverse coupling circuit and the direct current circuit includes both of said rectifiers whereby the peaks of amplitude in both the direct coupling circuit and the inverse coupling circuit are limited at levels determined by the potential of the rectified carrier, said direct and inverse coupling circuits being so proportioned that the cutoff level of the inverse coupling circuit substantially exceeds the cutoff level of the direct coupling circuit whereby the amplitude of inverted peaks may be reduced to zero.

4. A signal coupling and amplitude limiting system in accordance with claim 3 in which the direct current circuit includes a filter to eliminate modulation components and limit the current of such circuit to substantially direct currents.

5. A signal coupling and amplitude limiting system in accordance with claim 1 in which a rectifier is included in said inverse coupling circuit and the direct current circuit includes both of said rectifiers whereby the peaks of amplitude in both the direct coupling circuit and the inverse coupling circuit are limited at levels determined by the potential of the rectified carrier, said direct and inverse coupling circuits being so proportioned that the cutoff level of the inverse coupling circuit substantially exceeds the cutoff level of the direct coupling circuit whereby the amplitude of inverted peaks may be reduced to zero.

6. A signal coupling and amplitude limiting system in accordance with claim 1 in which the direct current circuit includes a filter to eliminate modulation components and limit the current of such circuit to substantially direct currents.

7. A signal coupling and amplitude limiting system in accordance with claim 1 in which the direct coupling circuit comprises, in addition to the rectifier, a resistor connected between one of the two opposing modulation voltages and one side of the rectifier, the other side of the rectifier is connected to an output terminal, the inverse coupling circuit comprises a capacitor and a resistor connected in series between the other of the two opposing modulation voltages and said output terminal, and the direct current circuit comprises a resistor, said rectifier, and a second resistor connected in series in the order named from one of said two opposing voltages to the other.

8. A signal coupling and amplitude limiting system in accordance with claim 7 in which a rectifier is included in said inverse coupling circuit and the direct current circuit includes both of said rectifiers.

9. A signal coupling and amplitude limiting system in accordance with claim 7 in which the two opposing modulation voltages are equal, the two resistors in the direct current path are equal, and the value of the resistor in the inverse coupling circuit is less than the values of the resistors in the direct-current circuit.

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No references cited.

Certificate of Correction

Patent No. 2,453,958.

November 16, 1948.

GILBERT J. C. ANDRESEN

It is hereby certified that errors appear in the printed specification of the above numbered patent requiring correction as follows:

Column 9, lines 4 and 5, for that portion of Equation 4 reading

$$“=\frac{G^3}{G_3-G_2}>1” \quad \text{read} \quad =\frac{G_3}{G_3-G_2}>1$$

line 30, for “that in” read *in that*; column 11, line 55, for the word “other” read *others*; column 15, line 7, for “single” read *signal*;

and that the said Letters Patent should be read with these corrections therein that the same may conform to the record of the case in the Patent Office.

Signed and sealed this 15th day of March, A. D. 1949.

[SEAL]

THOMAS F. MURPHY,
Assistant Commissioner of Patents.