

Sept. 27, 1949.

F. W. FRINK

2,483,271

FREQUENCY MODULATION TRANSMITTER

Filed Nov. 30, 1946

6 Sheets-Sheet 1

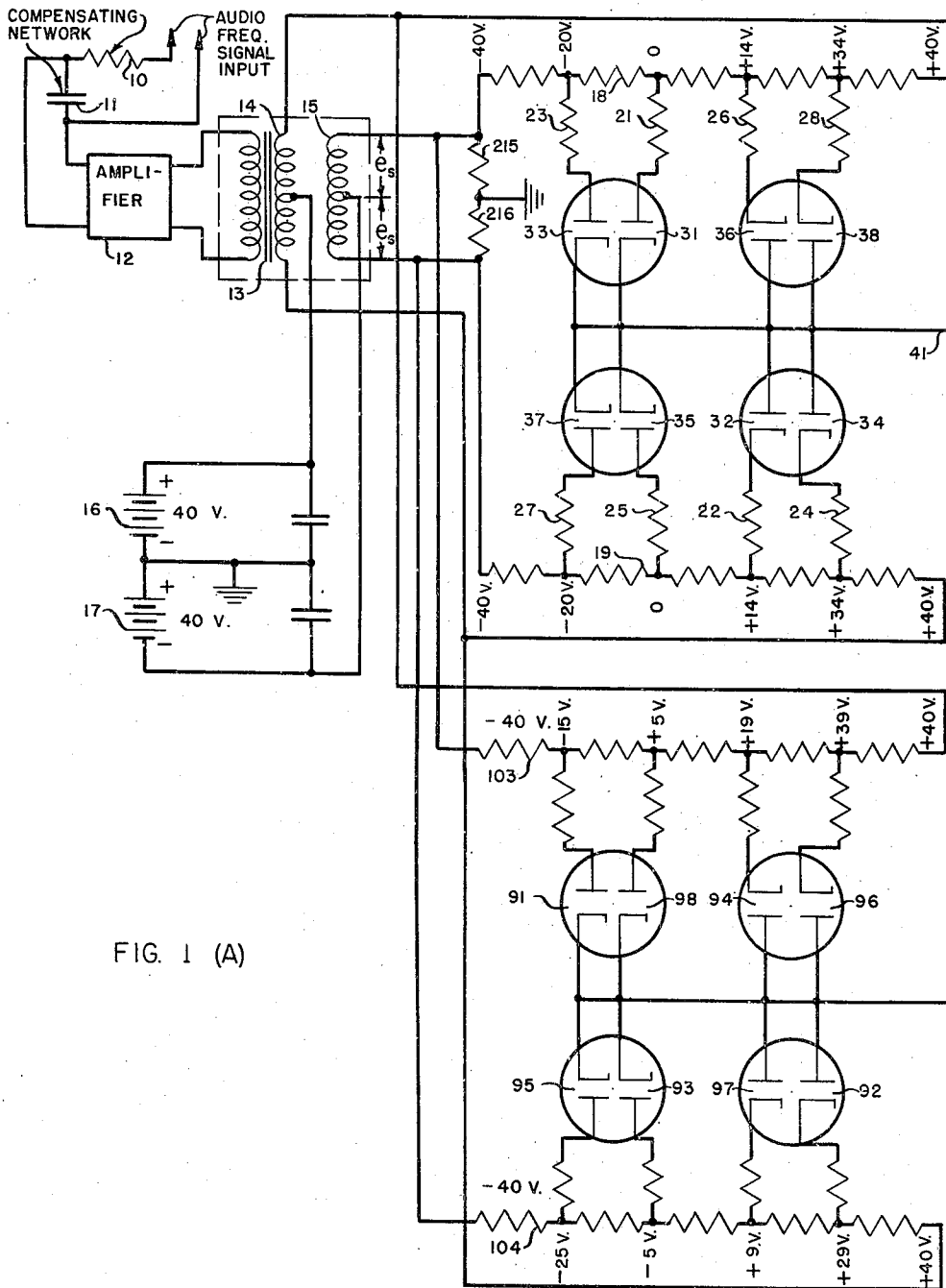


FIG. 1 (A)

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6 Sheets-Sheet 2

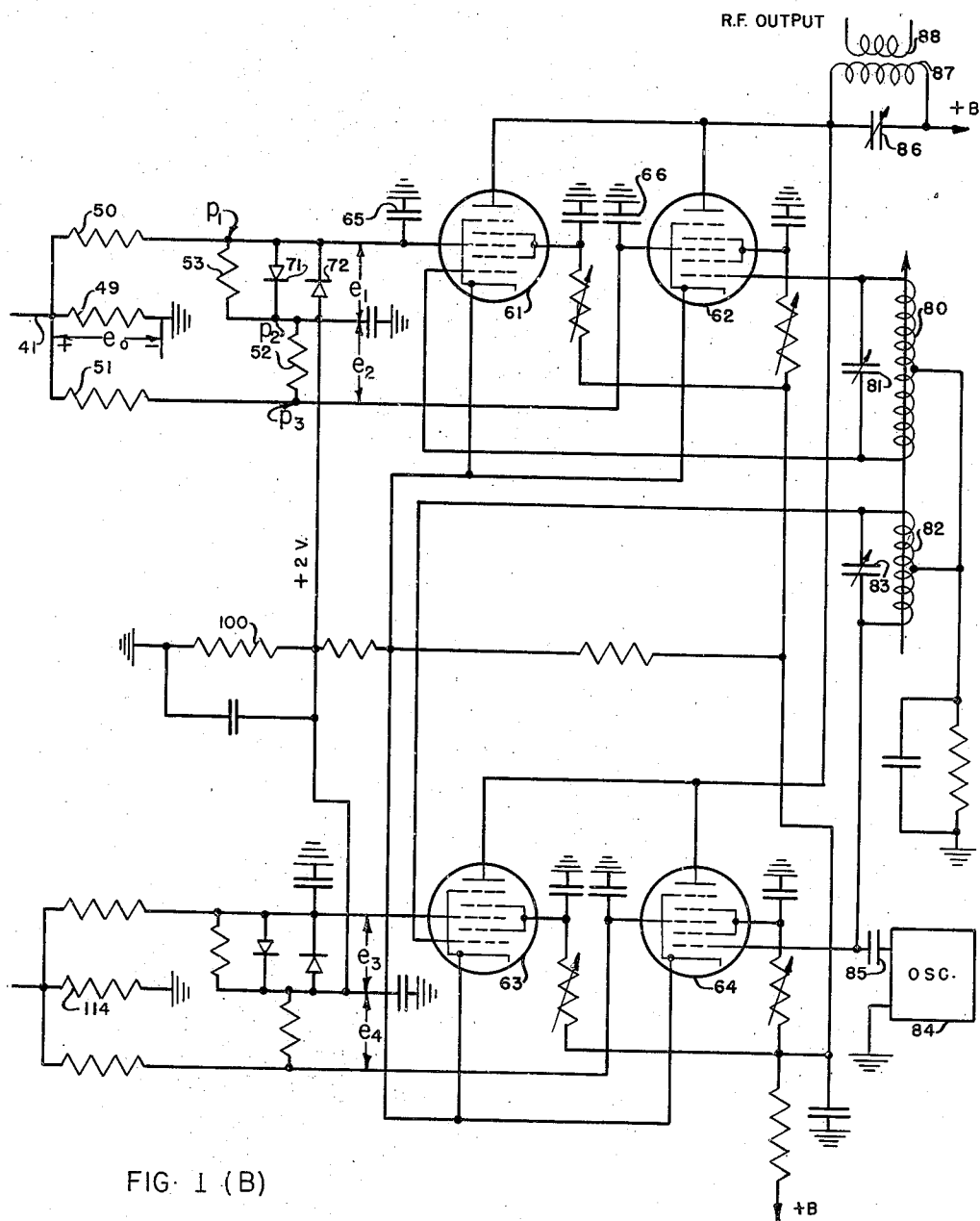


FIG. 1 (B)

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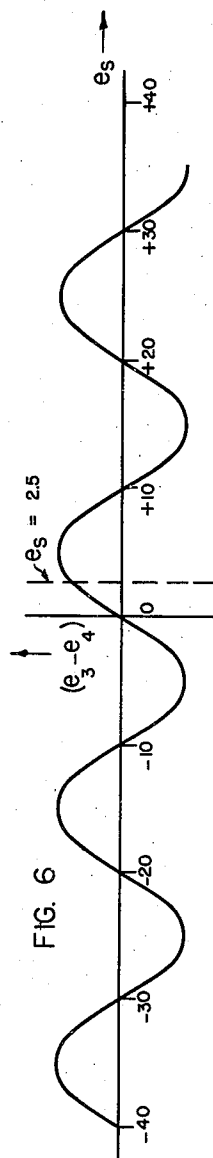
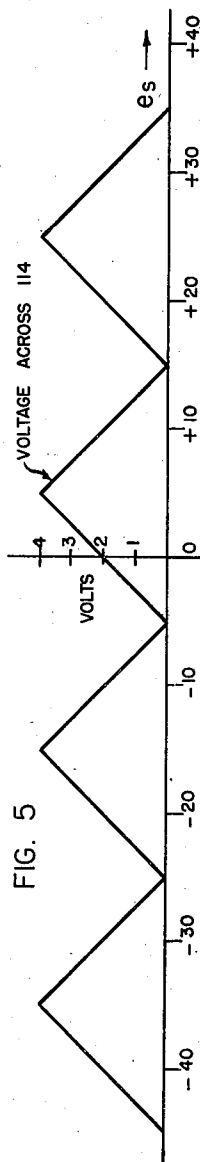
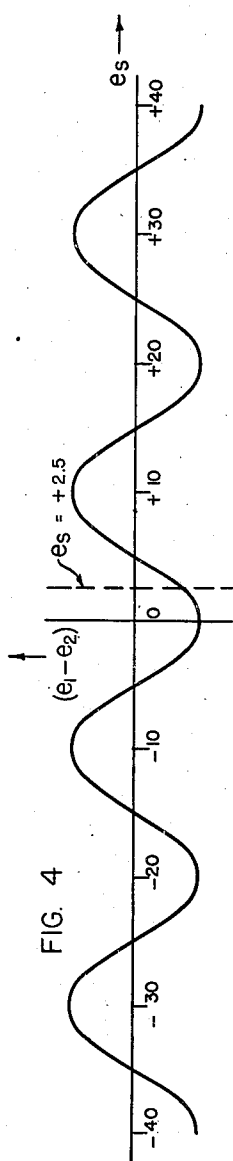
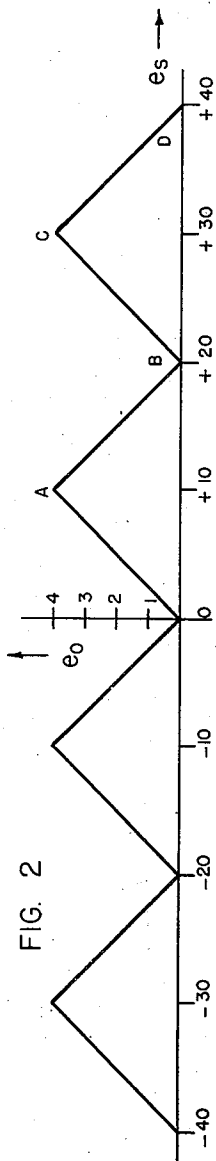
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FREQUENCY MODULATION TRANSMITTER

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6 Sheets-Sheet 3



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FREQUENCY MODULATION TRANSMITTER

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FIG. 3

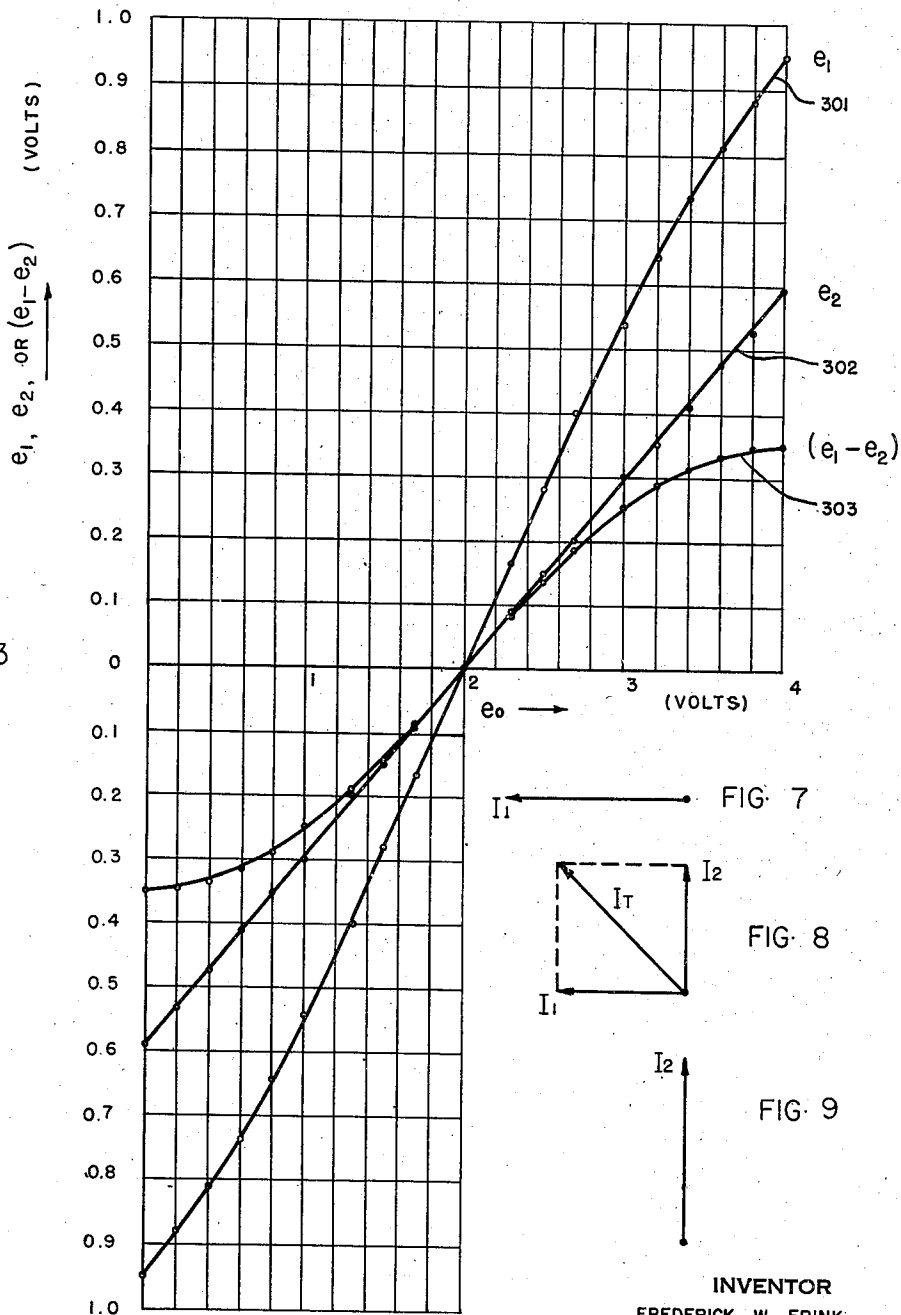


FIG. 7

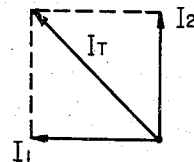


FIG. 8

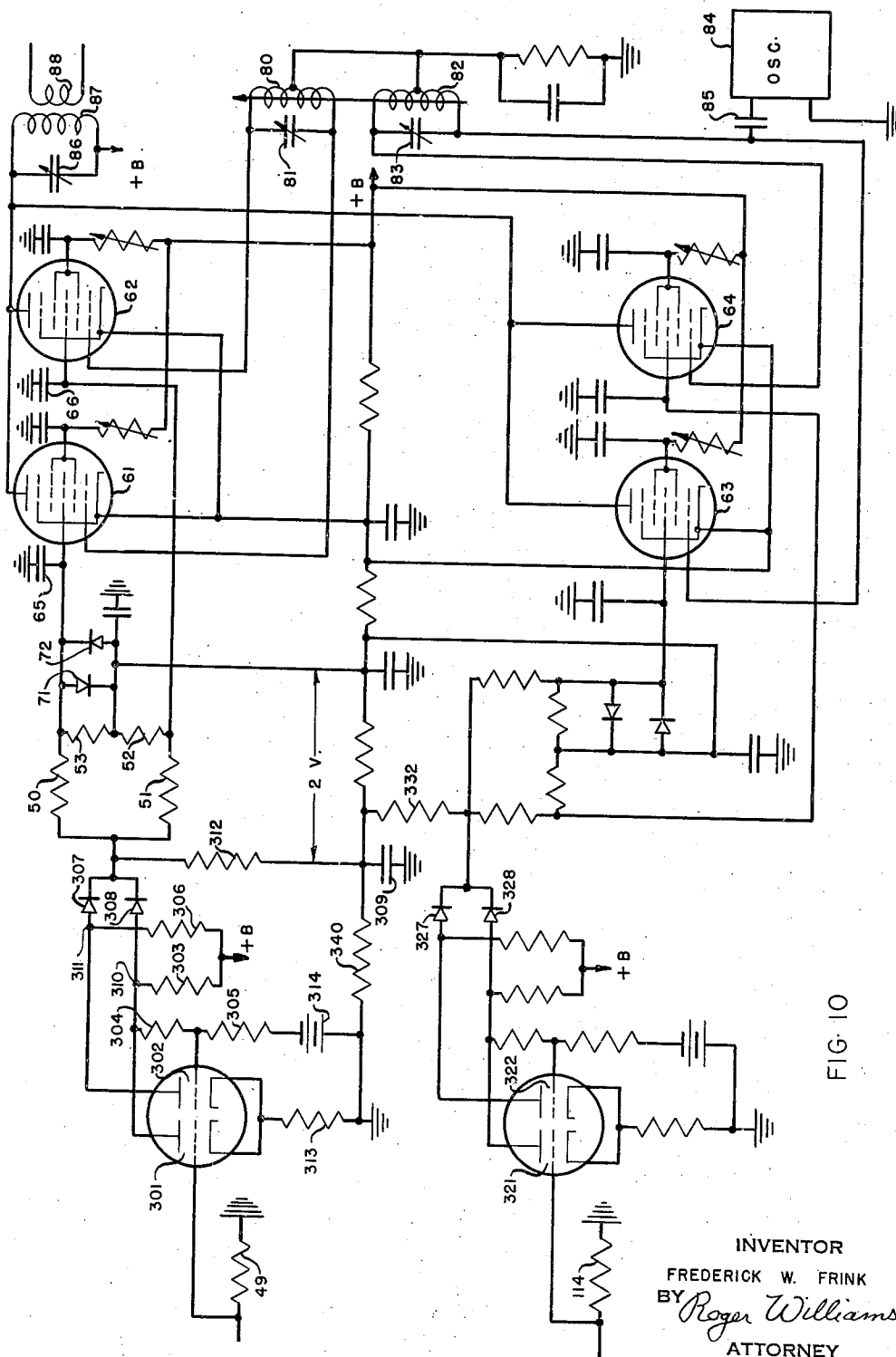
FIG. 9

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Filed Nov. 30, 1946

FREQUENCY MODULATION TRANSMITTER

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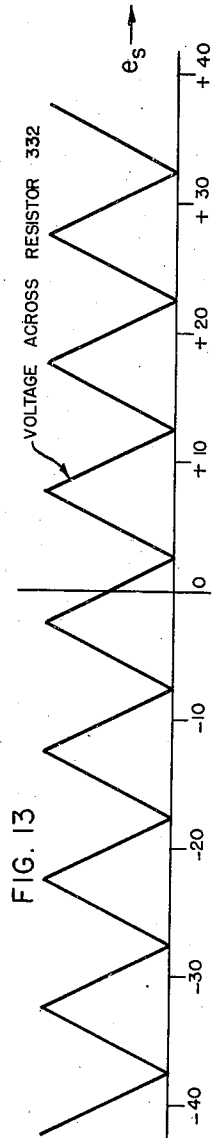
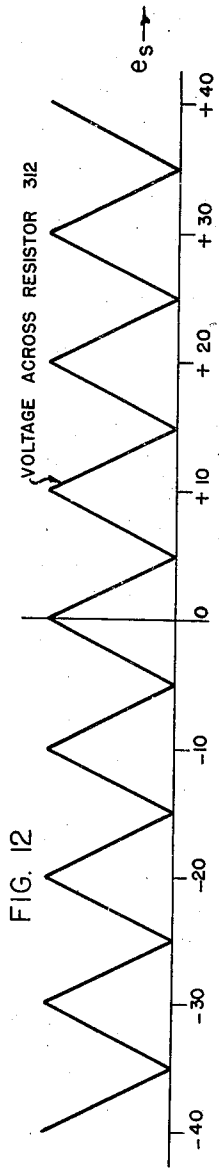
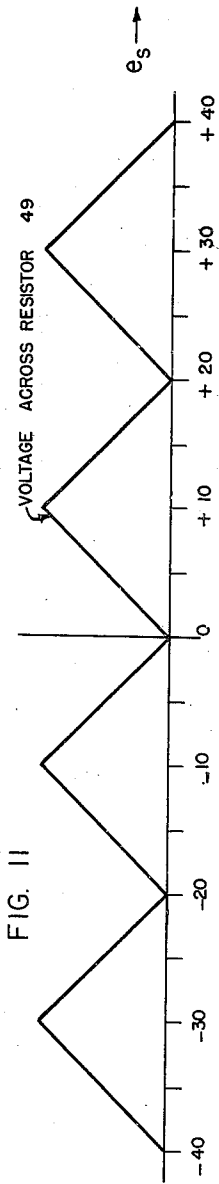
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FREQUENCY MODULATION TRANSMITTER

Filed Nov. 30, 1946

6 Sheets-Sheet 6



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UNITED STATES PATENT OFFICE

2,483,271

FREQUENCY MODULATION TRANSMITTER

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Application November 30, 1946, Serial No. 713,219

13 Claims. (Cl. 332-24)

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This invention pertains to a system and apparatus for the transmission of frequency-modulated radio signals. More particularly it pertains to radio transmitters of the type commonly referred to as amplitude to phase to frequency-modulated.

In transmitters of the type just referred to, the original modulation is of the amplitude type. This original modulation is caused to vary the phase of the carrier wave, such phase modulation giving rise to frequency-modulation, as is well known in the art.

A typical prior art system of this type passes the original audio-frequency through a corrective network having such characteristics that the output thereof is inversely proportional to the frequency of the audio-frequency voltage. The output of this network is supplied to a balanced modulator, where it modulates the radio-frequency current also supplied to this modulator, this latter being usually derived from a crystal oscillator. The modulated output of the balanced modulator is combined with unmodulated current from an identical source, after one or the other current has been shifted 90 degrees in phase, thus yielding phase modulation.

The system just described is capable of producing a base deviation approaching ± 90 degrees, but in practice it is impossible to extend the phase deviation to such a value, since extremely serious distortion would thereby occur. The fact that the phase angle is not directly proportional to the amplitude of the audio-frequency modulating signal is responsible for this distortion. In devices of the types in actual use, the extent of the deviation has been limited to values arbitrarily postulated as being the permissible maxima which can be used, while at the same time maintaining the distortion below a level which is considered permissible for a particular type of transmission. For example, high-fidelity broadcast transmitters usually limit the deviation to a maximum of ± 11.5 degrees, which is equivalent to ± 0.2 radian. Intelligible speech transmission may usually be obtained with deviations as great as ± 25 degrees, equivalent to ± 0.44 radian.

In frequency-modulation transmitters it is necessary to secure an ultimate deviation much greater than that corresponding to the values just given. In broadcast transmitters, there is usually desired at the ultimate output a 75-kilocycle deviation. In order to secure this, the output of the modulation system passes through an extensive cascaded chain of frequency-multipliers. For example, with a modulating frequency

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of 30 cycles per second, there initially results a frequency deviation of only 6 cycles per second, if the deviation be kept within the limits imposed by permissible distortion, as just stated, whence the need for extensive increase in deviation.

The employment of frequency-multipliers allows a crystal oscillator of 200 kc., for example, to be used as the primary frequency source, such an element being readily available. In order to secure the required output frequency, which latter is usually in the vicinity of 100 megacycles, the frequency-multipliers are employed. At the same time, the frequency multiplication thus obtained is still insufficient to yield the desired 75 kc. ultimate deviation, as the ultimate resting frequency simultaneously produced by the multiplication needed for 75 kc. deviation would be far above the order of 100 megacycles.

As one commonly used solution of this difficulty, the modulated signal is frequency-multiplied to a convenient value, then heterodyned to a lower value, which heterodyning does not reduce the absolute value of the deviation. Further frequency multiplication then simultaneously yields the desired ultimate resting frequency and the desired ultimate frequency deviation. However, such a solution poses several problems and introduces additional complications and difficulties. The need for extremely accurate control of more than one oscillator is a serious problem, and the various artifices which have been used to correct or compensate for oscillator frequency drift are characterized by the introduction of still further complexity in the apparatus, as typified by the well known dual-channel system.

The present invention overcomes the difficulties just recited by providing a phase modulator which produces an initial deviation so great that the number of frequency multipliers needed will be much less than the number ordinarily needed when a phase modulator of conventional type is used. Therefore, it will no longer be necessary to employ heterodyning, or dual-channel operation, while at the same time the distortion accompanying this increased phase deviation is kept at an extremely low value.

As one example, by the employment of apparatus according to this invention, initial phase deviation may extend to at least ± 11 radians, without introducing distortion above the maximum allowable, and theoretically this deviation may be extended to an even greater extent, if desired, although the necessary apparatus for effecting a greater range becomes relatively extensive. However, 11 radians' deviation allows

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the usually desired respective values, as above stated, of ultimate resting frequency and ultimate deviation to be obtained without recourse to heterodyning operations, and with a much lower degree of frequency multiplication than that usually required, which latter is in the neighborhood of 7 or 8 thousandfold.

One object of this invention is to provide a frequency modulation transmitter which will require the employment of a much smaller amount of frequency multiplication than has hitherto been found necessary in order to produce the requisite degree of ultimate phase deviation.

Another object of this invention is to provide a frequency modulation transmitting system in which an initial phase deviation many fold greater than that hitherto employed is used, while distortion is yet maintained below the maximum permissible in practice.

Yet another object of this invention is to provide a frequency-modulated transmitter in which complexity and consequent cost are reduced by reason of a substantial lessening of the degree of frequency multiplication required, while at the same time the standards required in practice with respect to distortion and ultimate resting frequency and phase deviation are met.

A further purpose of this invention is to provide a frequency-modulation transmitter in which heterodyning operations and the consequent employment of more than one oscillator are completely avoided.

A still further object of this invention is to provide a frequency-modulation transmitter in which the audio-frequency currents are so modified, and modulate the radio-frequency output of a single oscillator in such fashion, that modulation of the order of at least twenty times that permissible by hitherto used methods, may be obtained without substantial distortion.

An additional purpose of a variant form of this invention is to provide a transmitter of the type just identified, in which the audio-frequency currents are still further modified, so that the degree of ultimate modulation secured may be increased to an even greater value.

Other objects and features of the invention will be apparent from the following detailed description with reference to the accompanying drawing, in which:

Figure 1 is a schematic diagram of one embodiment of my invention, comprising Sheets A and B.

Figures 2, 3, 4, 5, and 6 are graphs showing the voltage variations occurring in certain circuits of the embodiment shown in Figure 1.

Figures 7, 8 and 9 are vector diagrams showing the phase conditions existing at certain times in one of the circuits of Figure 1.

Figure 10 is a schematic diagram of a portion of another embodiment of this invention, incorporating elements for producing additional phase deviation, and

Figures 11, 12 and 13 are graphs showing the variations of potential which occur at certain points in the embodiment shown in Figure 10.

Referring now to Figure 1, the audio-frequency signal voltage to be transmitted is passed through a compensating network similar to those which have already been used in the prior art frequency-modulation transmitters that employ phase modulation circuits. This compensating network consists of resistor 10 and capacitor 11, and the circuit constants are so chosen that the ratio of output voltage to input voltage is in-

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versely proportional to frequency, over the band of frequencies used, the reason for and function of this network being well known in the art of frequency modulation.

The output voltage of the compensating network is amplified by amplifier 12, which is a linear audio-frequency amplifier, and is then applied to the primary winding of transformer 13. This transformer has two substantially identical secondary windings, 14 and 15, each of which is tapped substantially at its midpoint. The midpoint of secondary winding 14 is connected to the positive end of battery 16, which has its negative end grounded, and the midpoint of secondary winding 15 is connected to the negative end of battery 17, which latter has its positive end grounded. Each of these batteries has a potential of about 40 volts, for the particular embodiment here shown. It is to be understood that the showing of batteries is purely exemplary, and that any other suitable potential source may be substituted therefor. This mode of connection allows a direct current supply of energy to be impressed upon voltage divider 18, as well as upon other voltage dividers hereinafter described, from a single potential source. At the same time, this particular connection allows such potential source to be grounded, or to have a relatively high capacity to ground, this last requirement being one which it would be difficult to meet, if individual batteries were employed for each voltage divider.

In the following description, it is assumed, for the sake of simplicity, that the form of the audio-frequency voltage is that of a sine wave at the maximum energy level.

A voltage divider 18, which consists of one or more resistors having a total of four taps, is connected from the upper end of secondary winding 14 to the upper end of secondary winding 15. Current from by-passed batteries 16 and 17 flows through this voltage divider, producing an overall D.-C. drop of about 80 volts, and the tap marked "O" is at D.-C. ground potential. The D.-C. potentials of the other taps on this voltage divider relative to ground are indicated by the respective numerals on the diagram. Voltage divider 19, which is similar to 18, connects the lower end of secondary winding 14 to the lower end of secondary winding 15.

The transformer secondary windings are wound in such a direction that the A.-C. potential of the upper end of secondary winding 14 relative to ground is equal, in magnitude and direction, to the A.-C. potential of the upper end of secondary winding 15 relative to ground. There is, therefore, never any A.-C. voltage applied across voltage divider 18 by transformer 13. Similarly, there is never any A.-C. voltage applied across voltage divider 19. All points on voltage divider 18 are, therefore, at the same A.-C. potential relative to ground, and a similar statement applies to 19.

The various taps on voltage dividers 18 and 19 are connected, through resistors 21 to 28 inclusive, to diodes 31 to 38 inclusive, and these diodes are connected to bus 41, which latter is in turn connected through resistor 49 to ground. To minimize the number of separate tubes required, four twin diodes may be used. Each twin diode consists of two independent diodes enclosed in the same envelope. However, wholly independent diodes may alternatively be used.

Let us suppose that the instantaneous voltage e_s existing across one half the respective sec-

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ondary winding 14 or 15 has just passed through zero, and is now increasing in such a direction that the upper end of 15 is positive with respect to the midpoint. Diode 31 immediately draws current through resistors 21 and 49, since the plate of this diode is at ground potential only when e_s is zero. It is assumed for illustrative purposes that the resistance of voltage divider 18 and the internal resistances of the various diodes are negligible compared with the resistances of resistors 21 to 28, and 49, inclusive. It is also assumed that the values of resistors 21 and 49 have been so chosen that when e_s has risen to an instantaneous value of 10 volts the voltage drop e_o across 49 will be 4 volts. If e_s continues to increase beyond 10 volts, diode 32 starts to draw current through resistors 22 and 49, because the transformer voltage e_s is then greater than the algebraic sum of e_o (4 volts) and the D.-C. cathode bias voltage of diode 32 (14 volts), which have opposing polarities so far as diode 32 is concerned.

The resistance value of resistor 22 is so chosen that the current through 22 builds up much faster than the current through 21, and by the time e_s has reached a value of 20 volts the current through 22 and 32 has reached a magnitude great enough to reduce the resultant current through 49 to zero, thus reducing voltage e_o to zero.

The method of determining correct resistance values for all of the resistors will be later explained, but at present it is merely assumed that when e_s increases above 20 volts, diode 33 starts drawing current, and this current builds up fast enough so that by the time e_s arrives at a value of 30 volts the voltage e_o across 49 will again have risen to a value of 4 volts. Furthermore, when e_s increases above 30 volts, diode 34 starts drawing current, and by the time e_s reaches a value of 40 volts this current has reached a high enough value so that the voltage e_o has again been reduced to zero.

Reference is now additionally made to Figure 2, where the variation of voltage e_o as a function of e_s has been plotted. From point O to point A on this graph, diode 31 is the only diode drawing current. From A to B, diodes 31 and 32 are both drawing current, from B to C diodes 31, 32, and 33 are drawing current, and from C to D diodes 31, 32, 33 and 34 are all drawing current. As voltage e_s increases from zero to 40 volts, the voltage e_o is always a linear function of e_s up to the point where an additional diode starts drawing current. The current drawn through the additional diode is always of such magnitude and direction that the slope of the curve of Figure 2 is changed from positive to negative, or vice versa.

Thus far the discussion has been limited to what happens during that period in which the upper end of secondary winding 15 is positive with respect to the midtap of this winding. During the other half cycle of e_s , when the upper end of 15 is negative with respect to the midpoint, diodes 31, 32, 33, and 34 do not draw any current, but diodes 35, 36, 37, and 38 function in substantially the same manner as previously described for diodes 31, 32, 33, and 34, and voltage e_o passes through substantially the same series of values that have been plotted in Figure 2 for negative values of e_s .

Resistors 50 and 51 have resistance values much greater than that of resistor 49, so that the magnitude of the voltage developed across 49

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is only slightly affected by any current flowing to ground through resistors 50 and 51. To compensate for these parallel currents it is necessary for 49 to have a resistance a little greater than 40,000 ohms, so that the resultant resistance from bus 41 to ground will be approximately 40,000 ohms. The ungrounded end of 49 is connected through resistor 50 to grid No. 3 (i. e. the third grid from the cathode) of vacuum tube 61, and is also connected through resistor 51 to grid No. 3 of vacuum tube 62. Capacitors 65 and 66 are small radio-frequency by-pass capacitors, connected to ground.

The effects produced by the voltages applied to the No. 3 grids of tubes 61 and 62, and the effects produced by rectifiers 71 and 72, hereinafter to be described, can best be understood after the functions of tubes 61, 62, 63, and 64 are first explained. Tubes 61, 62, 63, and 64 are of the type generally referred to as pentagrid converters. Grid No. 1 (i. e. the grid nearest the cathode) of 61, and grid No. 1 of 62 are connected to opposite ends of the tuned circuit composed of inductor 80 and condenser 81. Grid No. 1 of 63 and grid No. 1 of 64 are connected to opposite ends of the tuned circuit composed of inductor 82, and condenser 83. Radio-frequency voltage from a suitable oscillator 84 is applied to the tuned circuit composed of 82 and 83 through a very small capacitor 85, which latter couples this tuned circuit relatively loosely to the oscillator.

The tuned circuit composed of 80 and 81, and the tuned circuit composed of 82 and 83 are both resonant at the frequency of oscillator 84. Coils 80 and 82 are located in such a manner that there is inductive coupling between them, as here shown by the arrow, and the degree of this coupling is such that the total voltage developed across 80 is equal to that developed across 82. Furthermore, since the circuit composed of 80 and 81 is a resonant circuit, the total voltage consequently developed across 80 is approximately 90 degrees out of phase with the voltage across 82. The phase of the voltage across coil 80 can be adjusted so as to be exactly 90 degrees from that across coil 82 by slightly varying the capacitance of condenser 81.

Since the plates of tubes 61 and 62 are connected in parallel to the untuned circuit composed of condenser 86 and inductor 87, tubes 61 and 62 are equivalent to, and capable of functioning as a balanced modulator. If grid No. 3 of tube 61 is at the same potential as grid No. 3 of tube 62, no radio-frequency voltage will be produced across coil 87 and capacitor 86, because the radio-frequency voltage applied between grid No. 1 and cathode of tube 61 is 180 degrees out of phase with the radio-frequency voltage applied between grid No. 1 and cathode of tube 62, and the gain of 61 is the same as that of 62, this action being that well known in the art of balanced modulators.

If the potential of grid No. 3 of tube 61 is varied in a positive direction from its normal potential, the output currents of tubes 61 and 62 will no longer balance out, because the gain of 61 will be greater than that of 62, and a radio-frequency voltage will appear across 87. If the potential of grid No. 3 of tube 61 is varied in a negative direction from its normal potential, a radio-frequency voltage will appear across coil 87, but it will be 180 degrees out of phase with the voltage which would have been produced if the potential of grid No. 3 had been varied in a positive direction from normal. The effect of varying the potential of grid No. 3 of tube 62 is similar to that

of varying the potential of grid No. 3 of tube 61, except that, so far as the phase of the output voltage is concerned, the effect of raising the potential of grid No. 3 of tube 62 is the same as that of lowering the potential of grid No. 3 of tube 61.

In Figure 2 it was indicated that e_0 , the voltage across resistor 49, when plotted as a function of the transformer voltage e_s , has a triangular form. However this voltage is transmitted through the network composed of resistor 50, resistor 53, and copper-oxide rectifiers 71 and 72, before being applied between grid and cathode of tube 61. The copper-oxide rectifiers are of any suitable small instrument type having the characteristics hereinafter described. A plurality, for example eleven, rectifiers of this type are used in series to form rectifier unit 71, and eleven to form rectifier unit 72, and the two rectifiers are so connected that the polarity of 71 is opposite to that of 72. It is possible to substitute a single resistor, having similar non-linear characteristics for these rectifiers.

A D. C. bias voltage, of 2 volts, is applied between point P_2 and ground, by means of bypassed resistor 109, thus making point P_2 positive relative to ground. Therefore whenever the voltage e_0 across resistor 49 is 4 volts, point P_1 is positive with respect to point P_2 by an amount equal to 2 volts minus the voltage across 50. On the other hand, when the voltage across 49 is zero, point P_1 is negative with respect to P_2 by an amount equal to 2 volts minus the voltage drop across resistor 50. Because of the non-linearity of the copper-oxide rectifiers 71 and 72, the voltage e_1 between point P_1 and point P_2 varies as shown by the curve 301 of Figure 3, to which reference is now made, when e_0 is varied. The curvature of this latter curve is caused by the fact that the resistance of the copper-oxide rectifiers decreases as the voltage across them increases.

Voltage e_1 is applied between grid No. 3 and the cathode of tube 61. At the same time, the voltage e_2 developed across resistor 52 is applied between grid No. 3 and the cathode of tube 62, and e_2 is directly proportional to voltage e_0 , as indicated by curve 302 of Figure 3.

Since the excitation voltages applied to the No. 1 grids of tubes 61 and 62 from coil 80 are 180 degrees out of phase with each other, the radio-frequency output voltage appearing across coil 87, due to the action of modulators 61 and 62, is proportional to the difference between e_1 and e_2 . This difference has been plotted as curve 303 of Figure 3. All of the curves shown in Figure 3 were plotted from data obtained in actual measurements of a physical embodiment of this invention. To obtain the data for the $(e_1 - e_2)$ curve, a tube voltmeter was connected between points P_1 and P_3 , voltage e_0 was varied in small steps, and readings were taken of $(e_1 - e_2)$, which is equivalent to the voltage between point P_1 and point P_3 . Resistor 50 was 130,000 ohms, 51 was 470,000 ohms, 52 was 200,000 ohms, and 53 was 500,000 ohms. 71 and 72 each consisted of eleven "Conant" rectifiers, type BH, Series 160, connected in series. The circled points on the $(e_1 - e_2)$ curve of Figure 3 represent measured values of $(e_1 - e_2)$, and also represent substantially the values that $(e_1 - e_2)$ would have if $(e_1 - e_2)$ were a cosine function of e_0 , with maximum at $e_0 = +4$ and minimum at $e_0 = 0$.

Reference is now made to Figures 4 to 9, inclusive. It can be seen that $(e_1 - e_2)$ is substantially a cosine function of e_0 , which is the condi-

tion desired, as will later become apparent. Figure 4 shows $(e_1 - e_2)$ as a function of the transformer voltage e_s . Since e_0 is a linear function of e_s , except where reversals occur in the slope of the curve shown in Figure 2, and since $(e_1 - e_2)$ is substantially a cosine function of e_0 , then $(e_1 - e_2)$ is substantially a cosine function of e_s , as indicated in Figure 4.

Diodes 91 to 98 inclusive, together with voltage dividers 103 and 104, function in the same manner as do diodes 31 to 38 inclusive, except that diodes 91 to 98 are biased differently. Each tap on voltage divider 103 is located at a point 5 volts higher in potential than the corresponding tap on divider 18, and each tap on voltage divider 104 is located at a point 5 volts lower in potential than the corresponding point on divider 19. Because of this arrangement of biases, relative to one another, the voltage across resistor 114 varies in the manner shown in Figure 5, instead of in the manner shown in Figure 2. The curve of Figure 5 is substantially the same as that shown in Figure 2, except that it has been shifted to the left, so that when e_s is zero, the voltage across resistor 114 is 2 volts, instead of zero. Fig. 6 shows $(e_3 - e_4)$ as a function of e_s . Evidently, $(e_3 - e_4)$ is a sine function of e_s , instead of being a cosine function, as is $(e_1 - e_2)$ (Fig. 4). This causes a phase displacement between the two transmission channels.

Now considering the conditions existing in the common plate circuit of tubes 61 and 62 at the instant when e_s is zero, reference to Fig. 4 shows that $(e_1 - e_2)$ is at a negative peak, so that the resultant radio-frequency current, produced wholly by tubes 61 and 62 in the coil 87 may be represented by vector I_1 in Fig. 7. At the same instant $(e_3 - e_4)$ is zero, so tubes 63 and 64 are not producing any radio-frequency plate current. Therefore a vector I_2 , representing the current due to these last tubes, is at this instant, zero. The resultant radio-frequency plate current is therefore equal in magnitude and phase to I_1 . At the instant when $e_s = +2.5$ volts the output current I_1 of tubes 61 and 62 is equal in amplitude to the output current I_2 of tubes 63 and 64, because $(e_1 - e_2) = (e_3 - e_4)$, as a comparison of Fig. 4 and Fig. 6 will show. The resultant R.-F. plate current I_r is, therefore, as shown in Fig. 8. When voltage e_s arrives at a value of 5 volts, $(e_1 - e_2) = 0$, and $(e_3 - e_4)$ is a maximum. The resultant current is therefore equal to I_2 , which has the phase and value indicated in Fig. 9, while I_1 is now zero. As voltage e_s continues to increase, the resultant current I_r is continuously shifted in phase, in a clockwise direction, until finally, when e_s becomes +35 volts a total phase shift of $1\frac{3}{4}$ cycles, or 630 degrees, has been produced.

It is obvious that in order to make the resultant vector I_r rotate through 360 degrees or more without any change in length, it is necessary that each of the component vectors I_1 and I_2 shall in turn undergo a reversal of direction, corresponding to a phase reversal, when the resultant passes from each quadrant into the next succeeding quadrant, during the course of the modulation cycle. These reversals are able to occur because of the use of two balanced modulators, each of whose R.-F. output currents can be reversed in phase by reversing the polarity of voltage $(e_1 - e_2)$ or voltage $(e_3 - e_4)$.

Thirty-five volts is the maximum value which the voltage e_s can have without causing distortion in the transmitted signal, in the particular

embodiment here shown. When e_s returns from +35 volts to zero, the phase of the resultant current I_r varies in a counterclockwise direction instead of a clockwise direction. During that part of the cycle in which e_s is negative, the operation is similar to that for the positive half cycle. In either case, varying e_s in a positive direction always causes the phase of I_r to vary in a clockwise direction, and vice versa.

Coil 83 can of course be connected to an amplifier or a frequency multiplier and the radio-frequency signal voltage can be amplified, or the frequency can be multiplied, in substantially the same manner as in other frequency-modulation transmitters.

As previously mentioned, when the amplitude-phase-frequency system of modulation is used in broadcast transmitters, the phase deviation produced by the modulation system is usually limited to about ± 11.5 degrees, or ± 0.2 radian, to avoid any serious distortion. In my system, the maximum phase deviation that can be produced by the embodiment here shown, without serious distortion, is ± 630 degrees, or ± 11 radians, which is about 54.8 times as great as the figure of 11.5 degrees given above. It is obvious that several stages of frequency multiplication can be eliminated by the use of my system. Furthermore the large phase deviation obtained in my modulation system makes it unnecessary to use heterodyning or dual channel operation. For example, let it be supposed that the lowest audio-frequency to be transmitted is 20 cycles/sec. The frequency deviation produced by an 11-radian phase deviation at this modulating frequency would be 20×11 , or 220 cycles/sec. If the average output frequency of the transmitter is to be 100,000 kc. and the frequency of the oscillator is 200 kc., a total frequency multiplication of 500:1 is needed. This frequency multiplication will increase the frequency deviation to a value of 220×500 , or 110,000 cycles/sec., which is more than enough to meet the present requirements in the art, as fully described in an earlier portion of this specification.

The method of selecting resistance values for resistors 21 to 28 is now explained. The values of 21 and 49 are arbitrarily selected as 60,000 ohms and 40,000 ohms respectively, thus causing the voltage e_o across 49 to be 4 volts when e_s is +10 volts. It has already been decided that when e_s has risen to +20 volts, the current through 49 shall drop to zero, so that e_o will be zero (see Fig. 2). At this point the voltage across resistor 21 will be 20 volts, and the current through 21 will be $20/60,000 = 0.333$ milliamperes. Diode 32 starts drawing current when e_s is +10 volts, and by the time e_s is +20 volts the voltage across resistor 22 is $20 - 14 = 6$ volts. Since the current drawn through 22 must at this time be equal to that through 21, in order to obtain zero resultant current through 49, the resistance of 22 must be $6/(0.333 \times 10^{-3}) = 18,000$ ohms.

As indicated in Fig. 2, the network is to be designed so that when e_s is +30 volts the voltage e_o will be 4 volts. When e_s rises above +20 volts, diode 33 starts drawing current, and when e_s is +30 volts the voltage across 23 is $30 - 20 = 10$ volts. At the same time, the current through 21 is $(30 - 4)/60,000 = 0.4333$ milliamperes, and the current through 22 is $(30 - 14 + 4)/18,000 = 1.1111$ milliamperes. To obtain a voltage of 4 volts across resistor 49 we must have a resultant cur-

rent of 0.1000 milliamperes through 49. To obtain this resultant current, the current through 23 must be $0.1000 - 0.4333 + 1.1111 = 0.7778$ milliamperes. Since the voltage across 23 is 6 volts, the resistance of 23 must be $6/(0.7778 \times 10^{-3}) = 7,718$ ohms.

When e_s rises above +30 volts, diode 34 starts drawing current, and by the time e_s has reached a value of +40 volts the voltage across 24 is $(40 - 34) = 6$ volts. At the same time the current through 21 is $40/60,000 = 0.6667$ milliamperes, the current through 22 is $(40 - 14)/18,000 = 1.4444$ milliamperes, and the current through 23 is $(40 - 20)/7,718$ which equals 2.5913 milliamperes. The resultant current through 49 must be zero, and therefore, the current through 24 must be $0.6667 - 1.4444 + 2.5913 = 1.8136$ milliamperes. To obtain this current it is necessary that 24 must have a resistance of $6/(1.8136 \times 10^{-3}) = 3,308$ ohms.

The calculations made so far have pertained to diodes 31, 32, 33, and 34, which draw current during the half cycle when the upper end of secondary winding 15 is positive with respect to the midpoint. During the other half cycle the operation is similar to that on which the above calculations were based, but diodes 35, 36, 37, and 38 draw current instead of the above mentioned diodes. Therefore resistor 25 should be equal in value to 21, 26 should be equal to 22, 27 should be equal to 23, and 28 should be equal to 24. The resistors in series with diodes 31 to 38 inclusive should be equal in value to the corresponding resistors used in series with diodes 31 to 38 inclusive.

In determining the values of resistance as above calculated, certain factors which slightly affect the result are disregarded for the sake of simplicity. One of these is the contact potential of the diodes. This contact potential, which is usually only a fraction of one volt, produces the same effect as though a battery of this voltage were connected in series with the diode. Other factors which affect the result are the respective voltage drops produced by the diode currents in flowing through voltage dividers 18, 19, 103 and 104, although it is intended that the resistors used in the voltage dividers shall have a very much lower resistance than resistors 21 to 28 inclusive. To allow for these factors and any other factors affecting the results, it is best to make all of the series resistors adjustable, and all taps on the voltage dividers adjustable. After all resistors and taps have been adjusted according to the above calculated values, an alternating voltage should be applied to the input of amplifier 12, and adjusted in magnitude so as to produce a voltage of 80 volts peak across each entire secondary winding 14 or 15. The voltage across these windings should be applied to the horizontal deflection plates of a cathode-ray oscilloscope, or other equivalent device, using amplification if necessary. The voltage e_o across resistor 49 should be amplified, and applied to the vertical deflection plates of the same oscilloscope. The pattern on the screen of the oscilloscope should then appear the same as shown in Fig. 2. If adjacent peaks are not all separated by the same horizontal distance, or are not of equal height, the necessary corrections can be made by adjusting the taps on voltage dividers 18 and 19, or by adjusting the resistors in series with the various diodes.

Because of the fact that the various diodes do not all draw current during the entire half

cycle, the load on the secondary windings of transformer 13 is not a linear load. This tends to prevent voltage e_s from having exactly the same wave form as the input voltage of amplifier 12. This effect can be greatly reduced by loading transformer 13 with linear resistors 215 and 216. Any remaining effects produced by the non-linear load can be compensated by adjusting the taps on the voltage dividers, and the resistors in series with the diodes.

The calculated values of resistance for the resistors in series with the diodes, and the calculated position of the taps on the voltage dividers, should be considered as first approximations only, since changes may have to be made when the above described test is performed, and the final adjustments may be considerably nullified by the high resistances connected in series therewith.

Obviously, it would be possible to use copper-oxide, selenium, or other similar rectifiers, instead of the diodes here shown, although the resistance values of the various resistors used in various parts of the associated network might have to be somewhat different, because of the fact that the rectifiers, for example the copper-oxide or selenium type, do not have exactly the same characteristics as do thermionic diodes, as well known in the art. Such rectifiers need not be strictly linear, as variations are substantially nullified by the high resistances connected in series therewith.

Tubes 61, 62, 63, and 64, have been represented in Fig. 1 as pentagrid converter tubes. For example, tubes of the well known type 6SA7 or of the type 6BE6 could be used. Instead of using pentagrid converter tubes it would be possible to use various other types of tube, for example the type 6SJ7 sharp-cutoff pentode. In using this last type of tube the modulating voltage can be applied to the control grid (i. e., the grid nearest the cathode), while the oscillator voltage can be applied to the screen grid thereof.

Elements not specifically described in the lower transmission channel, correspond to those shown in the upper channel, and detailed description thereof is therefore not needed. The various grounds, bias resistors, by-pass condensers, etc. are likewise not discussed in detail, for purposes of clarity; as their respective functions will be apparent to one skilled in the art.

Reference is now made to Fig. 10, a schematic showing of a portion of an alternative system which makes it possible to produce an even greater phase deviation than that produced by the circuits of Fig. 1. The portions not shown in Fig. 10 may be identical with those illustrated in Fig. 1. Resistors 49 and 114 correspond to the resistors having the same reference numerals in Fig. 1. The voltages produced across these resistors have the same form as the voltages across the corresponding resistors in Fig. 1, i. e. they are as shown in Fig. 2 and Fig. 5. The circuits and elements used for producing these voltages across resistors 49 and 114 may be identical with those shown in Fig. 1, and operate in the same manner as already described in connection with Figs. 1, 2, and 5. For purposes of clarity they are omitted in Fig. 10.

The voltage across resistor 49 is amplified by triode 301, which is a linear amplifier, and is reproduced in amplified form across resistor 303. Some of the voltage across resistor 303 is applied, through the voltage divider consisting of

resistors 304 and 305, to the grid circuit of triode 302. This voltage is reproduced, in amplified form, across resistor 306, but the voltage across resistor 306 is 180 degrees out of phase with the voltage across resistor 303, because of the phase-inverter action of triode 302. The resistance values of resistors 304 and 305 are chosen so that the voltage produced across resistor 306 is equal in magnitude to that produced across resistor 303. While triodes 301 and 302 are shown as located within a single envelope, it is to be understood that discrete tubes may be employed.

The voltage of bias battery 314 and the values of the various resistors are also chosen so that when the voltage across resistor 49 is halfway between zero and maximum, point 310 will be at the same potential as point 311. Since these values depend, inter alia, upon tube characteristics, they must be ascertained, in any given case, by methods well known in the art. The D.-C. potential of point 309 is adjusted in a similar manner, so that it is equal to the potential of points 310 and 311 at the instant when the potential of point 310 is equal to the potential of point 311. Resistor 349 forms part of the network used to secure the proper voltage relationships. This enables rectifiers 307 and 308 to function as a full-wave rectifier, so that the voltage across output resistor 312 has the form shown in Fig. 12. Fig. 11, which has been plotted directly above Fig. 12 for purposes of comparison, shows the voltage across input resistor 49. The number of peaks in Fig. 12 is twice as great as the number in Fig. 11. Rectifiers 307 and 308 are preferably, though not necessarily, of the germanium crystal type, such as the type 1N34.

The network composed of resistors 51, 52, 53, and 54, and rectifiers 71 and 72, functions in the same manner as the similar network in Fig. 1, producing voltages having the same form as voltages e_1 and e_2 of Fig. 1. These voltages are applied to the No. 3 grids of balanced modulators 61 and 62, which likewise function in the same manner as the corresponding modulators of Fig. 1.

The action of triodes 321 and 322, rectifiers 327 and 328, and modulator tubes 63 and 64, is the same as that of the corresponding elements in the upper channel, i. e., the same as 301, 302, 307, 308, 61, and 62, respectively. The rectifier network between resistor 322 and tubes 63 and 64 may be substantially identical in structure and performance with that used in the upper channel, and therefore is not described in detail. Fig. 13 shows the voltage across resistor 322. It can be seen that the voltage across resistor 322 is displaced by a quarter cycle from the voltage across resistor 312. Because of the frequency-doubling action produced by rectifiers 327 and 328 it is necessary that the voltage across resistor 114 be displaced from that across resistor 49 by only one-eighth cycle, instead of one-quarter cycle. This may be accomplished by adjusting the circuit elements so as to secure certain biases different from those indicated in Fig. 1. Specifically, on voltage divider 103 the points in Fig. 1 marked -15 v., +5 v., +19 v., and +39 v., should now have potentials of -17.5 v., +2.5 v., +16.5 v., and +36.5 v., respectively. Similarly, on voltage divider 104 the points in Fig. 1 marked -25 v., -5 v., +9 v., and +29 v., should now have potentials of -22.5 v., -2.5 v., +11.5 v., and +31.5 v., respectively.

Since the number of peaks in Fig 12 and Fig. 13 is twice as great as the number in Figs. 2 and 5, it is evident that for an audio-frequency voltage e_s having a peak value of 35 volts, the total phase deviation in the radio-frequency current in coil L-3 will be twice as great, when the system shown in Fig. 10 is used, as it is when the system shown in Fig. 1 is used.

It will be obvious to one skilled in the art that by adding more amplifiers similar to 301, 302, 321, 322, and more rectifiers similar to 307, 308, 327 and 328, together with their associated resistors, it is possible to introduce additional rectifications which, in cascade, will result in still further phase deviation of the transmitted current.

From the foregoing description it can be seen that the essential points involved in this invention may be summarized as follows.

This invention has the advantage of providing a frequency-modulated transmitter of the phase modulation type in which the amount of frequency multiplication is greatly reduced from that found necessary in the prior art. Additionally, the necessity of frequency reduction by heterodyning is obviated.

The advantages just enumerated are secured by the employment of devices which allow the initial phase deviation to extend over a substantially unlimited range, instead of confining it to a maximum of about $\frac{1}{2}$ radian, as has been done in the prior art. This increased range is secured without transcending the degree of distortion considered allowable.

In the system of this invention the audio-frequency modulating signals are modified so as to act upon the two radio-frequency circuits carrying currents 90° out of phase with one another, in two distinctive manners, whereby the distortion which occurs when a balanced modulator of the conventional type is employed is substantially eliminated, even though the phase shift may reach a value of many radians.

Expressed in other terms, this system simultaneously modulates the two vectors involved in modulation, in such fashion that distortion is substantially absent. Furthermore, when the resultant of these two vectors leaves the first quadrant, the modulation undergoes a corresponding alteration. This same action takes place as the resultant of the vectors enters each succeeding quadrant, so that there is no theoretical limit to the number of times that the resultant of the vectors can describe a complete 360° revolution.

While the two balanced modulators are here shown as excited 90° out-of-phase with one another, by high frequency oscillator 84, it will be evident that both modulators could be excited in phase with each other, provided that the respective outputs of the modulators are shifted relatively to one another, so as to cause the desired 90° out-of-phase relationship, before these two outputs are combined with each other. The equivalence of these alternative modes of connection is well known in the art.

While there have been shown and described certain embodiments of this invention it is to be understood that many variations thereof will be apparent to those skilled in the art and accordingly the scope of this invention is limited only by the hereunto appended claims.

What is claimed as new and desired to be secured by Letters Patent of the United States, is:

1. In combination, an oscillator, a first balanced modulator receiving voltage from said os-

cillator, a second balanced modulator also receiving voltage from said oscillator, means for causing said last voltage to be ninety degrees out of phase with the first mentioned voltage, a unitary source of audio-frequency signalling voltage, means for reversing the phase of the output current of said first balanced modulator in accordance with a function of the magnitude of said signaling voltage at a plurality of differing values, means for subsequently reversing the phase of the output current of said second balanced modulator in accordance with a similar function, means for again reversing the phases of the output currents of the balanced modulators but in opposite sequence to the previous reversals, and means for combining the output currents of said balanced modulators.

2. In combination, an oscillator, a first balanced modulator receiving voltage from said oscillator, a second balanced modulator receiving voltage from said oscillator, means for causing said last voltage to be ninety degrees out of phase with the first mentioned voltage, a unitary source of audio-frequency signalling voltage, means for producing a series of reversals in the phase of the output currents of said balanced modulators in accordance with a function of the magnitude of said audio-frequency signaling voltage when said voltage is altering in value upon a single side only of the zero line, said reversals occurring in a predetermined sequence with the reversals of said first balanced modulator occurring at different times from the reversals of said second balanced modulator, means for reversing the sequence in which said reversals occur, and means for combining the output currents of said balanced modulators.

3. In combination, an oscillator, a first balanced modulator receiving voltage from said oscillator, a second balanced modulator receiving voltage from said oscillator, means for causing said last voltage to be ninety degrees out-of-phase with the first mentioned voltage, a source of signal voltage, means for causing the amplitude of the output current of said first balanced modulator to vary in proportion to a sine function of the instantaneous magnitude of said signal voltage, means for causing the amplitude of the output current of said second balanced modulator to vary in proportion to another sine function of the instantaneous magnitude of said signal voltage, said other sine function being displaced in phase from said first mentioned sine function, and means for combining the output currents of said balanced modulators.

4. A frequency modulated transmitter, including means for producing audio-frequency voltage, means for separating said voltage into two channels, means in the first channel for transforming said voltage into a first wave form triangular with respect to the instantaneous magnitude of said audio-frequency voltage, means in the second channel for transforming said voltage into a second triangular wave form, displaced in phase relationship with respect to said first triangular wave form, means in each channel for transforming the respective triangular wave forms therein into wave forms sinusoidal with respect to the instantaneous magnitude of said audio-frequency voltage, two balanced modulators each modulated by the output of one of said channels, means for feeding carrier voltage to both said modulators, means for withdrawing the respective outputs of both said modulators, means for rendering one output substantially

90° out-of-phase with respect to the other output, and means for combining both said out-of-phase outputs.

5. A transmitter according to claim 4, in which said means for transforming the respective voltages into triangular wave form comprises a plurality of rectifiers, means for biasing each rectifier to a different voltage, said rectifiers being oppositely disposed and being connected to a common load whereby the number of complete triangular wave forms per cycle of the audio-frequency voltage is a function of the amplitude of the audio-frequency voltage feeding said rectifiers.

6. A transmitter according to claim 4, in which said means for transforming said triangular wave forms into sinusoidal forms comprises a plurality of rectifiers having non-linear characteristics, connected in oppositely poled pairs.

7. A transmitter according to claim 4, additionally including means in each channel for doubling the number of cycles of said triangular wave form voltage, said last means comprising phase inversion means, two rectifiers, each located in one output circuit of said phase inversion means, so as to be 180° out of phase with one another, and means for biasing said last rectifiers, whereby they function alternately.

8. The method of frequency modulation which includes the steps of producing two carrier currents 90° out of phase with each other, producing an audio-frequency voltage, transforming said audio-frequency voltage into two voltages, the respective magnitudes of which are linear functions of said audio-frequency voltage, the respective slopes of said linear functions reversing in algebraic sign at a given number of differing predetermined voltage magnitudes of said audio-frequency voltage, transforming said two voltages into respective sine functions which have peaks occurring at different moments, modulating each of said carrier currents by a different one of said two voltages, and combining said two modulated carrier currents.

9. The method of frequency modulation according to claim 8, including the additional step of doubling the number of reversals of said linear functions with respect to the number of voltage magnitudes, whereby the phase deviation secured in said combined modulated carrier currents is likewise doubled.

10. A transmitter according to claim 4, in which said means for transforming said triangular wave forms into sinusoidal forms comprises a resistor formed of material having non-linear characteristics.

11. The method of frequency modulation which includes the steps of producing two carrier currents 90° out of phase with each other, producing an audio-frequency voltage, transforming said audio-frequency voltage into two voltages each of which has a rate of change whose magnitude is proportional to the rate of change of said audio-frequency voltage, reversing the algebraic sign of the rate of change of each of said voltages at predetermined magnitudes of said audio-frequency voltage, said predetermined magnitudes being different for each of said two voltages, transforming said two voltages into respective sine functions, modulating each of said carrier currents by a different one of the two transformed voltages, and combining the two modulated carrier currents.

12. A combination according to claim 1, in which said means for reversing the phase of the output of said first balanced modulator include a plurality of differently biased rectifiers and a plurality of non-linear resistances.

13. A combination according to claim 2 wherein said means for producing a series of reversals in the phase of the output current of at least one of said balanced modulators include an electrical resistance and rectifier network yielding an output voltage which is a periodic function of the magnitude of the input voltage thereof, means for changing said output voltage to a sine function of said input voltage, and means for applying said changed output voltage to one of said balanced modulators.

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