

Feb. 18, 1941.

G. GUANELLA

2,231,997

FREQUENCY DISCRIMINATOR

Filed June 14, 1938

3 Sheets-Sheet 1

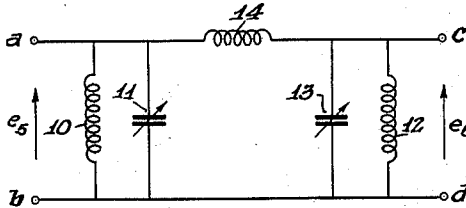


Fig. 1

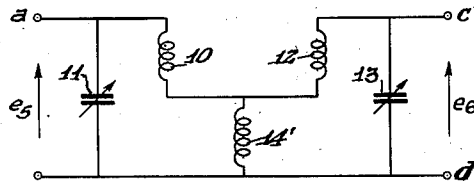


Fig. 2

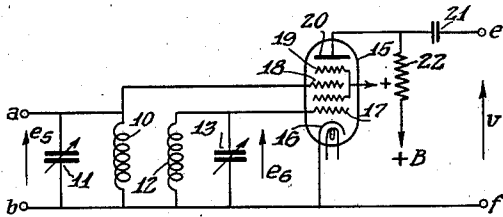


Fig. 3

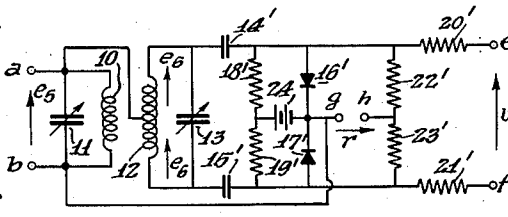


Fig. 4

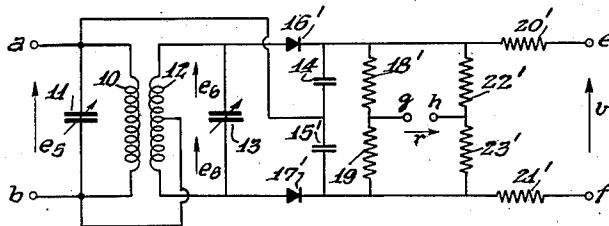


Fig. 5

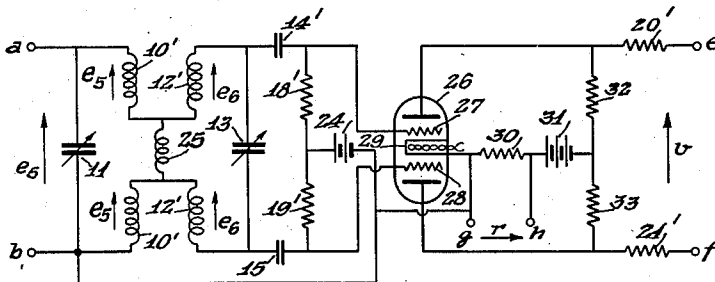


Fig. 6

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3 Sheets-Sheet 2

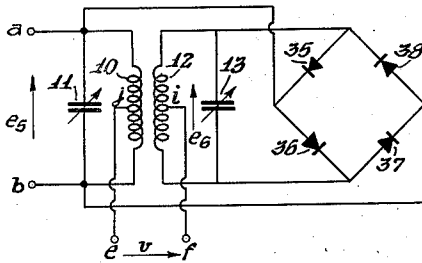


Fig. 7

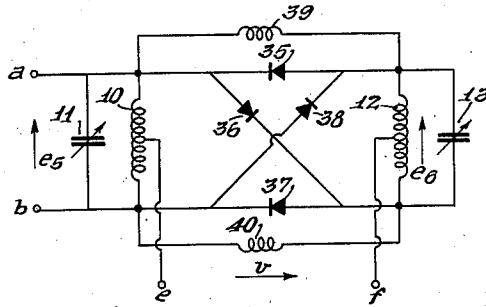


Fig. 8

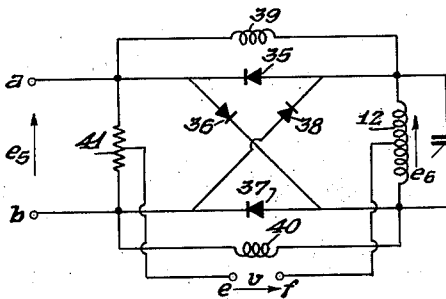


Fig. 9

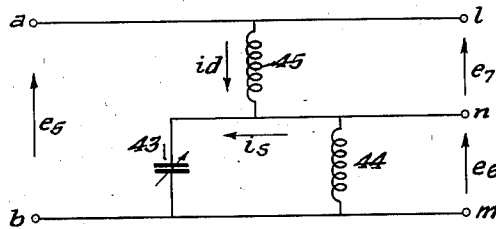


Fig. 10

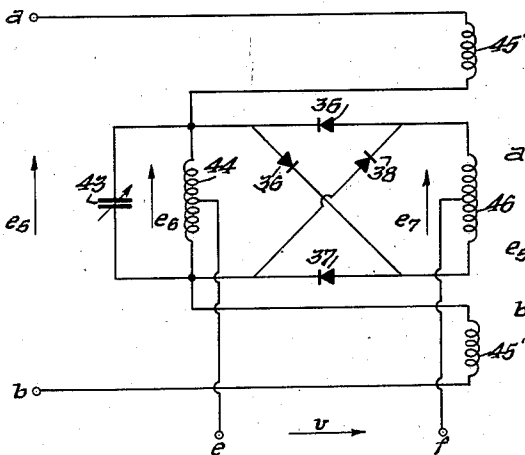


Fig. 11

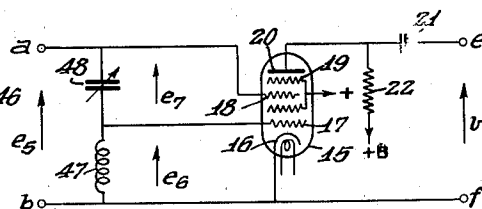


Fig. 12

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3 Sheets-Sheet 3

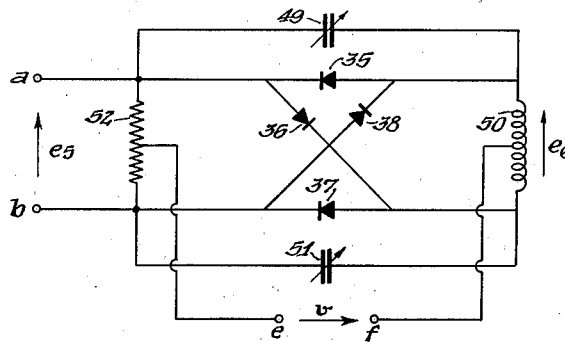


Fig. 13

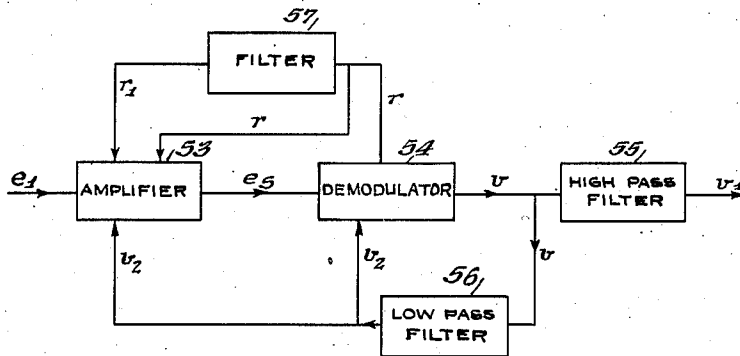


Fig. 14

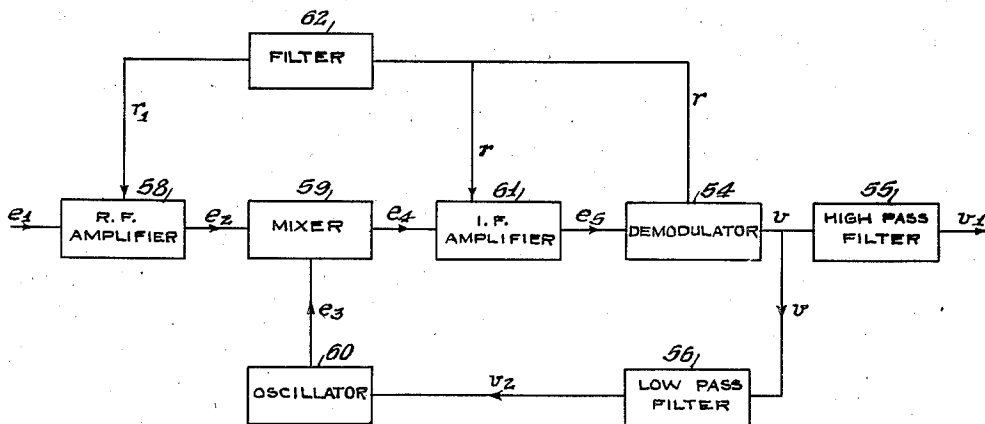


Fig. 15

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UNITED STATES PATENT OFFICE

2,231,997

FREQUENCY DISCRIMINATOR

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to Radio Patents Corporation, New York, N. Y.,
a corporation of New York

Application June 14, 1938, Serial No. 213,589
In Switzerland July 11, 1937

2 Claims. (Cl. 250-27)

This invention relates to new and improved means for and a method of receiving a high frequency carrier wave modulated in phase or in frequency by intelligence to be transmitted. The term intelligence is to include all kinds of signals such as audio, video and other signals, or impulses and the like to be transmitted from one point to another by the aid of a carrier wave.

As is well known, a frequency or phase modulated high frequency carrier wave may be regarded as a sine wave of substantially constant amplitude and continuously variable instantaneous frequency. In demodulating a wave of this type, that is in reproducing the low frequency signals or intelligence transmitted by the aid of such a wave, it is necessary to generate a potential which varies in dependence upon the instantaneous deviation of the frequency of the wave from a predetermined average (carrier) frequency. Several methods have been proposed to accomplish this purpose.

Thus, it has been suggested to provide a resonant circuit detuned relatively to carrier frequency of the received high frequency wave in such a manner that all the instantaneous frequencies fall within the uniformly curved or practically straight portion of one branch of the resonance curve of the resonant circuit. In this manner the frequency modulation is converted into an amplitude modulation whereupon a demodulation may be carried out in the usual manner by means of any one of the known detectors or demodulating arrangement for receiving amplitude modulated carrier waves.

According to the present invention demodulation of a phase or frequency modulated high frequency wave is effected by comparing the phases of the high frequency potentials or currents at different points such as in the input and output of a resonant circuit or network. Such a network may comprise two mutually coupled tuned circuits one of which is excited by the signal waves to be demodulated. In carrying out the invention the arrangement is preferably such that the input and output phases differ exactly by 90° for waves having a frequency corresponding to the resonant (carrier) frequency of the circuit, and deviate in either sense from a 90° mutual phase angle according to the sense and in proportion to the amount of detuning of the impressed frequency relative to the resonant frequency of the circuit. Thus, if the circuit is exactly tuned to the carrier frequency of the waves being received, the input and output phases

will be in quadrature relation for the unmodulated or carrier wave, but the relative phase angle will vary if the impressed carrier frequency varies in dependence upon variations of a lower frequency or modulating signal being transmitted. The input and output potentials of varying mutual phase differences are then applied to a device or system adapted to generate an output potential varying in dependence upon the mutual phase angle between the applied potentials, that is according to the low frequency signal wave, or intelligence impressed upon the carrier wave by frequency modulation.

The nature of the invention and the manner of its operation will become more apparent from the following detailed description of several specific embodiments taken in conjunction with the accompanying drawings forming part of this specification and throughout which like reference characters indicate like parts.

In the drawings,

Figures 1 and 2 illustrate resonant circuits suited for effecting phase rotation for the purpose of the invention,

Figure 3 shows a modified resonant circuit with a multi-grid vacuum tube as a demodulating device for carrying out the invention,

Figure 4 is a modification of Figure 3 embodying a rectifier bridge as a demodulating device,

Figure 5 is a modification of Figure 4,

Figure 6 is a further modification employing vacuum tube rectifiers,

Figures 7, 8 and 9 illustrate further modifications of the invention embodying a modulator or rectifier bridge circuit as a demodulating device,

Figure 10 shows a further resonant circuit for effecting phase rotation suited for the purpose of the invention,

Figure 11 shows a combination of a circuit according to Figure 10 embodying a rectifier or modulator bridge as a demodulating device,

Figure 12 illustrates a further resonant circuit in combination with a vacuum valve demodulator,

Figure 13 is a modification of Figure 12 employing a rectifier bridge as a demodulating arrangement,

Figures 14 and 15 illustrate in block diagram form improved systems for receiving frequency or phase modulated high frequency waves in accordance with the invention.

The novel features of the invention and its function will be further understood from the following theoretical discussion.

Let it be assumed that f_0 is the average (carrier) frequency of a high frequency wave being received from which the instantaneous frequency f deviates by an amount f_1 . This frequency deviation f_1 in case of frequency modulation is representative of the instantaneous value u of the low frequency or modulating signal wave (intelligence) being transmitted and is related therewith according to a definite, preferably linear law as follows:

$$f_1 = Au \quad (I)$$

wherein A is a constant.

The phase shift b effected by the phase rotating resonant circuit may then be expressed as follows:

$$b = b_0 + b_1 \quad (II)$$

wherein b_0 represents the normal relative phase angle for the frequency f_0 ($\pm 90^\circ$) while b_1 represents the instantaneous phase deviation from this normal angle depending upon the frequency deviation f_1 and being functionally related to the latter as follows:

$$b_1 = g(f_1) = g(Au) \quad (III)$$

wherein g signifies the functional relation between b_1 and f_1 .

Let it be further assumed that $a = a_0 a_1$ represents the amplitude transmission ratio or propagation factor of the phase rotating circuit, that is the ratio of the output amplitude to the input amplitude whereby a_0 corresponds to the average (carrier) frequency f_0 and a_1 corresponds to the frequency deviation f_1 . Accordingly, there exists the following relationship:

$$a_1 = h(f_1) = h(Au) \quad (IV)$$

wherein h signifies the functional relation between a_1 and f_1 .

As a result the potential e_s (see Figure 1) obtained at the output of the phase rotating circuit is phase shifted relative to the input potential e_5 applied to the input of the phase rotating circuit by an amount $b = b_0 + b_1$. Furthermore, the amplitude E_6 of the high frequency potential e_6 differs from the amplitude E_5 of the input high frequency potential e_5 by an amount $a = (a_0 a_1)$.

By mutual modulation or combination in a non-linear device of the two potentials e_5 and e_6 there is obtained a resultant product potential containing a low frequency component v which varies in accordance with the low frequency or modulating signal and represented by the following expression:

$$\begin{aligned} v &= K E_5^2 a \cos(b_0 + b_1) \\ &= V a \sin b_1 \\ &= V h(Au \sin g(Au)) \end{aligned} \quad (V)$$

wherein K and V are constants and $b_0 = 90^\circ$ as above pointed out. By suitably choosing the propagation characteristics g and h of the phase rotating circuit as a function of the frequency, the desired linear relation

$$v = k u \quad (VI)$$

may be obtained between v and the frequency deviations u resulting in a linear demodulating of the frequency modulated high frequency wave.

A demodulation is further possible by separate linear rectification of the sum and the difference potentials ($e_5 + e_6$ and $e_5 - e_6$ respectively) and subsequent differential combination of the rectified potentials (balanced demodulation). Assuming that $a = 1$ for the entire range of frequencies

employed, so that $E_5 = E_6 = E$, the amplitude of the sum potential ($e_5 + e_6$) will be as follows:

$$E_7 = 2E \cos(45^\circ - b_{1/2}) \quad (VII)$$

and the amplitude of the difference potential ($e_5 - e_6$) will be as follows:

$$E_8 = 2E \cos(45^\circ + b_{1/2}) \quad (VIII)$$

By rectification of E_7 and E_8 the peak voltages obtained are as follows:

$$E_7 \sqrt{2} \text{ and } E_8 \sqrt{2}$$

respectively and by differentially combining the latter the low frequency (modulating) potential is obtained as follows:

$$v = (E_7 - E_8) \sqrt{2} = 4E \sin b_{1/2} \quad (IX)$$

By properly choosing the propagation characteristics of the phase rotating circuit it is possible in this manner to secure a linear relation between v and u and to obtain linear demodulation without distortion.

Referring to Figures 1 and 2 of the drawings, there are shown band-pass filters suited as phase rotating networks for the purpose of the invention and comprising a pair of tuned or resonant circuits mutually coupled with each other inductively, capacitatively or in any other suitable way. The network shown in Figure 1 comprises a first resonant circuit constituted by an induction coil 10 shunted by a condenser 11 which may be fixed or variable as in the example shown. To this resonant circuit is applied through input terminals $a-b$ a high frequency input potential e_5 which may be a phase or frequency modulated carrier potential. The resonant network further comprises a second resonant circuit constituted by an induction coil 12 shunted by a condenser 13 and having its lower side connected directly and having its upper side connected through a suitable coupling reactance such as an induction coil 14 in the example illustrated with the corresponding sides or terminals of the resonant circuit 10-11. The output potential e_6 is derived from the resonant circuit 12-13 through terminals $c-d$.

In the modification according to Figure 2, there is provided a common induction coil 14' in series with the induction coils 10 and 12 of the resonant circuits forming a common coupling link therebetween with or without additional inductive coupling between the coils 10 and 12. This circuit may be further modified by omitting the induction coil 14' and providing an inductive coupling only between the coils 10 and 12 thereby obtaining a band-pass filter of known design as shown in the embodiment according to Figure 3. The band-pass filters may also be constructed symmetrically as shown in Figures 7 and 8.

In resonant networks or band-pass filters of the above described type the phase of a high frequency output potential (e_6) relative to the phase of the impressed input potential (e_5) varies in dependence upon the frequency within the range from 0 to 180° . If both resonant circuits 10-11 and 12-13 are in tune with the impressed frequency the mutual phase angle between the input and output potentials (e_5 and e_6) is exactly 90° and varies in either direction from this value in dependence upon the sense and in proportion to the amount of detuning relative to the impressed frequency.

According to the present invention, the input and output potentials e_5 and e_6 are combined by means of a non-linear or modulating device adapted to produce a resultant current or poten-

tial which is a function of the product of the potentials being combined and which contains a steady or direct current component varying according to the relative phase angle between the impressed potentials that is in turn according to the instantaneous frequencies of the impressed high frequency carrier signal e_s . Since in case of a frequency modulated carrier wave the frequency variations correspond to the instantaneous values of a low frequency or signal wave there is obtained in this manner a demodulated or low frequency signal directly without necessitating an intermediate conversion into a corresponding amplitude modulated carrier as required in previously known demodulating systems for high frequency modulated carrier signals.

A complete demodulating arrangement according to the invention utilizing an electronic converter or mixer valve is shown in Figure 3, wherein the mixer valve 15 has a cathode 16, a first control grid 17, a second control grid 18, the latter being shielded from the control grid 17 by a positively biased screen 19 in a manner well known, and an anode or plate 20. The input potential e_s developed by a frequency modulated wave impressed upon the resonant circuit 10—11 is applied to the cathode and control grid 18 of the valve, while the output potential e_o of variable phase depending on the instantaneous values of the low frequency or demodulating signal and developed across the tuned circuit 12—13 is applied to the cathode and control grid 17 of the valve 15. There is thus produced by the action of the two grids 17 and 18 in the output or plate circuit of the valve a current which is a function of the product of the potentials e_s and e_o impressed upon the grids 18 and 17, respectively, which current contains a component v varying according to the relative phase difference between the impressed potentials or in turn according to the low frequency or demodulating signal variations. The latter may be applied to a further utilization or translation device connected to the terminals $e-f$ through a coupling condenser 21 and coupling impedance such as a high ohmic resistance 22 inserted between the plate 20 and the positive pole of a high tension source for supplying the plate current as indicated by the +B sign.

According to another modification, the sum and difference potentials $e_s + e_o$ and $e_s - e_o$, respectively, may be rectified separately by means of suitable rectifiers and the rectified potentials combined differentially in a balanced demodulator arrangement such as of the type shown in Figure 4. In this manner too an output current containing the low frequency or signal component is obtained.

Referring more particularly to Figure 4, there are provided a pair of rectifiers 16' and 17' serially connected with their conductive directions opposed to each other across the secondary tuned circuit 12—13 through a pair of blocking condensers 14' and 15'. The opposite sides of the primary circuit 10—11 are connected to the electrical center point of the secondary circuit 12—13 such as to the mid-point of the induction coil 12 as shown in the example and to the common terminal of the rectifiers 16' and 17'. In this manner, there are formed two rectifying circuit paths each including one of the rectifiers 16' and 17' and one of which has impressed thereon the sum ($e_s + e_o$) and the other of which has impressed thereon the difference ($e_s - e_o$) of the relatively dephased input and output potentials developed in the primary and secondary circuits 10—11 and 12—13, respectively. In this manner, a differ-

tial rectified potential v is obtained from the output terminals $e-f$ of rectifiers 16' and 17' preferably through smoothing choke coils or resistances 20' and 21'. This circuit is also suited for producing an amplitude responsive (AVC) potential r which latter in the example illustrated is derived from the common terminal (g) of the rectifiers 16' and 17' on the one hand and from the common terminal of a pair of additional high ohmic resistances 22' and 23' serially connected across the rectifiers. The condensers 14' and 15' serve for blocking the demodulated potential v from the resonant circuits of the filter network. In order to afford a constant discharge of these condensers there are provided a pair of leak resistances 18' and 19' connected between the right-hand terminals of the condensers 14' and 15' and the common terminal of the rectifiers 16' and 17' preferably through a common potential source 24 providing a biasing potential in the conductive direction of the rectifiers. Alternatively, the leak resistances 18' and 19' may be connected in parallel to the condensers 14' and 15', respectively, as will be readily understood.

In view of the relatively rapid variations of the frequency and accordingly of the output potential v it is necessary that the time constant of the condensers 14' and 15' in combination with their associate leak resistances should be small compared with the oscillating period of the highest low frequency or signal component to be expected.

The rectifiers 14' and 15' may be of any known type such as dry rectifiers or valve rectifiers having a sharply pronounced blocking potential. Alternatively, rectifiers with a gradually curved characteristic such as those with a square law characteristic may be provided as is customary in push-pull modulators. Moreover, grid leak or audion rectification may be employed as shown in Figure 6.

In specific cases it may be advisable to use an arrangement of the rectifiers as shown in Figure 5 wherein the position of condensers 14' and 15' and the rectifiers 16' and 17' are exchanged and the former are charged additionally with the relatively slow varying output potential supplied through the inductance coils and resistances 20' and 21'. In this case too it is necessary that the time constant of the leak resistances 18' and 19' and associate condensers 14' and 15', respectively, is kept sufficiently low to enable the output potential v to follow the low frequency fluctuations of a received high frequency carrier wave.

Referring to the embodiment of Figure 6, there is shown therein a band-pass filter of the type according to Figure 2. The induction coils 10' and 12' may or may not be inductively coupled to provide additional coupling between the primary and secondary resonant circuits in addition to the coupling afforded by coil 25 arranged symmetrically between the coils 10' and 12'. In place of the induction coil 25 a capacitive coupling reactance or condenser may be provided to form a common coupling impedance between the primary and secondary resonant circuits. The balanced demodulation of the primary and secondary potentials e_s and e_o is carried out by grid leak detection by means of a twin triode valve 26 having a pair of control grids 27, 28 cooperatively arranged with a common cathode 29. The potentials developed at the opposite sides of the output resonant circuit are applied to the grids 27 and 28 through coupling condensers 14' and 15', the grids 27 and 28

being further provided with grid leak resistances 18' and 19' in series with a common grid biasing source 24. The cathode 29 is further connected to one end (terminal b) of the input resonant circuit. The anodes of the two triode units of tube 26 are connected through a pair of resistances 32 and 33 the common terminal of which is connected to the cathode 29 through a high tension source 31 and a further resistance 30. The latter serves to supply an amplitude responsive (AVC) potential r derived from terminals g and h . In this manner the average grid biasing potential of both grids 27 and 28 is controlled according to the impressed sum or difference potential ($e_s + e_c$ and $e_s - e_c$), respectively, and an amplified demodulated or signal potential v is obtained at the output terminals $e-f$ connected to the plates of the valve through smoothing impedances 20' and 21'.

A further demodulating arrangement according to the invention is shown in Figure 7. The latter also comprises a band-pass filter constituted by a pair of resonant circuits 10-11 and 12-13 in inductive coupling relation with each other and connected to a rectifier or modulator bridge for producing a demodulated or frequency responsive output potential. The rectifier bridge comprises four rectifiers 35, 36, 37, 38 connected in series in like sense as regards their conductive directions to form a closed circuit. The opposite sides of the input resonant circuit 10-11 are connected to one pair of diagonal points of the rectifier bridge and the opposite ends of the secondary resonant circuit 12-13 of the band-pass filter are connected to the remaining pair of diagonal points of the bridge, while the demodulated or frequency responsive output potential v is derived from the electrical centers of the circuits 10-11 and 12-13 such as the center tap points j and i of the induction coils 10 and 12, respectively, as shown in the example illustrated. The rectifiers 35, 36, 37, 38 are preferably dry rectifiers whose impedance may be increased by providing high ohmic resistances in series therewith to reduce the damping imposed upon the associate filter circuits.

In an arrangement of the aforescribed type, if the received instantaneous frequency is low, the potentials e_s and e_c developed by the primary and secondary circuits are substantially in phase. In this case, due to the rectifying effect the center tap i of the secondary 12 assumes a positive charge relative to the center tap j of the primary 10. If, on the other hand, the received instantaneous frequency is higher than the average or carrier frequency f_0 , the potentials e_s and e_c will be in counter-phase and the point i will become negative relative to the point j . For the average or carrier frequency f_0 the mutual phase angle between the potentials is 90° whereby the output potential v between i and j becomes zero. In this manner there is supplied at the terminals i, j or e, f , respectively, a demodulated or output potential v varying according to the frequency deviations f_1 of the received high frequency carrier wave from a predetermined average or carrier frequency (f_0). The damping of the circuits should be chosen in such a manner that the phase rotation and in turn the output potential v is linearly related to the frequency variations to be dealt with.

The circuit according to Figure 7 may be derived from the principal circuit diagram according to Figure 2 by reducing the inductance 14' to a small value. If a band filter coupling is employed with large coupling impedances according to Figure 1, an arrangement is obtained as shown in Figure 8 wherein 39 and 40 represent sufficiently

high inductive or capacitive coupling impedances. The latter should be designed in such a manner to prevent a short circuit or circulatory current path for the output or demodulated potential v occurring between terminals $e-f$. The operation of the rectifier or demodulator bridge 35, 36, 37, 38 in Figure 8 is substantially similar to the embodiment of Figure 7.

In Figures 1 to 8, the input impedance of the band-pass filter is determined essentially by the impedance of the primary resonant circuit 10-11. A frequency responsive or output potential v may be obtained also in the case that the condenser 11 is omitted and replaced by a sufficiently high inductive or ohmic impedance. An arrangement of this type is shown in Figure 9 wherein a tuned primary circuit is dispensed with. There is provided in place of the tuned circuit a sufficiently high impedance such as an ohmic resistance 41 connected across the input terminals $a-b$ and 20 having a center tap leading to the output terminal e . This impedance serves merely for restoring the symmetry of the terminal e relative to the terminals $a-b$ which symmetry is required for producing the demodulated potential v in view of the described function and operation of the rectifier bridge 35, 36, 37, 38.

Referring to Figure 10 there is shown a different type of resonant network for effecting a phase rotation suited for the purpose of the invention. The circuit shown comprises a series combination connected across the input terminals $a-b$ including a parallel tuned circuit constituted by an induction coil 44 shunted by a condenser 33 and an induction coil 45 in series with the tuned circuit. In such an arrangement, a potential derived from the circulatory current i_s within the tuned circuit 43, 44 is compared with a potential derived from the total current i_a flowing into the tuned circuit. The currents i_s and i_a will be in phase in case of resonance of the circuit 43, 44 with the impressed input signal e_s (frequency f_0). In case of detuning of the circuit 43, 44 relative to the instantaneous impressed frequency, the phase relation between the currents will assume values differing from zero. If the impedance 45 is inductive or capacitive for all the frequencies to be dealt with the potential e_7 between the terminals l and n and the potential e_6 between terminals n and m will be phase shifted by 90° relative to each other in case of resonance while the phase rotation in case of detuning deviates in either sense and in proportion to the amount of detuning from this value. Thus, a frequency responsive or demodulated output potential may be obtained by differential demodulation of the two potentials in a manner similar to the embodiments described hereinbefore.

Referring to Figure 11, there is shown a complete demodulating circuit arrangement employing a phase rotating resonant network of the type according to Figure 10 and a rectifier or modulator bridge of the type described previously. In order to maintain the symmetry of the circuit, the impedance 45 in the example chosen an inductive impedance is split into two halves 45' arranged serially at opposite sides of the tuned circuits 43-44. Both inductances 45' are in symmetrical coupling relation with a further inductance coil 46 serving to supply the potential e_7 to be combined with the potential e_6 developed by the circuit 43-44 by means of the rectifier bridge 45, 46, 47, 48 thereby to produce a frequency responsive or demodulated potential v at the output terminals

$e-f$ connected to the center tap points of the inductances 44 and 46, respectively.

According to a further modification of the invention, a simple series tuned circuit is employed as a phase rotating circuit for carrying out the method according to the invention. An arrangement of this type is illustrated in Figure 12. In the latter, a series tuned circuit comprising an induction coil 47 and a condenser 48 is connected across the input terminals $a-b$ whereby the total potential e_s impressed upon the circuit may be compared with the potential e_s developed by the inductance 47 by impressing both potentials upon the grids 18 and 17, respectively, of an electronic modulator or mixer valve in substantially a similar manner as described in connection with Figure 3. In case of resonance (carrier frequency f_0) the resonant circuit 47, 48 offers a pure ohmic impedance to the impressed potential whereby the potential drop e_s across the inductance 47 is in phase quadrature to the total input potential e_s . In case of deviation of the impressed frequency from the resonance frequency of the circuit the relative phase between the potentials e_s and e_s varies in either direction in dependence upon the sense and in proportion to the amount of detuning whereby a frequency responsive or demodulated output potential v is obtained at terminals $e-f$ in a manner well understood from the above. As is obvious, the same result is obtained by comparing and combining the potential drop e_r developed across the condenser 48 with the total input potential e_s .

Referring to Figure 13, there is shown a circuit arrangement embodying a series tuned circuit and a rectifier bridge for effecting balanced modulation in place of a direct conversion obtained by a mixer valve provided in the arrangement according to Figure 12. In the example illustrated, the series tuned circuit is constituted by an induction coil 50 in series with condensers 49 and 51 of equal capacity and arranged symmetrically to and at opposite sides of the coil 50 to maintain the symmetry of the system. There is further provided an impedance in the example shown a high ohmic resistance connected across the input terminals $a-b$ and having its center tap connected to one of the output terminals (e) the other output terminal (f) being connected to the center tap of the inductance 50 of the series tuned circuit. The potentials e_s and e_s are impressed upon different pairs of diagonal points of the rectifier bridge 35, 36, 37, 38 in substantially the same manner as described hereinbefore to produce a demodulated or frequency responsive output potential v .

The arrangements hereinbefore described constitute examples of the invention for deriving currents or potentials between definite points of a tuned electrical network and for combining the same with the aid of dry or valve rectifiers or other modulating devices to produce a resultant current or potential varying according to the deviations of the frequency of a carrier wave impressed upon said network relative to a predetermined frequency such as the carrier frequency in the case of a frequency modulated signal. As is understood, numerous variations and modifications differing from the exemplifications illustrated may be resorted to coming within the broader scope and spirit of the invention.

The frequency demodulating or conversion systems as described may be energized by the high frequency signals received directly or the received signals may be converted to signals of

a fixed intermediate frequency by combination with local signals of auxiliary frequency in which case the conversion or demodulation of the signals of varying frequency is carried out at intermediate frequency whereby the demodulating circuit may be fixedly designed and adjusted for this frequency.

In carrying out the process according to the invention it has been found desirable that the amplitude of the received frequency modulated signals should have a substantially constant value. In practice rapid amplitude variations may occur for one reason or another liable to interfere with the demodulation or conversion of the signals. In most cases a certain amplitude modulation in addition to the frequency modulation of the transmitted carrier wave cannot be avoided. In other cases the amplitude of the amplified high frequency carrier varies in accordance with fluctuations of the operating potentials if the supply potentials are insufficiently filtered or smoothened. Moreover, an apparent amplitude modulation of the received carrier wave may be caused by interference with oscillations received from disturbing transmitters or other interfering sources. In these cases, it is advisable to regulate the amplitude of the potentials applied to the frequency demodulation or conversion circuit by a special limiting arrangement so as to maintain the amplitudes at a substantially constant value. This can be effected by the aid of transmission elements having a pronounced saturation characteristic such as dry rectifiers suitably biased or electron valves having a bent characteristic operating curve. In the latter case the high frequency oscillations of sinusoidal shape are converted into substantially rectangular curves of constant height from which the fundamental wave may be segregated by the aid of tuned circuits and low-pass filters if required. It is furthermore possible to employ a limiting impedance varying according to the amplitude of the current passed therethrough in the form of a hot wire resistance to obtain the desired effect.

The regulation to a constant amplitude may also be effected by means of a control system of the type of the anti-fading or automatic volume control arrangements by regulating the amplification of a control valve in accordance with an amplitude responsive control potential. The time constants of this arrangement should be sufficiently low to ensure equalization of the very rapid amplitude fluctuations corresponding to the low frequency modulation. The arrangements for receiving frequency modulated carrier waves may furthermore be combined with a system for automatic frequency control (AFC) to the average or carrier frequency of the received oscillations. For this purpose the resonant circuits are controlled by a tuning responsive potential which may be generated in a simple manner by proper smoothing or filtering of the output potential v .

A principal arrangement of the aforementioned type is shown in block diagram form in Figure 14. In the latter, the frequency modulated high frequency potential e_1 is amplified in an amplifier 53 to a value e_s sufficient for impression upon the frequency converter or demodulator 54 which may be of any of the previously described types and which serves to produce a frequency responsive or output potential v in the manner described. Since this potential v is proportional to the deviations of the received frequency from a normal

(carrier) frequency, it corresponds to the low frequency or modulating signal except for an additional direct current component. This direct current component is caused by small deviations of the tuning of the receiver relative to the average (carrier) frequency of the received signals and may be suppressed by means of a high-pass filter 55 to obtain a true demodulated or low frequency output signal v_1 from the output of the filter 55. At the same time, this direct current component is applied through a low-pass smoothing filter 56 which latter is impermeable to all speech and modulating signal frequencies, to the variable tuning elements of the selective or resonant circuits of the amplifier 52 and the demodulator 54 (potential v_2) in such a manner that the amount of detuning of the latter relative to the average (carrier) frequency f_0 is automatically maintained at a predetermined small value.

In addition, the degree of amplification or gain of the amplifier 53 may be controlled in accordance with an amplitude responsive control potential r obtained for instance in a manner described in connection with Figures 4 to 6 to substantially suppress any disturbing and undesired amplitude modulation of the received signals. An additional automatic volume control to compensate slow variations of the carrier amplitude due to fading and similar causes may be employed by impressing a normal AVC potential r_1 obtained from the potential r by suitable filtering by means of a filter 57. Thus, the amplitude control potential r is obtained by ordinary rectification of the high frequency potential and under circumstances may be derived from the demodulator 54 as shown in Figures 4 to 6.

Referring to Figure 15, there is shown a similar receiving arrangement to Figure 14 as applied to a superheterodyne receiver. The frequency modulated input potential e_1 is applied to a radio frequency preamplifier 58 and the amplified potential e_2 impressed upon a frequency converter or mixer 59 to be combined therein with a locally generated auxiliary potential e_3 of different frequency (oscillator 60) resulting in a beat or intermediate frequency potential e_4 in a manner well known. The latter is amplified by the aid of an intermediate frequency amplifier 61 and the amplified intermediate frequency potential e_5 applied to a demodulator 54 in substantially the same manner as in the case of Figure 14. The demodulated or output potential v is again impressed upon a high-pass filter 55 to obtain a final output potential v_1 free from any direct current component while the unfiltered potential v is passed through a low-pass filter 56 and the filtered output v_2 thereof applied to a frequency controlling element of the local oscillator 60 in such a manner that the intermediate average frequency never deviates to an undesirable degree from the predetermined value. The amplitude control potential in the example shown is derived from the demodulator 54 and applied to a gain control element of the intermediate frequency amplifier 61 to suppress undesirable amplitude modulation of the potential e_5 . The slow variations of potential r derived by means of a filter 62 are applied to a gain control element of the preamplifier 53 to effect a normal automatic volume control (AVC) in a manner similar to the arrangement in Figure 14.

The arrangements according to the invention are also suited for the reception of a phase modu-

lated carrier wave as it is well known that a phase modulation is equivalent to a frequency modulation with a linearly distorted low frequency or modulating wave. In this case, it is only necessary to provide a special filter in the low frequency section having a propagation factor inversely proportional to the frequency of the modulating or low frequency signal components to compensate for the linear amplitude distortion in dependence upon the frequency.

It will be obvious from the above that the invention is not limited to the specific exemplifications and methods disclosed herein for illustration and that many possible embodiments and changes may be made in the embodiments above set forth. Accordingly, it is to be understood that all matter hereinbefore set forth or shown in the accompanying drawings is to be interpreted as illustrative and not in a limiting sense.

I claim:

1. In a frequency discriminator, a network comprising a capacitive and an inductive reactance element in series, said network being resonant to a predetermined frequency, means for impressing a high frequency signal potential upon said network, the relative frequency of said signal potential with respect to the resonant frequency of said network being varied in either direction from said predetermined frequency, an electron discharge tube having a cathode and an anode for producing an electron space current, a pair of control grids located in the path of said space current, a further grid located between said control grids, means for maintaining said further grid at a positive potential with respect to said cathode, an output circuit connected to said anode, means for impressing the total signal potential developed by said network upon one of said control grids, further means for impressing signal potential developed by one of said reactance elements upon said other control grid, and impedance means in said output circuit adapted to develop output potential having an amplitude varying in sign and magnitude proportionately to the frequency departure of said signal potential from the frequency to which said network is resonant.

2. In a frequency discriminator, a network comprising a capacitive and an inductive reactance element in series, said network being resonant to a predetermined frequency, means for impressing a high frequency signal potential upon said network, the relative frequency of said signal potential with respect to the resonant frequency of said network being varied in either direction from said predetermined frequency, an electron discharge tube having a cathode and an anode for producing an electron space current, a pair of control grids located in the path of said space current, a further grid located between said control grids, means for maintaining said further grid at a positive potential with respect to said cathode, an output circuit connected to said anode, means for impressing the total signal potential developed by said network upon one of said control grids, further means for impressing signal potential developed by said inductive reactance element upon said other control grid, and impedance means in said output circuit adapted to develop output potential having an amplitude varying in sign and magnitude proportionately to the frequency departure of said signal potential from the frequency to which said network is resonant.

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