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(54) An arrangement for amplifying a modulated carrier signal

(57) A modulated carrier signal e.g. O.Q.P.S.K, SSB, the amplitude variations of which are smaller than the amplitude of the unmodulated carrier, is not very suitable for transmission over radio links because the required use of non-linear components, such as class-C amplifiers, result in an unwanted widening of the transmitted spectrum. According to the invention an arrangement for amplifying such a signal comprises two oscillators (16, 17) the desired relative phase of which is adjusted in a loop including a control circuit 22. On assembling the oscillator signals v'_k and v'_g, there is produced a resulting signal r which is a replica of the input signal v having the amplitude variations. As the oscillators 16 and 17 produce signals having a constant amplitude, these signals can be easily amplified, for example by class-C amplifiers. The control circuit 22 may include phase or amplitude comparators for comparing the input signal or signals derived therefrom with signals derived from the oscillators.

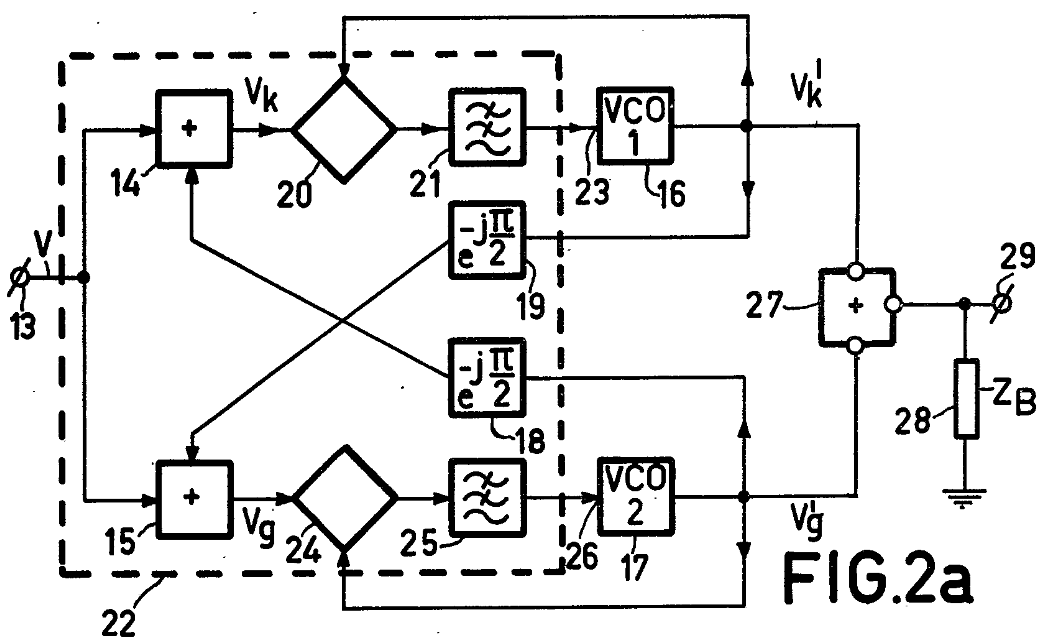

FIG.2a

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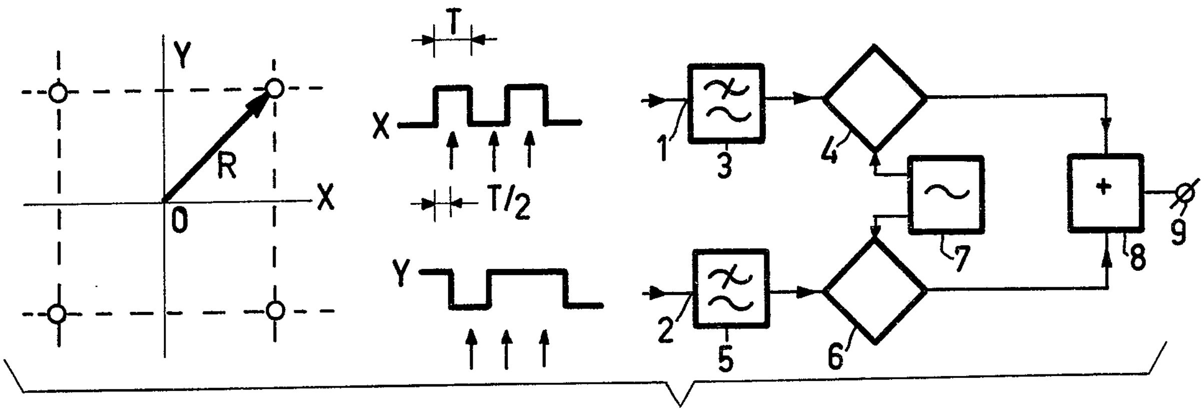

FIG.1a

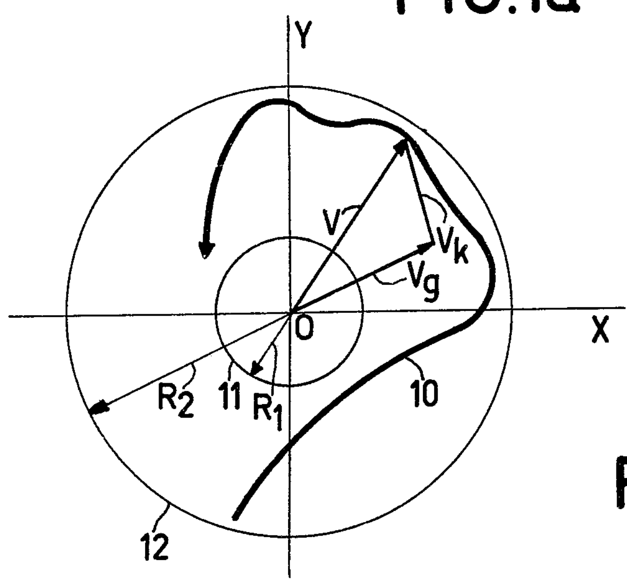

FIG.1b

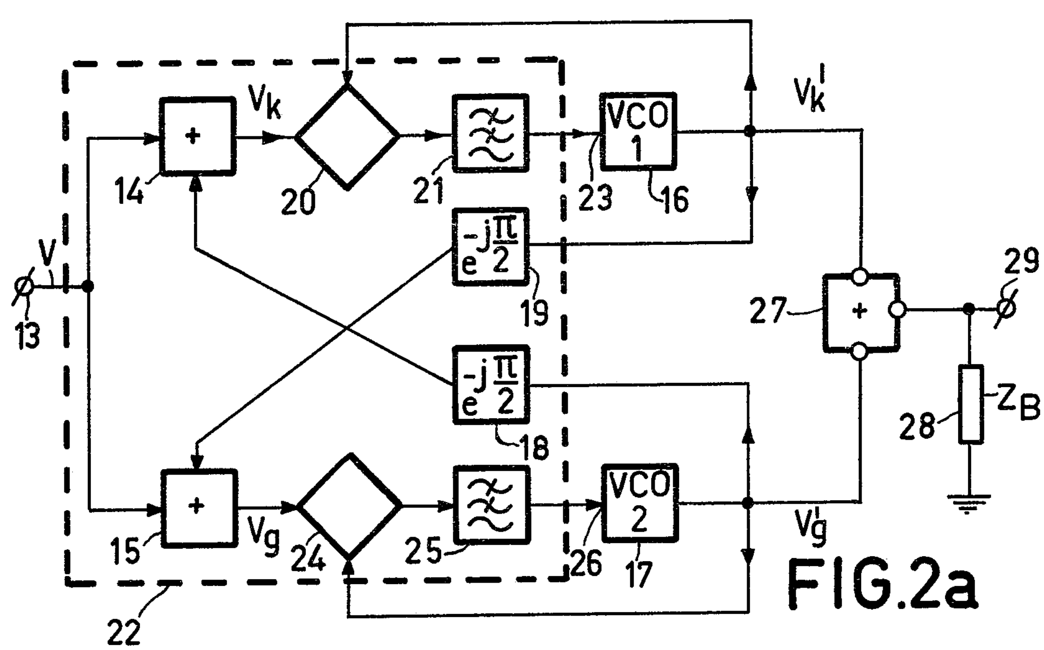

FIG.2a

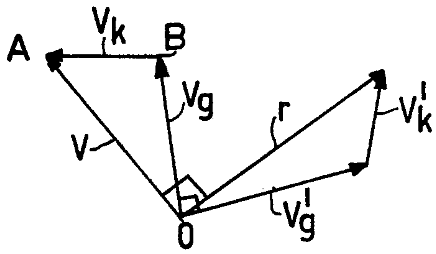

FIG.2b

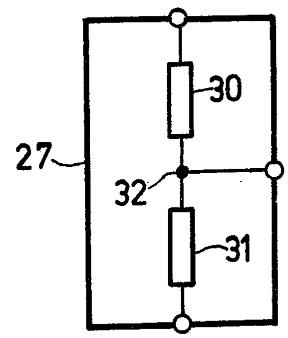

FIG.2c

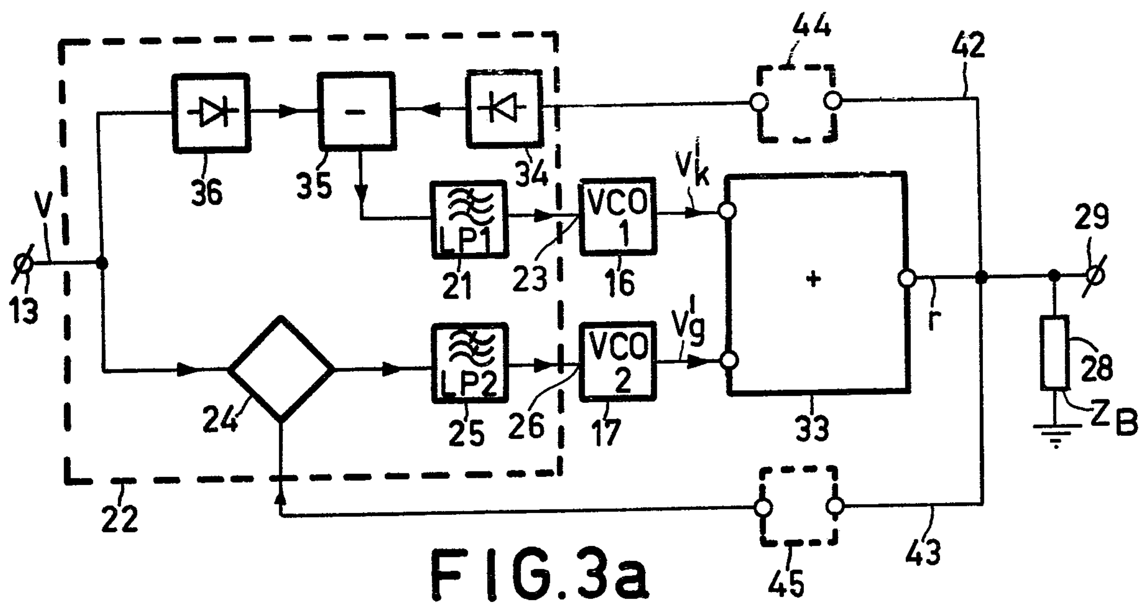

FIG.3a

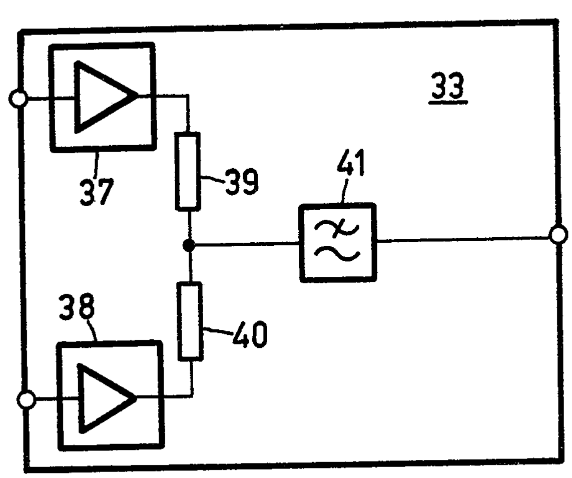

FIG.3b

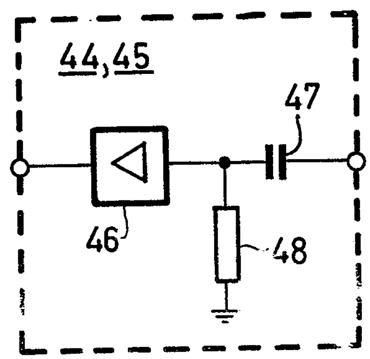
44,45
47
46
48

FIG.3c

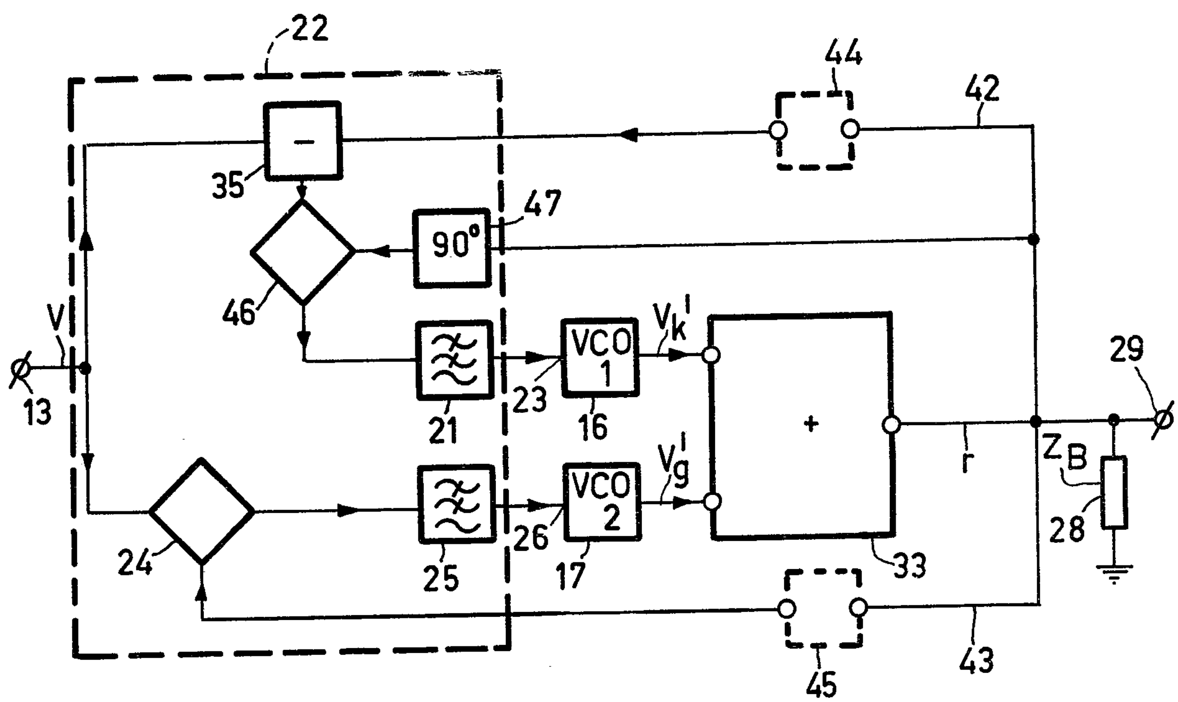
22
44
42
35
47
90°
46
V
VCO 1
V_k'
29
13
21
23
16
+
r
Z_B
VCO 2
V_g'
24
25
26
17
33
28
45
43

FIG.4

SPECIFICATION

Arrangement for amplifying a modulated carrier signal

The invention relates to an arrangement of a type suitable for amplifying a modulated carrier signal the amplitude variations of which are smaller than the amplitude of the unmodulated carrier.

In recent years numerous modulation methods aimed at achieving efficient data transmission over telephone lines have been designed and introduced. In almost all cases these modulation methods have resulted in a modulated carrier signal having amplitude variations (i.e. amplitude modulation), and they use linear modulators and amplifiers.

However, these modulation methods are not so suitable for data transmission over radio links because, in radio communication systems, the need for a high power economy requires the use of components having a non-linear amplitude transfer function and the spectrum at the output of such a component, for example a class C amplifier, will be wider than the spectrum at the input when the amplitude at the input varies. Therefore, in radio communication systems preference is given to modulation methods which result in a modulated carrier signal of a substantially constant amplitude, which means the use of angle modulation. See, for example, reference (1).

The invention has for its object to provide an arrangement of the type referred to, by which a modulated carrier signal having amplitude variations is amplified, the arrangement allowing the use of components having a non-linear amplitude transfer function–such as class-C amplifiers–while the spectrum at the output of the arrangement is not wider to any significant extent than at the input.

According to the invention, an arrangement of the type referred to is characterized by comprising a first and a second controlled oscillator, each having a control input and an output, which oscillators are connected by means of their respective control inputs to a control circuit and are operable to oscillate at a substantially constant amplitude and at a frequency which substantially corresponds to the carrier frequency, the output of each oscillator being coupled to the control circuit for generating control signals for the oscillators from comparisons between the modulated carrier signal and the oscillator signals, the arrangement also having an output circuit connected to the outputs of the oscillators for vectorially assembling an output signal.

The arrangement of a type suitable for amplifying a modulated carrier signal in accordance with the invention may be further characterized in that, the control circuit comprises a first and a second phase comparator circuit, each having a first and a second input and an output, together with a first and a second low-pass filter, that the output of the first oscillator is connected to the second input of the first phase comparator circuit and the output of the first phase comparator circuit is connected to the control input of the first oscillator through the first low-pass filter, that the output of the second oscillator is connected to the second input of the second phase comparator circuit and the output of this second phase comparator circuit is connected to the control input of the second oscillator through the second low-pass filter, that the control circuit further comprises a first and a second adder circuit, each having a first and a second input and an output, together with a first and a second delay element, that the first input of the two adder circuits are interconnected and connected to an input of the arrangement, that the second input of the first adder circuit is connected to the output of the second oscillator through the first delay element and the output of the first adder circuit is connected to the first input of the first phase comparator circuit and that the second input of the second adder circuit is connected to the output of the first oscillator through the second delay element and the output of the second adder circuit is connected to the first input of the second phase comparator circuit.

It should be noted here that such a control circuit is indeed disclosed in United States Patent Specification 3,873,931, but said Patent Specification relates to a FM demodulator for separating the useful, original, transmitted signal from the noise signal.

Embodiments of the invention and their advantages will now be further explained by way of example with reference to the accompanying drawings.

In the drawings:

Figure 1a is a block schematic circuit diagram of a known modulation stage;

Figure 1b shows the variation of a signal vector in the phase plane;

Figure 2a is a block schematic circuit diagram of a first embodiment of an arrangement according to the invention;

Figure 2b is a phase diagram for illustrating the operation of Figure 2a;

Figure 2c shows an output circuit for use in the first embodiment shown in Figure 2a;

Figure 3a is a block schematic circuit diagram of a second embodiment of an arrangement according to the invention;

Figure 3b shows an output circuit for use in the second embodiment;

Figure 3c shows a feed-back network for use in the second embodiment; and

Figure 4 is a block schematic circuit diagram of a third embodiment of an arrangement according to the invention.

Corresponding elements in the Figures have been given the same reference symbols.

1. F. de Jager, C. B. Dekker, "Tamed Frequency Modulation, a novel method to achieve spectrum economy in digital transmission" IEEE Trans. Comm. Vol. CDM-26, No. 5, May 1978, pages 534-542.

2. S. A. Rhodes, "Effect of noisy phase reference on coherent detection of offset QPSK signals", IEEE Trans. Comm. Vol. CDM-22, No. 8, Aug. 1974, pages 1046-1055.

3. S. A. Gronemeyer, A. L. McBride, "MSK and Offset QPSK modulation" IEEE Trans. on Comm. Vol. CDM-24, No. 8, Aug. 1976, pages 809-820.

Figure 1a shows a block schematic circuit diagram of a known modulation stage for a modulation method which is referred to in the literature as "Offset Quadrature Phase Shift Keying", abbreviated to OQPSK (see reference (2)). This modulation method differs from quadrature-phase modulation in that the data signal of the Y-channel is shifted by a time T/2 with respect to the X-channel data signal. As a result, the X- and Y-channels must be sampled in a receiver at different instants, for example at the instants as indicated by means of the arrows in the data signals X and Y of Figure 1a. The modulation stage has an input 1 for data signal X and an input 2 for data signal Y. After filtering by a low-pass filter 3, the signal X is applied to an input of a modulator 4. After filtering by a low-pass filter 5, the signal Y is applied to an input of a modulator 6. Signals from a carrier oscillator 7 are applied to a further input of modulators 4 and 6, respectively, these signals having a relative phase of 0° and 90°, respectively. The output signals of the modulators 4 and 6 are added together in an adder 8 and applied to an output 9. The signal vector of the modulated carrier signal available at the output 9 has the property that the amplitude variations are smaller than the unmodulated carrier amplitude. To illustrate this, Figure 1b shows in the phase plane a portion of a path 10 of the signal vector v. It is seen that the amplitude of signal vector v is invariably situated in the area between the circles 11 and 12, which have a radius R1 and R2, respectively, wherein R2 > R1 > 0.

In practice, it appears that for OQPSK the radius R1 is approximately equal to 0.5 R and the radius R2 is approximately equal to 1.5 R, R being the distance from the origin to the characteristic phase points in 4-phase modulation. When the input signals X and Y have the value +1 or −1 at the sampling instants, R is equal to $\sqrt{2}$.

In the arrangement according to the invention use is made of the property that the final point of the signal vector v remains at a certain distance from the origin (in OQPSK even at a considerable distance, as described above).

It should be noted that, in addition to OQPSK, the invention is also applicable to other modulation methods used for digital signals, provided the indicated property is present. It further appears that the principle described hereinafter may alternatively be used for analogue signals, single-side band signals in particular. By way of example, the invention will be further explained with reference to the example of OQPSK given in Figure 1, although the invention is not restricted thereto.

As shown in Figure 1b the arbitrary final point of the signal vector v may be represented by giving two vectors v_g and v_k–each having a constant amplitude and consequently a fixed amplitude ratio with respect to each other–the desired phase angles. If, for example, a value 0.6 v_g is chosen for v_k, then the amplitude of the resulting vector v may be varied by a factor of 4 between minimum and maximum amplitudes, which is sufficient to handle 3 to 1 amplitude ratio 1.5 R:0.5 R as required for OQPSK. E(2) Description of the preferred embodiments.

Figure 2a shows a block schematic circuit diagram of a first embodiment of an arrangement according to the invention. The arrangement comprises a first voltage-controlled oscillator 16 and a second voltage-controlled oscillator 17, each of which forms part of a loop and is connected to a control circuit 22. The oscillators 16 and 17 oscillate a frequency which substantially corresponds to the carrier frequency of the input signal. An input terminal 13 of control circuit 22, to which input an OQPSK input signal is applied, is connected to a first input of both a first adder circuit 14 and a second adder circuit 15. The oscillator signal of the second oscillator 17, shifted 90° by a first delay element 18 is applied to a second input of the first adder circuit 14. The sum of the signals applied to the two inputs of the first adder circuit 14 is applied to a first input of a first phase comparator circuit 20 to compare it with the oscillator signal of oscillator 16, which is applied to a second input of the first phase comparator circuit 20. The output signal of the first phase comparator circuit 20, which is a measure of the phase difference between the input signals applied thereto, is filtered by a first low-pass filter 21 and applied to a control input 23 of the first oscillator 16. Likewise, the oscillator signal of the first oscillator 16, shifted 90° by a second delay element 19 is applied to a second input of the second adder circuit 15. The sum of the signals applied to the two inputs of the second adder circuit 15 is applied to a first input of a second phase comparator circuit 24 to compare it with the oscillator signal of the second oscillator 17, which is applied to a second input of the second phase comparator circuit 24. The output signal of the second phase comparator circuit 24, which is also a measure of the phase difference between the two input signals applied thereto, is filtered by a second low-pass filter 25 and applied to a control input 26 of the second oscillator 17. The oscillator signals of the two oscillators 16 and 17 are further applied to a combining circuit 27, which is loaded by an impedance 28. Impedance 28 may be, for example, a transmitting aerial or the input impedance of an output circuit–not shown–which is to be connected to output terminal 29: further power amplification and conversion, if necessary, to a desired radio frequency band, is carried out in this output circuit.

The arrangement shown in Figure 2a will now be further explained, reference also being made to the vector diagram shown in Figure 2b. Let the oscillator signal of oscillator 16 be v'_k and the oscillator signal of oscillator 17 be v'_g. In addition, let the desired signal vector v be applied at a low power, in correspondence with OQPSK modulation, to input terminal 13. The signal at the output of adder circuit 14 is then equal to

$$v + v'_g e^{\frac{-j\pi}{2}}.$$

In response to the phase control in control loop 17, 24, 25, the signal v'_g was subjected to a 90° phase shift with respect to the signal v_g, so $v'_g = V_g \exp(-j\pi/2)$. It therefore holds that:

$$v + v'_g e^{\frac{-j\pi}{2}} = v - v_g = v_k$$

In a similar manner it holds that:

$$v + v'_k e^{\frac{-j\pi}{2}} = v - v_k = v_g$$

So, when the voltages v_k and v_g are generated from v, v'_k and v'_g then it is effected by means of the two phase loops that the output voltages v'_g and v'_k, respectively, remain at a 90° angle with respect to the input voltages v_g and v_k, respectively.

To generate a high power, the input signal v, is supplied, with a low power, the high power oscillators or, for example, an oscillator followed by a class-C amplifier being controlled with this low-power input signal. An advantage is that co-operation of the two oscillators creates the possibility to introduce not only phase variations but also amplitude variations in the output signals, while the amplitudes of the signals to be amplified are constant, so that the non-linear amplitude transfer function of the components, for example class-C amplifiers, is no longer important.

The following points should be considered when implementing a circuit for the arrangement shown in Figure 2a. For a proper operation, it is necessary that the phase loop can react quickly to variations in the input signal. The band width of the low-pass filters 21 and 25 must therefore be relatively large. In addition, the loop gain must preferably also be large to keep deviations from the desired phase angles of 90° at a minimum. When this circuit is used to generate a high power, the delay elements 18 and 19 must also comprise attenuators. It was further found that a 5:3 ratio between the amplitudes of v_g and v_k is an advantageous ratio. In that case an average of approximately 75% of the total power is produced by oscillator 17 and approximately 25% by oscillator 16.

The output voltages v'_k and v'_g must be added together in a fixed ratio by output circuit 27. This is, for example, realised by connecting–as shown in Figure 2c–the output of oscillator 16 to connecting point 32 *via* a coupling impedance 30 and the output of oscillator 17 also to connecting point 32 *via* a coupling impedance 31. In order to ensure that the components v'_g and v'_k contribute in the desired ratio towards the output signal, the coupling impedances 30 and 31 must be accurately equal to each other.

A disadvantage of the arrangement shown in Figure 2a is that the precision which is required with respect to the equality of the coupling impedances 30 and 31 in the high-frequency range is difficult to realize. This results in noise components being introduced into the spectrum of the output signal. A further disadvantage is that, even when the coupling impedances 30 and 31 are perfectly equal to each other, difficulties may still occur as a result of the so-called interaction effect of the two oscillators 16 and 17. This is caused by the fact that, in general, the output voltage produced by an oscillator depends on the impedance by which it is located. Therefore, in Fig. 2a the impedance by which one oscillator is loaded depends on the voltage produced by the other oscillator. A phase control of oscillator 17, say, will therefore produce a voltage variation in the oscillator 16. This effect is indeed reduced by the negative feed-back provided, but the magnitude of this interaction effect must be taken into consideration when rating the loop again to be introduced therein, in order to be able to obtain the required precision.

The second embodiment of an arrangement for amplifying a modulated carrier signal is shown in Figure 3a. In the embodiment shown in Figure 2a, the information required for the two phase controls was derived directly from the output voltages of the individual oscillators, but in the second embodiment this information is derived from a point where the two components have already been combined, i.e. after assembly in an output circuit 33 at output 29. The relevant information must then be selected from the composite signal for each of the two oscillators. To this end the phase of the oscillator signal v'_k, produced by oscillator 16, is controlled in such manner that the amplitude of output signal r is correct and the phase of the oscillator signal v'_g, produced by oscillator 17, is controlled in such manner that the phase of output signal r with respect to v is correct (90°). The first-mentioned object is accomplished by applying output signal r after detection by a first amplitude detector 34 (for example a rectifier circuit) to an amplitude comparator circuit 35, to compare this output signal with the input signal v, which was processed by an amplitude detector 36. The oscillator 16 is adjusted to the correct value by means of the difference signal from amplitude-comparator circuit 35, which is applied to it *via* a low-pass filter 21. Simultaneously, the output signal r is compared with the input signal v in phase comparator circuit 24 and when there is deviation from the desired phase relationship of 90°, oscillator 17 is readjusted by applying the output signal of phase comparator circuit 24, after filtering in a low-pass filter 25, to the control input of oscillator 17. In this manner the resultant output signal r is made equal to the signal vector v, required for OQPSK-modulation, with the aid of a phase and an amplitude control from the final stage.

Figure 3b shows an embodiment of the output circuit 33 of the arrangement as shown in Figure 3a. The output signal v'_k, which is of a constant amplitude, of oscillator 16 is amplified by a class-C amplifier 37 and added to the output signal v'_g, which was amplified by a class-C amplifier 38, of oscillator 17 by means of coupling impedances 39 and 40.

As mentioned in the foregoing, phase variations which are passed on to coupling impedance 40 via class-C amplifier 38 are introduced in oscillator 17 in the rhythm of the information frequencies. In response thereto, the output impedance of class-C amplifier 37 varies and as a result thereof the composition of the higher harmonics changes. This means that the shape of the output signal is changed in such manner that the amplitude and the phase of the resulting signal r cannot be taken into account in the desired manner without further measures at the feedback. This produces extra side bands in the vicinity of the central frequency, that is to say an unwanted widening of the spectrum to be transmitted. In order to prevent this a low-pass filter having a cut-off frequency of, for example 1.5 × the carrier frequency, may be included in the feedback leads 42,

43. As will be apparent from Figure 3b only one low-pass filter is used, namely by arranging it between the common junction point of the coupling impedances 39 and 40 and the load impedance 28. This has the additional advantage that the higher harmonics in the output signal are also suppressed.

Separation networks 44 and 45 may be included in the return lines of the arrangement shown in Figure 3a, the main object being the prevention of non-linear impedances in parallel with the output impedance 28. As shown in Figure 3c, each separation network 44 and 45 comprises an isolation amplifier 46 and a coupling network consisting of a capacitor 47 in series with the return line and a resistor 48 between the input of amplifier 46 and ground.

The following should be noted as regards the implementation of the arrangement shown in Figure 3.

The arrangement shown in Figure 3 is based on a simultaneous modulation of amplitude and phase. Requirements must be imposed on the synchronization of phase and amplitude modulation. Let the desired amplitude be a(t) and the desired phase $\theta(t)$, then in agreement with the input signal v(t):

$$v(t) = a(t)e^{j\theta(t)}$$

If there is a time difference τ between a(t) and $\theta(t)$ then the output signal is given by

$$v(t) = a(t-\tau)e^{j\theta(t)}$$

which will produce unwanted side bands in the transmitted spectrum. From a worst-case analysis it appeared that in order to keep the amplitude of the first component, which is located outside the desired spectrum, approximately 80 dB below the nominal amplitude, $f_b\tau$ must be less than 10^{-3}. For $f_b = 16$ kb/s this means that $\tau < 60$ ns, wherein f_b is the bit frequency of the digital information signal. In practice, the time delays τ_0 and τ'_0, respectively, of low-pass filters 21 and 25, respectively, which each have a cut-off frequency of approximately $5 \times f_b$, are of importance.

However, because of the feedback

$$\frac{1}{A_1+1} \quad \text{and} \quad \frac{1}{A_2+1},$$

respectively, the time delays τ_1 and τ'_1, respectively, are smaller when A_1 is the loop gain in the loop formed by the elements 21, 16, 42, 34, 35, 21 and A_2 is the loop gain in the loop formed by the elements 25, 17, 43, 24, 25.

When $\tau_1 = \tau_2$, then there is only a constant delay between the signals r and v. When τ_1 and τ_2 differ then, in correspondence with the foregoing, in order to obtain 80 dB suppression, the following condition must be satisfied:

$$\left| \frac{\tau 0}{A_1+1} - \frac{\tau 0'}{A_2+1} \right| . f_b < 10^{-3}$$

When the cut-off frequency of the low-pass filters 21 and 25 is chosen equal to five times the bit rate then the condition

$$\tau_0 = \tau'_0 = (10\,\pi f_b)^{-1}$$

changes into

$$\left| \frac{1}{A_1+1} - \frac{1}{A_2+1} \right| < 0.03,$$

which can be easily satisfied.

As was described for the arrangement shown in Figure 2, the 5:3 ratio between the amplitude of v'_g and v'_k is an advantageous ratio. This also holds for Figure 3. However, if so desired, the stability of the loop of which oscillator 16 forms part can be increased by choosing a greater amplitude for v'_k. As will be apparent from the vector diagram shown in Figure 1b, this reduces the required variation of the angle between v_g and v_k. The stability of the loop of which oscillator 17 forms part is only endangered when the amplitude of the input signal becomes too small. As is also apparent from Figure 1b, this is avoided as the signal vector is always in an area for which it holds : $|v| > R_1$. So, choosing a higher value of v_k has an advantageous influence on the first-mentioned loop, while the stability of the last-mentioned loop is not endangered.

If so required v'_k, may be chosen equal to v'_g, although this will be not so desirable because of the fact that then two high-power oscillators are required.

Figure 4 shows a variant of the embodiment shown in Figure 3; more particularly, in accordance with this Figure 4, the amplitude difference between the signal vectors v and r is not measured by means of two amplitude detectors (34, 36) but by means of a modulator 46. The signal vectors v and r have almost the same phase and consequently the difference between the two (high-frequency) components, which difference is available at the output of amplitude comparator circuit 35, may be applied to the modulator 46 to which a carrier is applied which has approximately the same phase as v or r. The action of the fed-back VCO causes the carrier phase of r to differ by approximately 90° from the carrier phase of v. When a 90° shifting network 47 is used, the desired phase of the carrier for modulator 46 is obtained. The differential voltage detected by the modulator 46 is applied to low-pass filter 21 and processed in the manner described in the foregoing.

It should be noted that combining the output signals of the generator 16 and 17 cannot only be carried out in the manner as shown, for example, in Figure 2c, but that it can alternatively be done by means of a hybrid circuit. This results, however, in a loss of approximately 3 dB.

CLAIMS

1. An arrangement for amplifying a modulated carrier signal the amplitude variations of which are smaller than the amplitude of the unmodulated carrier, characterized by comprising a first and a second controlled oscillator, each having a control input and an output, which oscillators are connected by means of their respective control inputs to a control circuit and are operable to oscillate at a substantially con-

stant amplitude and at a frequency which substantially corresponds to the carrier frequency, the output of each oscillator being coupled to the control circuit for generating control signals for the oscillators from comparisons between the modulated carrier signal and the oscillator signals, the arrangement also having an output circuit connected to the outputs of the oscillators for vectorially assembling an output signal.

2. An arrangement as claimed in Claim 1, characterized in that, the control circuit comprises a first and a second phase comparator circuit, each having a first and a second input and an output, together with a first and a second low-pass filter, that the output of the first oscillator is connected to the second input of the first phase comparator circuit and the output of the first phase comparator circuit is connected to the control input of the first oscillator through the first low-pass filter, that the output of the second oscillator is connected to the second input of the second phase comparator circuit and the output of this second phase comparator circuit is connected to the control input of the second oscillator through the second low-pass filter, that the control circuit further comprises a first and a second adder circuit, each having a first and a second input and an output, together with a first and a second delay element, that the first input of the two adder circuits are interconnected and connected to an input of the arrangement, that the second input of the first adder circuit is connected to the output of the second oscillator through the first delay element and the output of the first adder circuit is connected to the first input of the first phase comparator circuit and that the second input of the second adder circuit is connected to the output of the first oscillator through the second delay element and the output of the second adder circuit is connected to the first input of the second phase comparator circuit.

3. An arrangement as claimed in Claim 1, characterized in that the control circuit comprises a phase and an amplitude comparator circuit, each having a first and a second input and an output, a first and a second low-pass filter, a first and a second amplitude detector, and a first and a second return line, that the output circuit is connected to the second input of the phase comparator circuit through the second return line and the output of the phase comparator circuit is connected to the control input of the second oscillator through the second low-pass filter that the output circuit is connected to the second input of the amplitude comparator circuit through the first return line and *via* the first amplitude detector, that the output of the amplitude comparator circuit is connected to the control input of the first oscillator through the first low-pass filter, that the input of the second amplitude detector is connected to the first input of the phase comparator circuit and to an input of the arrangement and that the output of the second amplitude detector is connected to the first input of the amplitude comparator circuit.

4. An arrangement as claimed in Claim 1, characterized in that the control circuit comprises a phase and an amplitude comparator circuit, each having a first and a second input and an output, a first and a second low-pass filter and a first and a second return line, that the output circuit is connected to the second input of the phase comparator circuit through the second return line and the output of the phase comparator circuit is connected to the control input of the second oscillator through the second low-pass filter, that the output circuit is connected to the second input of the amplitude comparator circuit by means of the first return line and the first input of this amplitude comparator circuit is connected to an input of the arrangement and to the first input of the phase comparator circuit, that the control circuit further comprises a modulator and a phase shifting network, the modulator having a first and a second input and an output, that the output of the amplitude comparator circuit is connected to the first input of the modulator and the output circuit is connected to the second input of the modulator *via* the phase shifting network and that the output of the modulator is connected to the control input of the first oscillator through the first low-pass filter.

5. An arrangement as claimed in Claim 3 or 4, characterized in that the output circuit comprises a first and a second class-C amplifier, a first and a second coupling impedance and a third low-pass filter, that the input of the first amplifier is connected to the output of the first oscillator and the input of the second amplifier is connected to the output of the second oscillator and the input of the third low-pass filter is connected to the output of the first amplifier through the first coupling impedance and to the output of the second amplifier through the second coupling impedance, the output of the third low-pass filter being connected to the output of the output circuit.

6. An arrangement as claimed in Claim 3, Claim 4 or Claim 5, characterized in that the first and the second return line each have an isolating amplifier and a coupling network.

7. An arrangement for amplifying a modulated carrier signal, substantially as herein-before described with reference to the accompanying drawings.

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