Model Answer of Communication System-I

B. Tech. V Sem (ECE - 3105)

Part - A

(I) - (b) TV Broad casting

(II) - (d) Balance Modulator.

(III) - (b) lower bandwidth is required

(IV) - (c) 180 KHZ

(V) - (b) Permit better adjacent channel rejection

(VI) - (b) Amplifying the higher audio frequencies.

(VII) - (d) 72

(VIII) - $\frac{S_o}{N_o} = \frac{3}{2}\beta^2 \cdot \frac{S_i}{N M}$

(IX) - 4.16

(X) (b) random nature of electrons

Ans:— 3 (c) we know that all odd numbered sideband pairs give rise to phasors

$$A_n = J_n(\beta) \sin n\omega_m t \quad \text{where } n \text{ is odd}$$

which are perpendicular to the carrier phasor and all even numbered sideband pairs give rise to phasors

$$A_n = J_n(\beta) \cos n\omega_m t \quad n \text{ is even}$$

which are parallel to the carrier phasor

So

$$A_0 = J_0(6) \cos \omega_c t = .15 \cos \omega_c t = .15$$

$$A_1 = J_1(6) \sin \omega_m t = -.28 \sin \omega_m t = -.28$$

$$A_2 = J_2(6) \cos 2\omega_m t = -.24 \cos \cancel{2} \times \pi/\cancel{2} = .24$$

$$= .24$$

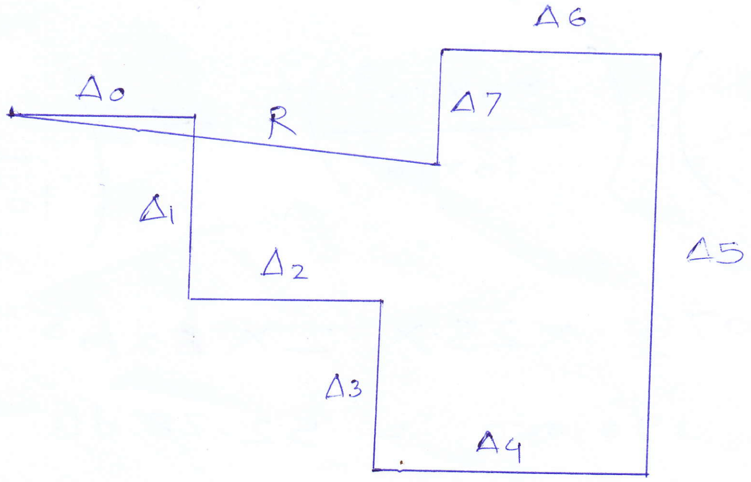
$$A_3 = J_3(6) \sin 3\omega_m t = .11 \sin 3\pi/2 = -.11$$

$$A_4 = J_4(6) \cos 4\omega_m t = .36 \cos 4 \times \pi/2 = .36$$

$$A_5 = J_5(6) \sin 5\omega_m t = .36 \sin 5 \times \pi/2 = .36$$

$$A_6 = J_6(6) \cos \omega_m t = .24 \cos 6 \times \pi/2 = -.24$$

$$A_7 = J_7(6) \sin 7\omega_m t = .13 \sin 7 \times \pi/2 = -.13$$



Ans - 5 (b)

$$S_o/N_o = \left(\frac{3}{4\pi^2}\right) \cdot \left(\frac{k^2 \overline{m^2(t)}}{f_M^2}\right) \cdot \left(\frac{A^2/2}{\eta f_M}\right)$$

here, the last term in the bracket represents received input signal to Noise ratio as its numerator gives strength of signal ~~si/~~ ($s_i = A^2/2$) and denominator strength of Noise in the band of intrest.

The numerator in the second term is related to maxinum freq. deviation Δf.

Now $\Delta f = |f - f_c| = |f_c + k|m(t)|_{max} - f_c|$

$= k|m(t)|_{max} = k$ since, message $|m(t)| \leq 1$

where f = instantaneous freq., f_c = carrier freq.

Thus the above eqn. can be written as

$$\left(\frac{S_o}{N_o}\right) = \left(\frac{3}{4\pi^2}\right) \cdot \left(\frac{\Delta f}{f_m}\right)^2 \cdot \overline{m^2(t)} \cdot \left(\frac{S_i}{\eta f_M}\right)$$

$$= \left(\frac{3}{4\pi^2}\right) \cdot \left(\frac{50\times10^3}{10\times10^3}\right)^2 \cdot \cancel{.5}\ .2 \left(\frac{.5}{10^{-10} \times 10\times10^3}\right)$$

$$= .076 \times 25 \times .2 \times 5\times10^5$$

$$= 190{,}000 = 52.78 \text{ dB}$$

(b) Required SNR at output > 40 dB

from (a) $\quad S_i/.5 \geq \frac{10000}{190000}$

$$S_i = \frac{.5 \times 10000}{190000} = .026 \text{ W}$$

Since channel has 30 dB (1000) loss, required transmitter power > $.026 \times 1000 = 26.31$ W

Ans- 6 (a)

(1) $$B_N = \frac{1}{2g_{a0}} \int_{-\infty}^{\infty} g_a(f)\, df$$

and

$$G_a(f) = \frac{k}{2}(T_{ant} + T_e)$$

$$= \frac{1.38 \times 10^{-23}}{2}(4 + 100)$$

$$G_a(f) = 52 \times 1.38 \times 10^{-23} = 7.17 \times 10^{-22}$$

The Noise Bandwidth of the RC filter is $\pi/2$ times its 3 dB bandwidth f_c.

$$B_N = \frac{\pi}{2} f_c \qquad \text{where } f_c = 3 \text{ dB bandwidth}$$

So $$B_N = \frac{\pi}{2} \times 10^7$$

$$= 1.57 \times 10^7 = 15.7 \times 10^6 \text{ Hz}$$

(ii) and $$P_{ao} = g_{ao}\, k\, (T_{ant} + T_e) \cdot B_N$$

$$= 10^{10} \times 1.38 \times 10^{-23} \times (104) \times 15.7 \times 10^6$$

$$= 2.25 \times 10^{-4} = 225 \times 10^{-6} = 225\ \mu\text{W}$$

Ans- 4 (c) image freq. is given by

$$f_{si} = f_s + 2f_i$$

where f_s = station freq.

f_i = intermediate freq.

f_{si} = image freq.

~~wh~~ given $f_i = 455 \text{ kHz}$

$f_s = 1050 \text{ kHz}$

So $f_{si} = (1050 + 2 \times 455) \text{ kHz}$

$= 1050 + 910$

$= 1960 \text{ kHz}$

$$\rho = \frac{f_{si}}{f_s} - \frac{f_s}{f_{si}}$$

$$= \frac{1960}{1050} - \frac{1050}{1960}$$

$= 1.866 - .536$

$= 1.87 - .54$

$= 1.33$

rejection ratio α is given by

$$\alpha = \sqrt{1 + Q^2 \rho^2}$$

$$= \sqrt{1 + 100^2 \times 1.33^2}$$

$\sqrt{17691} = 133$

$\alpha = 133$

(b) for 20 MHZ

$$f_{si} = 20\times10^6 + 2\times455\times10^3$$

$$= (20 + .91)\,10^6 = 20.91\ MHZ$$

and $\rho = \dfrac{20.91}{20} - \dfrac{20}{20.91}$

$$= 1.045 - .956$$

$$= \cancel{1.05}\ .089$$

and $\alpha = \sqrt{1 + Q^2\rho^2}$

$$= \sqrt{1 + 100^2\times(.089)^2}$$

$$= \sqrt{1+79.21} = \sqrt{80.21} = 8.956$$

$$\alpha = 8.96$$

Ans- 2(a)

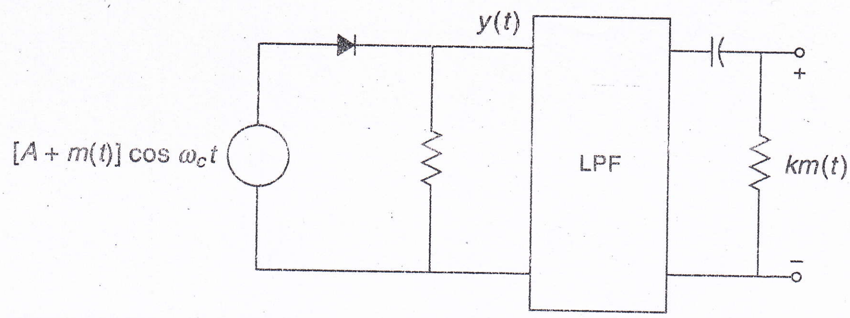


Fig. 3.10 Rectifier detector circuit for AM (DSB-C).

Thus $y(t)$ has a dc term A/π coming from $\cos^2 \omega_c t$ term and $m(t)$ between 0 to ω_c which is passed by LPF. And after dc blocker capacitor we only get the demodulated message signal.

The square law demodulator: An alternative method of recovering the baseband signal which has been superimposed as an amplitude modulation on a carrier is to pass the AM signal through a nonlinear device. Such demodulation is illustrated in Fig. 3.11. We assume here for simplicity that the device has a square-law relationship between input signal x (current or voltage) and output signal y (current or voltage). Thus $y = kx^2$, with k a constant. Because of the nonlinearity of the transfer characteristic of the device, the output response is different for positive and for negative

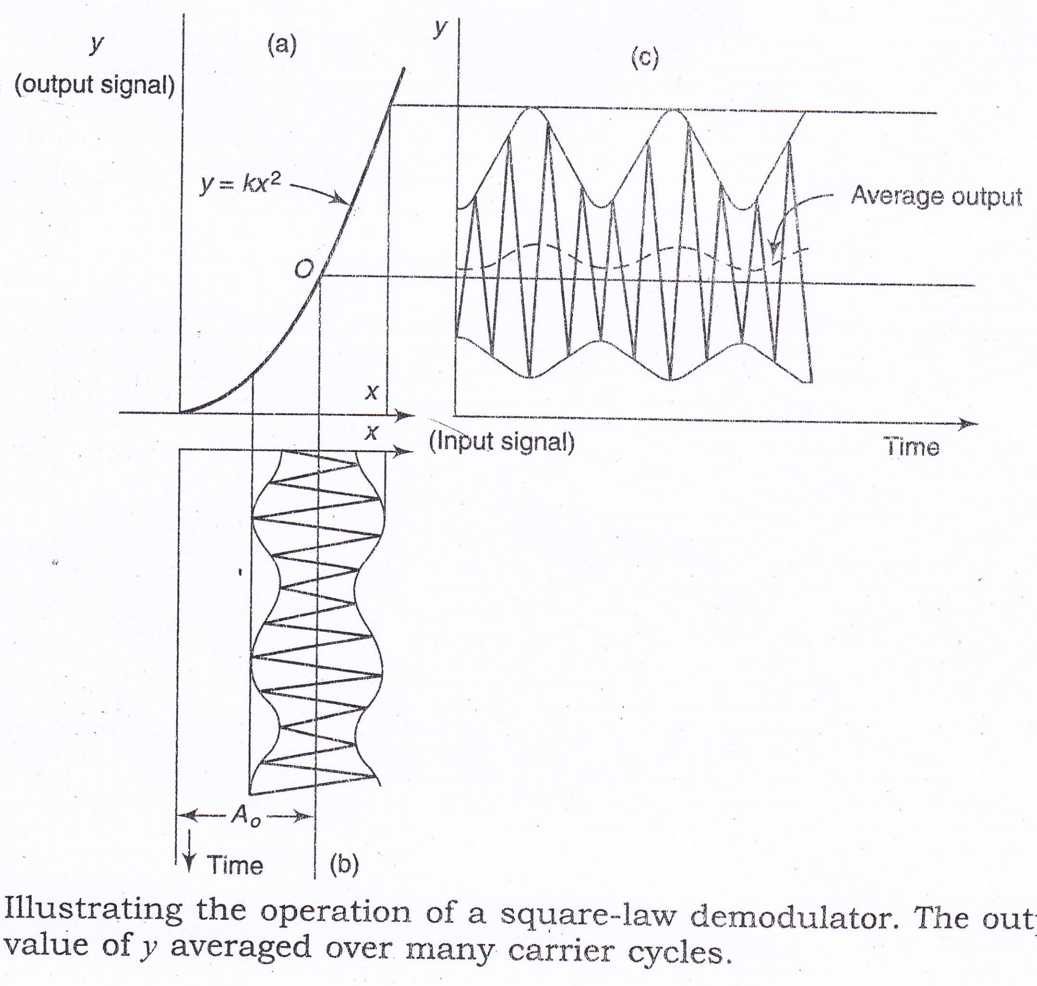


Fig. 3.11 Illustrating the operation of a square-law demodulator. The output is the value of y averaged over many carrier cycles.

Ans- 2(a)

excursions of the carrier away from the quiescent operating point O of the device. As a result, and as is shown in Fig. 3.11c, the output, when averaged over a time which encompasses many carrier cycles but only a very small part of the modulation cycle, has the waveshape of the envelope.

The applied signal is

$$x = A_o + A_c[1 + m(t)] \cos \omega_c t \tag{3.11}$$

Thus the output of the squaring circuit is

$$y = k\{A_o + A_c[1 + m(t)] \cos \omega_c t\}^2 \tag{3.12}$$

Squaring, and dropping dc terms as well as terms whose spectral components are located near ω_c and $2\omega_c$, we find that the output signal $s_o(t)$, that is, the signal output of a low-pass filter located after the squaring circuit, is

$$s_o(t) = kA_c^2\left[m(t) + \frac{1}{2}m^2(t)\right] \tag{3.13}$$

Observe that the modulation $m(t)$ is indeed recovered but that $m^2(t)$ appears as well. Thus the total recovered signal is a distorted version of the original modulation. The distortion is small, however, if $\frac{1}{2}m^2(t) \ll |m(t)|$ or if $|m(t)| \ll 2$.

There are two points of interest to be noted in connection with the type of demodulation described here; the first is that the demodulation does not depend on the nonlinearity being square-law. Any type of nonlinearity which does not have odd-function symmetry with respect to the initial operating point will similarly accomplish demodulation. The second point is that even when demodulation is not intended, such demodulation may appear incidentally when the modulated signal is passed through a system, say, an amplifier, which exhibits some nonlinearity.

3.3.3 Maximum Allowable Modulation for Rectifier Detection

If we are to avail ourselves of the convenience of demodulation by the use of the simple diode circuit of Fig. 3.9, we must limit the extent of the modulation of the carrier. That such is the case may be seen from Fig. 3.12. In Fig. 3.12a is shown a carrier modulated by a sinusoidal signal. It is apparent that the envelope of the carrier has the waveshape of the modulating signal. The modulating signal is sinusoidal; hence $m(t) = m \cos \omega_m t$, where m is a constant. Equation (3.8) becomes

$$v(t) = A_c(1 + m \cos \omega_m t) \cos \omega_c t \tag{3.14}$$

In Fig. 3.12b we have shown the situation which results when, in Eq. (3.14), we adjust $m > 1$. Observe now that the diode demodulator which yields as an output the positive envelope (a negative envelope if the diode is reversed) will not reproduce the sinusoidal modulating waveform. In this latter case, where $m > 1$, we may recover the modulating waveform but not with the diode modulator. Recovery would require the use of a coherent demodulation scheme such as was employed in connection with the signal furnished by a multiplier.

It is therefore necessary to restrict the excursion of the modulating signal in the direction of decreasing carrier amplitude to the point where the carrier amplitude is just reduced to zero. No such similar restriction applies when the modulation is increasing the carrier amplitude. With sinusoidal modulation, as in Eq. (3.14), we require that $|m| \leq 1$. More generally in Eq. (3.8) we require that the maximum negative excursion of $m(t)$ be -1.

Ans - 2 (b)

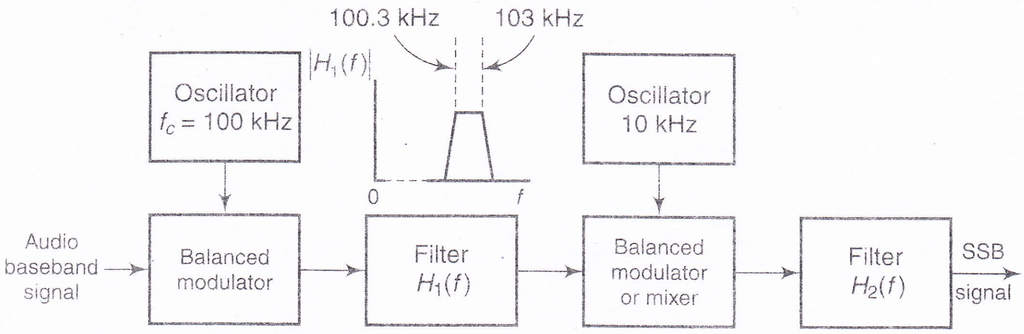


Fig. 3.16 Block diagram of the filter method of generating a single-sideband signal.

suppress the carrier itself. Of course, in principle, no carrier should appear at the output of a balanced modulator. In practice, however, the modulator may not balance exactly, and the precision of its balance may be subject to some variation with time. Therefore, even if a pilot carrier is to be transmitted, it is well to suppress it at the output of the modulator and to add it to the signal at a later point in a controllable manner.

Now consider that we desire to generate an SSB signal with a carrier of, say, 10 MHz. Then we require a passband filter with a selectivity that provides 40 dB of attenuation within 600 Hz at a frequency of 10 MHz, a percentage frequency change of 0.006 percent. Filters with such sharp selectivity are very elaborate and difficult to construct. For this reason, it is customary to perform the translation of the baseband signal to the final carrier frequency in several stages. Two such stages of translation are shown in Fig. 3.16. Here we have selected the first carrier to be of frequency 100 kHz. The upper sideband, say, of the output of the balanced modulator ranges from 100.3 to 103 kHz. The filter following the balanced modulator which selects this upper sideband need now exhibit a selectivity of only a hundredth of the selectivity (40 dB in 0.6 percent frequency change) required in the case of a 10 MHz carrier. Now let the filter output be applied to a second balanced modulator, supplied this time with a 10 MHz carrier. Let us again select the upper sideband. Then the second filter must provide 40 dB of attenuation in a frequency range of 200.6 kHz, which is nominally 2 percent of the carrier frequency.

We have already noted that the simplest physical frequency-translating device is a multiplier or mixer, while a balanced modulator is a balanced arrangement of two mixers. A mixer, however, has the disadvantage that it presents at its output not only sum and difference frequencies but the input frequencies as well. Still, when it is feasible to discriminate against these input signals, there is a merit of simplicity in using a mixer rather than a balanced modulator. In the present case, if the second frequency-translating device in Fig. 3.16 were a mixer rather than a multiplier, then in addition to the upper and lower sidebands, the output would contain a component encompassing the range 100.3 to 103 kHz as well as the 10 MHz carrier. The range 100.3 to 103 kHz is well out of the range of the second filter intended to pass the range 10,100,300 to 10,103,000 Hz. And it is realistic to design a filter which will suppress the 10 MHz carrier, since the carrier frequency is separated from the lower edge of the upper sideband (10,100,300) by nominally a 1-percent frequency change.

Altogether, then, we note in summary that when a single-sideband signal is to be generated which has a carrier in the megahertz or tens-of-megahertz range, the frequency translation is to be done in more than one stage—frequently two but not uncommonly three. If the baseband signal has spectral components in the range of hundreds of hertz or lower (as in an audio signal), the first stage invariably employs a balanced modulator, while succeeding stages may use mixers.

1.1.3 Modulation and Multiplexing

Ans - 2 (c)

As discussed in Sec. 1.1.1, modulation is the process of varying one attribute of a signal by message signal. In electronic communication modulation plays a very important role and a good number of chapters in any communication text is devoted to it. Why at all we need to modulate the message signal? There are several reasons. The first comes from an important practical consideration. For wireless communication the electromagnetic wave needs to be radiated. This requires an antenna the diameter of which is required to be approximately one-tenth of signal wavelength or of that order. Now, let us do a quick calculation.

> If c is speed of light and frequency of electromagnetic wave to be transmitted is f, then its wavelength $\lambda = c/f$.
>
> To radiate a signal of frequency 4 kHz the antenna diameter needs to be $0.1 \times 3 \times 10^8/4000 =$ 75000 meter or 7.5 km while a 1 GHz signal needs an antenna size of $0.1 \times 3 \times 10^8/10^9 =$ 0.03 meter or 3 cm.

Of course, the first one is impractical and we should have a mechanism to translate original baseband signal to a very high frequency to reduce the antenna size. This is usually achieved by linearly or proportionately varying one attribute of a high frequency signal called *carrier* (called so as *carries message* or *message rides on it*) be it amplitude, frequency or phase. Note that, such frequency translation is required at the transmitter end for both analog and digital signal before the signal is put to channel unless it is a baseband transmission.

There is another advantage behind modulating message signal. The original message signal occupies lower baseband frequencies, which for example in speech is 0.3 to 3.4 kHz. If two persons want to communicate over same channel simultaneously then two baseband signals will interfere with each other. However, if one signal is placed between 0 to 4 kHz and the other say 4 to 8 kHz then there is no such conflict. Using a carrier to shift the frequency band of a message signal for simultaneous transmission is known as frequency division multiplexing (FDM). For digital signal, multiplexing can be achieved by dividing the time between two samples of signals in various time slots and using each time slot to send one digital signal. Such a method of simultaneous transmission is known as time division multiplexing (TDM).

Ans- 3 (a)

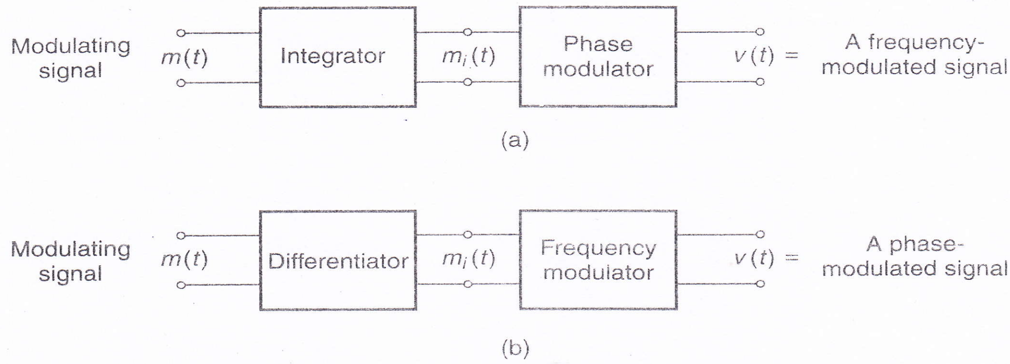


Fig. 4.2 Illustrating the relationship between phase and frequency modulation.

in which k'' is also a constant. Then with $k = k'k''$ we have

$$v(t) = A \cos\left[\omega_c t + k\int_{-\infty}^{t} m(t)\, dt\right] \tag{4.7}$$

The instantaneous angular frequency is

$$\omega = \frac{d}{dt}\left[\omega_c t + k\int_{-\infty}^{t} m(t)\, dt\right] = \omega_c + km(t) \tag{4.8}$$

The deviation of the instantaneous frequency from the carrier frequency $\omega_c/2\pi$ is

$$v \equiv f - f_c = \frac{k}{2\pi} m(t) \tag{4.9}$$

Since the deviation of the instantaneous frequency is directly proportional to the modulating signal, the combination of *integrator* and *phase modulator* of Fig. 4.2*a* constitutes a device for producing a *frequency-modulated* output. Similarly, the combination in Fig. 4.2*b* of the differentiator and frequency modulator generates a *phase-modulated output*, i.e. a signal whose phase departure from the carrier is proportional to the modulating signal.

In summary, we have referred generally to the waveform given by Eq. (4.1) as an *angle-modulated* waveform, an appropriate designation when we have no interest in, or information about, the modulating signal. When $\phi(t)$ is proportional to the modulating signal $m(t)$, we use the designation *phase modulation* or PM. When the time derivative of $\phi(t)$ is proportional to $m(t)$, we use the term *frequency modulation* or FM. In an FM waveform, the form of Eq. (4.7) is of special interest, since here the instantaneous frequency deviation is directly proportional to the signal $m(t)$ which appears explicitly in the expression. In general usage, however, we find that such precision of language is not common. Very frequently the terms angle modulation, phase modulation and frequency modulation are used rather interchangeably and without reference to, or even interest in, the modulating signal.

4.1.3 Phase and Frequency Modulation

In the waveform of Eq. (4.1) the maximum value attained by $\phi(t)$, that is, the maximum phase deviation of the total angle from the carrier angle $\omega_c t$, is called the *phase deviation*. Similarly the maximum departure of the instantaneous frequency from the carrier frequency is called the *frequency deviation*.

Ans- 3 (a)

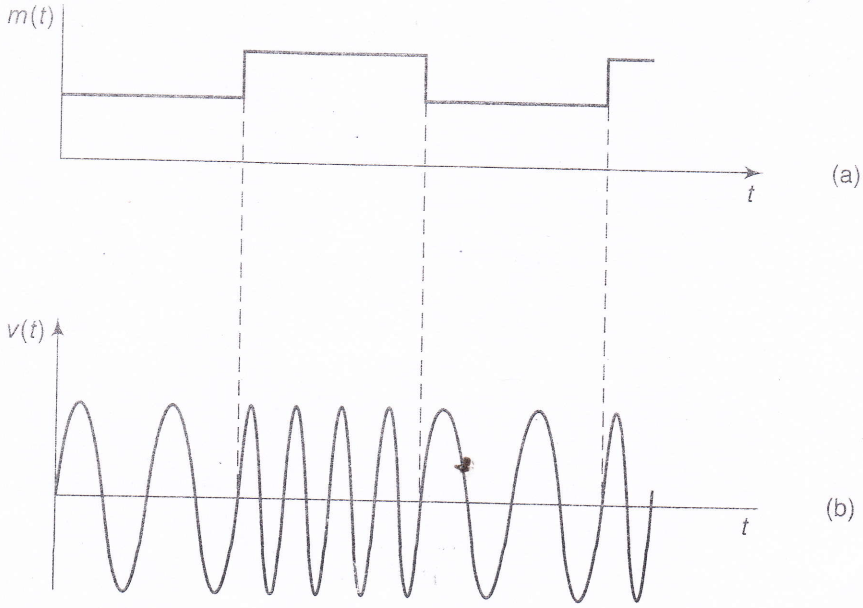


Fig. 4.1 An angle-modulated waveform. (a) Modulatign signal. (b) Frequently-modulated sinusodial carrier signal.

Among the possibilities which suggest themselves for the design of a modulator are the following. We might arrange that the phase $\phi(t)$ in Eq. (4.1) be directly proportional to the modulating signal, or we might arrange a direct proportionality between the modulating signal and the derivative, $d\phi(t)/dt$. From Eq. (4.3), with $f_c = \omega_c/2\pi$

$$\frac{d\phi(t)}{dt} = 2\pi(f - f_c) \tag{4.4}$$

where f is the instantaneous frequency. Hence in this latter case the proportionality is between modulating signal and the departure of the instantaneous frequency from the carrier frequency. Using standard terminology, we refer to the modulation of the first type as *phase* modulation, and the term *frequency modulation* refers only to the second type. On the basis of these definitions it is, of course, not possible to determine which type of modulation is involved simply from a visual examination of the waveform or from an analytical expression for the waveform. We would also have to be given the waveform of the modulating signal. This information is, however, provided in any practical communication system.

4.1.2 Relationship Between Phase and Frequency Modulation

The relationship between phase and frequency modulation may be visualized further by a consideration of the diagrams of Fig. 4.2. In Fig. 4.2*a* the phase-modulator block represents a device which furnishes an output $v(t)$ which is a carrier, phase-modulated by the input signal $m_i(t)$. Thus

$$v(t) = A \cos[\omega_c t + k' m_i(t)] \tag{4.5}$$

k' being a constant. Let the waveform $m_i(t)$ be derived as the integral of the modulating signal $m(t)$ so that

$$m_i(t) = k'' \int_{-\infty}^{t} m(t)\, dt \tag{4.6}$$

Ans-3(b)

to 5. Where higher orders of multiplication are required, multipliers may be cascaded. Cascaded multipliers of order n_1, n_2, n_3, ... yield an overall multiplication of order $n_1 n_2 n_3$ ....

Now let us consider the result of the application to a frequency multiplier of an FM signal. If we think of an FM signal as a "sinusoidal" signal in which the frequency changes from moment to moment, then at each instant we may expect that the output frequency will be n times the input frequency. Hence, if the input consists of a carrier of frequency f_c which ranges through a frequency deviation $\pm\Delta f$, the output will have a carrier frequency nf_c and will range through the deviation $\pm n\ \Delta f$. The multiplier multiplies both the carrier and the deviation frequency. Also, since the modulation index is proportional to the frequency deviation, for a fixed modulation frequency, the multiplier increases the modulation index by the same factor n.

By way of example, consider the case of commercial FM broadcasting in the United States. Here, the allowable frequency deviation is 75 kHz, so that for a modulation frequency $f_m = 50$ Hz, $\beta = \Delta f/f_m = 1500$. Even if we allow ϕ in Fig. 4.6a to attain a maximum value as large as $\phi = 0.5$, then the multiplication needed is $1500/0.5 = 3000$. On the other hand, at the high-frequency end of the baseband spectrum, say, $f_m = 15$ kHz, $\beta = 75/15 = 5$. Correspondingly, with a multiplication by a factor of 3000, the phase ϕ in Fig. 4.6a need to attain only a maximum value $5/3000 = 1.7 \times 10^{-3}$. Thus, it is to be seen that the required multiplication is determined by the low-frequency limit of the baseband spectrum.

4.4.4 An Example of an Armstrong FM System

The block diagram of Fig. 4.11 represents an Armstrong FM system which supplies a signal whose carrier is at 96 MHz (which is near the center of the commercial FM broadcasting band). It allows direct *phase modulation* of the carrier, before multiplication, to the extent of $\phi_m = 0.5$. Thus, at $f_m = 50$ Hz, we have $\Delta f = 25$ Hz. Note that at higher modulating frequencies, ϕ_m is less than 0.5 rad. The carrier frequency before multiplication has been selected at 200 kHz. a frequency at which very stable crystal oscillators and balanced modulators are readily constructed.

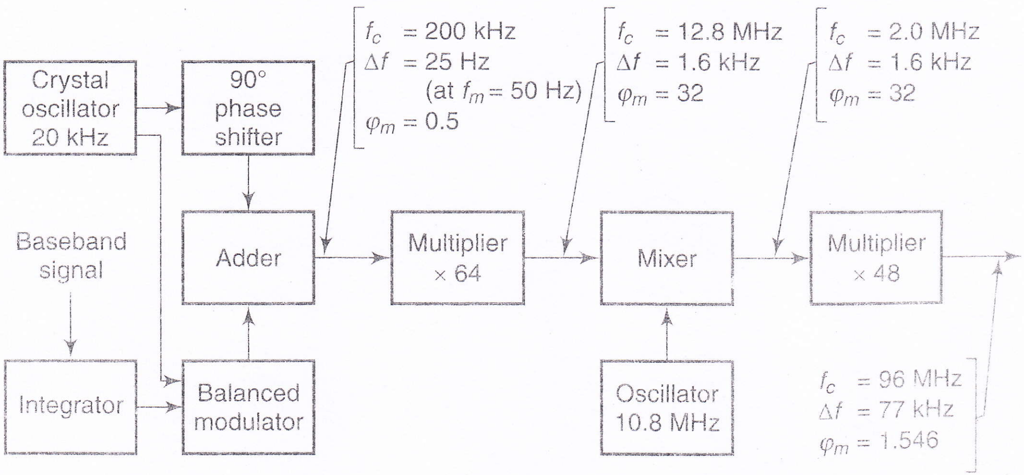


Fig. 4.11 Block diagram of an Armstrong system of generating an FM signal using multipliers to increase the frequency deviation.

As already noted, if we require that $\Delta f \approx 75$ kHz, then a multiplication by a factor of 3000 is required. In Fig. 4.11 the multiplication is actually $3072 (= 64 \times 48)$. The values were selected so

4.4.2 FM Generation by Armstrong's Indirect Method

Ans-3(b)

A phase-modulated waveform in which the modulating waveform is $m(t)$ is written $\cos[\omega_c t + m(t)]$. If the modulation is narrowband [$|m(t)| \ll 1$], then we may use the approximation

$$\cos[\omega_c t + m(t)] \cong \cos \omega_c t - m(t) \sin \omega_c t \tag{4.66}$$

The term $m(t) \sin \omega_c t$ is a DSB-SC waveform in which $m(t)$ is the modulating waveform and $\sin \omega_c t$ the carrier. We note that the carrier of the FM waveform, that is, $\cos \omega_c t$, and the carrier of the DSB-SC waveform are in quadrature. We may note in passing that if the two carriers are in phase, the result is an AM signal since

$$\cos \omega_c t + m(t) \cos \omega_c t = [1 + m(t)] \cos \omega_c t \tag{4.67}$$

A technique used in commercial FM systems to generate NBFM, which is based on our observation in connection with Eq. (4.66), is shown in Fig. 4.9. Here a balanced modulator is employed to generate the DSB-SC signal using $\sin \omega_c t$ as the carrier of the modulator. This carrier is then shifted in phase by 90° and, when added to the balanced modulator output, thereby forms an NBFM signal. However, the signal so generated will be phase-modulated rather than frequency-modulated. If we desire that the frequency rather than the phase be proportional to the modulation $m(t)$, then, as discussed in Sec. 4.1.2 and illustrated in Fig. 4.2, we need merely integrate the modulating signal before application to the modulator.

If the system of Fig. 4.9 is to yield an output signal whose phase deviation is directly proportional to the amplitude of the modulating signal, then the phase deviation must be kept small. That

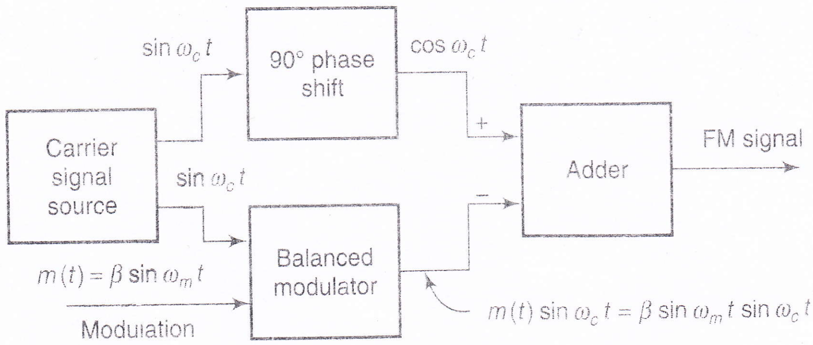


Fig. 4.9 Illustrating the principle of the Armstrong system of generating a PM signal.

Ans - 4(a)

Fig. 6.1 gives the block diagram of amplitude modulation radio transmitter using modulation at high power level of the carrier. The function of each constituent stage is briefly given below :

(*i*) **Master Oscillator.** It generates oscillations of desired frequency with high constancy of frequency. The generated frequency is required to remain constant within close limits inspite of variations in the supply voltage, ambient temperature, temperature of components of load. Further frequency variations with time and with age of the tube (or transistor) are to be avoided.

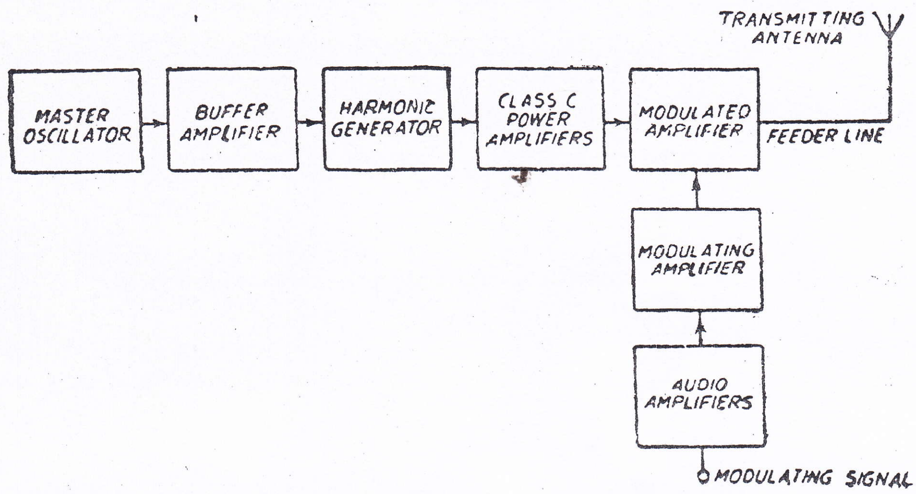


Fig. 6.1. Block diagram of amplitude modulation radio transmitter using modulation at high carrier power level.

(*ii*) **Buffer Amplifier or Isolating Amplifier.** If the master oscillator directly drives a harmonic generator or class C power amplifier, which may draw input current (grid current or base current in *CE* amplifier), then power is drawn from the master oscillator. This results in loading of master oscillator which in turn causes variation of effective resistance of the tank circuit of the oscillator and hence results in frequency variation. Accordingly a buffer amplifier or isolating amplifier is placed between the master oscillator and the harmonic generators. This buffer amplifier does not draw any input current and hence causes no loading of the master oscillator. Changes in carrier frequency due to variations in loading are thus avoided.

(*iii*) **Harmonic Generators.** Usually master oscillator generates voltage at a frequency which is a sub-multiple of the carrier frequency *i.e.* the frequency of the radiated power. Basically these harmonic generators are class C tuned amplifiers in which the output R.F. voltage is first distorted through class C operation and then the tuned circuit in the output circuit of amplifier selects the desired harmonic frequency.

(*iv*) **Class C Amplifiers.** R.F. voltage generated by the master oscillator has usually very small power, of the order of a few watts. The power level is required to be raised to the final high value in a chain of class C amplifier having high output circuit efficiency of the order of 70%. In general, first few stages of class C amplifier act as harmonic generators as well.

(*v*) **Modulated Amplifier.** This is a class C tuned amplifier usually of pushpull type and is modulated by audio modulating voltage from modulating amplifier. High efficiency series plate modulation is most popularly used in high power radio broadcast and radio telephone transmitters. Grid bias modulation and suppressor grid modulation are sometimes used particularly for modulation at low power levels. In small transistorized radio transmitter, collector modulation or base modulation or both may be used.

(*vi*) **Modulating Amplifier.** This is usually a class B pushpull amplifier and feeds audio power into the modulation amplifier in the plate circuit, control grid circuit or suppressor grid circuit depending upon the method of modulation used. Class B operation is generally used because of high plate circuit efficiency. However, class A modulating amplifier are also sometimes used particularly in low power transmitters.

9.7. Constituent Stages of a Superheterodyne Receiver

Ans - 4 (b)

Fig. 9.3 gives the block diagram of a superheterodyne receiver. Functions of different stages are briefly given below :

Antenna or Aerial. It intercepts the electromagnetic waves. Voltages induced in the antenna are communicated to the receiver input circuit by means of a feeder wire or lead-in wire. A parallel tuned circuit at the input of the receiver responds only to voltage at the desired carrier frequency and rejects voltages at all other frequencies. The voltage so picked up is fed to the input of the RF amplifier stage.

R.F. Amplifier. This stage is generally a tuned voltage (small-signal) amplifier tuned to the desired carrier frequency. The chief functions of RF amplifier stage are :

(*i*) To amplify the input signal voltage to a suitably high level before feeding it to the frequency mixer which contributes large noise. Thus signal/noise ratio is improved.

(*ii*) To provide discrimination or selectivity against image frequency signal and intermediate frequency signal.

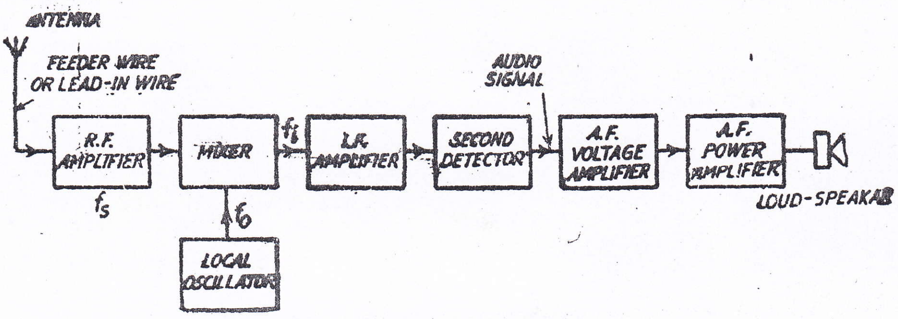


Fig. 9.3. Block diagram of superheterodyne receiver.

Frequency Converter Stage. This consists of a local oscillator and frequency mixer. To the frequency mixer are fed both the local oscillator voltage as well as signal voltage. The mixer, being a non-linear device, produces at its output the various intermodulation terms. The difference frequency voltage is picked up by the tuned circuit in the output circuit of the mixer. This difference frequency is called the intermediate frequency, the value of which is constant for a receiver. For all-wave receivers, typical value of intermediate frequency is 465 kHz or 456 kHz. Sometimes two separate transistors are used as local oscillator and frequency mixer but more often only one transistor functions both as local oscillator and frequency mixer. Such a transistor is then referred to as a frequency converter transistor. Thus with the help of frequency converter stage, RF signal of any carrier frequency is converted into similarly modulated fixed frequency IF signal.

[redacted] Ans-5(a)

If the modulation is sinusoidal, with $m(t) = m \cos 2\pi f_m t$ (m a constant), then

$$s_i(t) = A(1 + m \cos 2\pi f_m t) \cos 2\pi f_c t \tag{8.39}$$

In this case $\overline{m^2(t)} = m^2/2$ and

$$\frac{S_o}{N_o} = \frac{m^2}{2+m^2}\frac{S_i}{\eta f_M} \tag{8.40}$$

When the carrier is transmitted only to synchronize the local demodulator waveform $\cos 2\pi f_c t$, relatively little carrier power need be transmitted. In this case $m \gg 1$, $m^2/(2 + m^2) \cong 1$, and the signal-to-noise ratio is not greatly reduced by the presence of the carrier. On the other hand, when envelope demodulation is used (Sec. 3.3), it is required that $m \leq 1$. When $m = 1$, the carrier is 100 percent modulated. In this case $m^2/(2 + m^2) = \frac{1}{3}$, so that of the power transmitted, only one-third is in the sidebands which contribute to signal power output.

A Figure of Merit

We observe that in each demodulation system considered so far, the ratio $S_i/\eta f_M$ appeared in the expression for output SNR [see Eqs. (8.12), (8.20), and (8.37)]. This ratio is the output signal power S_i divided by the product ηf_M. To give the product ηf_M some physical significance, we consider it to be the noise power N_M at the input, measured in a frequency band equal to the *baseband frequency*. Thus

$$N_M \equiv \frac{\eta}{2} 2f_M = \eta f_M \tag{8.41}$$

The ratio $S_i/\eta f_M$ is, therefore, often referred to as the *input signal-to-noise ratio* S_i/N_M. It needs to be kept in mind that N_M is the noise power transmitted through the IF filter only when the IF filter bandwidth is f_M. Thus N_M is the true input noise power only in the case of single sideband.

For the purpose of comparing systems, we introduce the *figure of merit* γ, defined by

$$\gamma \equiv \frac{S_o/N_o}{S_i/N_M} \tag{8.42}$$

The results given above may now be summarized as follows:

$$\gamma = \begin{cases} 1 & \text{SSB-SC} \quad (8.43) \\ 1 & \text{DSB-SC} \quad (8.44) \\ \dfrac{\overline{m^2(t)}}{1+\overline{m^2(t)}} & \text{DSB} \quad (8.45) \\ \dfrac{m^2}{2+m^2} & \text{DSB with sinusoidal modulation} \quad (8.46) \end{cases}$$

8.4 DOUBLE SIDEBAND WITH CARRIER (DSB-C) Ans - 5(a)

Let us now consider the case where a carrier accompanies the double-sideband signal. Demodulation is achieved synchronously as in SSB-SC and DSB-SC. The carrier is used as a *transmitted reference* to obtain the reference signal cos $\omega_c t$ (see Prob. 8.10). We note that the carrier increases the total input-signal power but makes no contribution to the output-signal power. Equation (8.20) applies directly to this case, provided only that we replace S_i by $S_i^{(SB)}$, where $S_i^{(SB)}$ is the power in the sidebands alone. Then

$$\frac{S_o}{N_o} = \frac{S_i^{(SB)}}{\eta f_M} \tag{8.33}$$

Suppose that the received signal is

$$\begin{aligned} s_i(t) &= A[1 + m(t)] \cos 2\pi f_c t \\ &= A \cos 2\pi f_c t + Am(t) \cos 2\pi f_c t \end{aligned} \tag{8.34}$$

where $m(t)$ is the baseband signal which amplitude-modulates the carrier $A \cos 2\pi f_c t$. The carrier power is $A^2/2$. The sidebands are contained in the term $Am(t) \cos 2\pi f_c t$. The power associated with this term is $(A^2/2)\overline{m^2(t)}$, where $\overline{m^2(t)}$ is the time average of the square of the modulating waveform. We then have that the total input power S_i is given by

$$S_i = \frac{A^2}{2} + S_i^{(SB)} = \frac{A^2}{2}[1 + \overline{m^2(t)}] \tag{8.35}$$

Eliminating A^2, we have

$$S_i^{(SB)} = \frac{\overline{m^2(t)}}{1 + \overline{m^2(t)}} S_i \tag{8.36}$$

or, with Eq. (8.33),

$$\frac{S_o}{N_o} = \frac{\overline{m^2(t)}}{1 + \overline{m^2(t)}} \frac{S_i}{\eta f_M} \tag{8.37}$$

In terms of the carrier power $P_c \equiv A^2/2$, we get, from Eqs. (8.35) and (8.37), that

$$\frac{S_o}{N_o} = \overline{m^2(t)} \frac{P_c}{\eta f_M} \tag{8.38}$$

9.6 THRESHOLD IN FREQUENCY MODULATION Ans - 5(c)

In a communication system in which the modulation is linear and demodulation is accomplished by coherent detection (for example, SSB and DSB-SC), we have the result that [see Eqs. (8.43) and (8.44)]

$$\frac{S_o}{N_o} = \frac{S_i}{N_M} \tag{9.52}$$

or equivalently

$$10 \log \frac{S_o}{N_o} \equiv \left[\frac{S_o}{N_o}\right]_{\text{dB}} = \left[\frac{S_i}{N_M}\right]_{\text{dB}} \equiv 10 \log \frac{S_i}{N_M} \tag{9.53}$$

In FM we have [see Eq. (9.27)]

$$\frac{S_o}{N_o} \equiv \frac{3}{2}\beta^2 \frac{S_i}{N_M} \tag{9.54}$$

or equivalently

$$\left[\frac{S_o}{N_o}\right]_{\text{dB}} = \left[\frac{S_i}{N_M}\right]_{\text{dB}} + 10 \log \frac{3}{2}\beta^2 \tag{9.55}$$

In Fig. 9.12 we have plotted Eq. (9.53) in a coordinate system in which the coordinate axes are marked off in decibels. The plot is a straight line passing through the origin. We have also plotted Eq. (9.55) for two values of $\beta\left(>\sqrt{\frac{2}{3}}\right)$. These plots, as indicated by the dashed extensions, are also straight lines. In the FM case a plot for a value of β is raised by the amount 10 log $(3\beta^2/2)$ above the plot for a linear coherent modulation system. The quantity 10 log $(3\beta^2/2)$ expresses in decibels precisely the improvement afforded by the FM system in return for a sacrifice of bandwidth.

Experimentally it is determined, however, that the FM system exhibits a threshold. Thus, as indicated by the solid-line plots, for each value of β, as S_i/N_M decreases, a point is reached where S_o/N_o falls off much more sharply than S_i/N_M. The threshold value of S_i/N_M is arbitrarily taken to be the value at which S_oN_o falls 1 dB below the dashed extension. We note that for larger β the threshold S_i/N_M is also higher. Suppose, then, that we are operating with a modulation index β_1 above the threshold for β_1 but below the threshold for β_2. Suppose, further, that at this input S_i/N_M we should increase β from β_1 to β_2 hoping thereby to improve the output-signal-to-noise ratio S_o/N_o by a sacrifice of bandwidth. We would find, however, as is apparent from Fig. 9.12, that

Ans- 5(c)

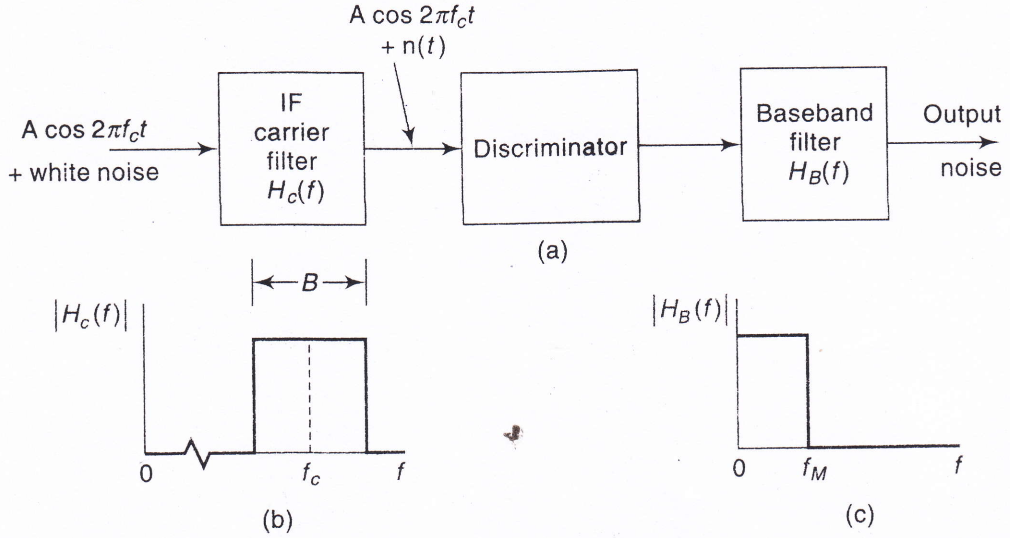


Fig. 9.15 (a) An FM discriminator and associated filters. (b) The bandpass range of the carrier filter. (c) The passband of the baseband filter.

The notation of Eq. (9.56) is especially appropriate when $n(t)$ is to be represented as a combination of phasors in a coordinate system rotating counterclockwise with angular frequency ω_c. In such a coordinate system $n(t)$ is represented by the phasor sum of $x(t)$ in the horizontal direction (i.e. the x direction) and $y(t)$ in the vertical direction (i.e. the y direction). In terms of this new notation, the carrier and noise output of the IF filter, which is the input $v_i(t)$ to the demodulator, is given by

$$v_i(t) = A \cos \omega_c t + x(t) \cos \omega_c t - y(t) \sin \omega_c t \qquad (9.57)$$

This equation can be rewritten in phasor notation as

$$v_i(t) = \text{Re}\ \{[A + x(t)]\, e^{j\omega_c t} + y(t)\, e^{j[\omega_c t + (\pi/2)]}\}$$

$$= \text{Re}\ \{e^{j\omega_c t}\underbrace{[A + x(t) + jy(t)]}_{\text{phasors}}\} \qquad (9.58)$$

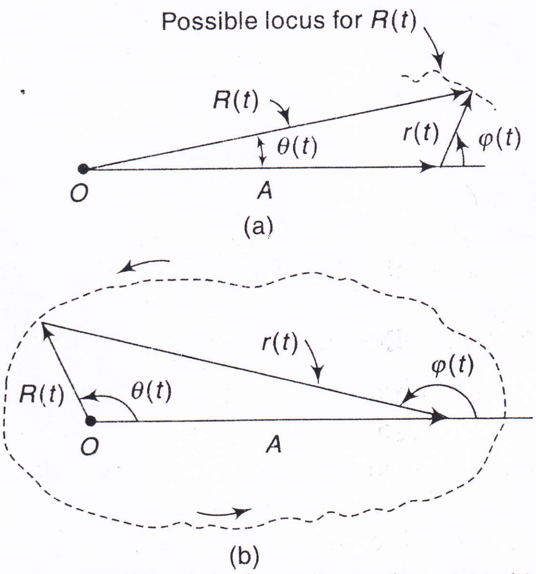


Fig. 9.16 (a) A noise phasor $r(t)$ is added to a carrier phasor of amplitude A. The sum is the resultant phasor $R(t)$. (b) A locus for the end point of $R(t)$ which will give to a spike.

The phasor $A + x(t)$ lies along the horizontal axis and $y(t)$ lies along the vertical axis.

It is convenient, when discussing the occurrence of spikes, to talk about the *amplitude* of the noise $r(t) = \sqrt{x^2 + y^2}$ and its random phase angle $\phi(t) = \tan^{-1}(y(t)/x(t))$. [It was shown in Sec. 2.3.3 that the density of $r(t)$ is Rayleigh and that the density of $\phi(t)$ is $1/2\pi$ from $-\pi$ to $+\pi$.]

The phasor diagram for $v_i(t)$ is shown in Fig. 9.16. Here we see that the phasor sum of signal A and noise $r(t)$ is defined as $R(t)$. The angle that $R(t)$ makes with the horizontal axis is called $\theta(t)$. Then

$$A + x + jy = A + r(t)e^{j\phi(t)} = R(t)e^{j\theta(t)} \qquad (9.59)$$

[redacted] Ans - 5(c)

and that $v_i(t)$ in Eq. (9.16) is

$$v_i(t) = \text{Re}[e^{j\omega_c t} R(t) e^{j\theta(t)}] = R(t) \cos[\omega_c t + \theta(t)] \tag{9.60}$$

As was discussed in Sec. 9.2, the output of the discriminator is proportional to $\dot{\theta}(t) = d\theta/dt$.

Let us now consider Fig. 9.16*a* to see how $\theta(t)$, and hence $\dot{\theta}(t)$, are affected as the noise $r(t)$ and $\phi(t)$ vary. If the noise power is small in comparison with the carrier power, we expect that $r(t) \ll A$ most of the time, and that the end point of the resultant phasor $R(t)$ will never wander far from the end point of the carrier phasor. Thus, as $\phi(t)$ changes, the angle $\theta(t)$ remains small.

If, however, the ratio $S_i/\eta f_M$ decreases, the likelihood of $r(t)$ being much less than A also decreases. When $r(t)$ becomes comparable in magnitude to the carrier amplitude A, the locus of the end point of the resultant phasor $R(t)$ moves away from the end point of the carrier phasor and may, as shown in Fig. 9.16*b*, even rotate about the origin. The variation of the phase angle $\theta(t)$, at and near the time of occurrence of this event, is shown in Fig. 9.17*a*. If the rotation of the end point of $R(t)$ around the origin occurs in the interval between t_1 and t_2, the angle $\theta(t)$ changes by 2π rad during this time interval. Preceding t_1 and following t_2, $r(t) \ll A$, and the usual small random variations of $\theta(t)$ occur.

We are interested in the discriminator output, which is proportional to the instantaneous frequency $d\theta/dt$, and we have therefore plotted $d\theta/dt$ in Fig. 9.17*b*. Notice that when $\theta(t)$ changes by 2π rad, $d\theta/dt$ appears as a sharp *spike* or *impulse* with area 2π. To show that the area under the spike waveform is indeed 2π, we simply integrate $d\theta/dt$ over the time interval, t_1 to t_2 during which $\theta(t)$ changes by 2π. Thus

$$\text{Area} = \int_{t_1}^{t_2} \frac{d\theta}{dt}\, dt = \theta \Big|_{t_1}^{t_2} = 2\pi \tag{9.61}$$

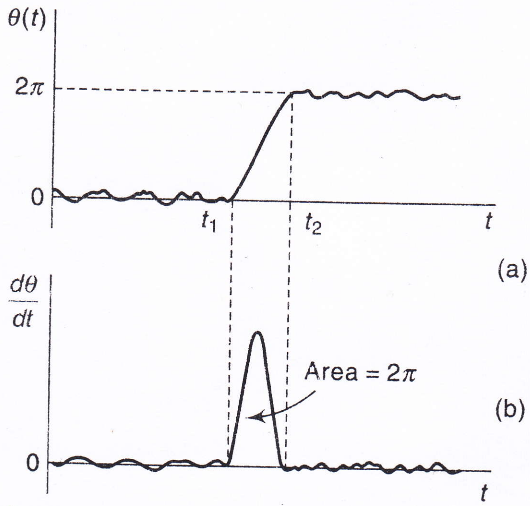


Fig. 9.17 (a) A plot of $\theta(t)$ for a case in which the end point of $R(t)$ in Fig. 9.16*b* executes a rotation around the origin. (b) A plot of $d\theta/dt$ as a function of time.

The waveform shown in Fig. 9.17*b* is the *spike* noise referred to in Sec. 9.6 and is the waveform which is presented to the input of the baseband filter. Since these spikes occur only rarely and are impulse-like in character, they represent a *shot* noise phenomenon. The noise power at the baseband filter output is calculated in Sec. 9.7, using the results given in Sec. 2.5.3.

9.6.2 Spike Characteristics

When the noise amplitude $r(t)$ is comparable to the carrier amplitude A, the resultant phasor $R(t)$ may clearly execute all sorts of random, wide-ranging excursions which will cause $\theta(t)$, and hence the frequency $d\theta/dt$, to experience large changes. Why then do we single out for special consideration the excursion which carries the resultant in a complete rotation around the origin? The reason for our special concern may be seen by comparing the noise outputs for the two cases shown in Fig. 9.18*a*. The path of $R(t)$, marked *spike path*, carries the end point of $R(t)$ completely around the origin as in Fig. 9.16*b* and results in a waveform for $\theta(t)$ as shown in Fig. 9.17*a* and in a waveform

Ans - 6(b)

If we use Eq. (14.43), the overall noise figure of the cascade is

$$F = \frac{1}{g_a}\frac{N_o}{N_i} = \frac{1}{g_{a1}g_{a2}}\frac{N_o}{N_i} \tag{14.48}$$

$$= F_1 + \frac{F_2 - 1}{g_{a1}} \tag{14.49}$$

from Eq. (14.47). If the calculation leading to Eq. (14.49) is extended to a cascade of k stages, the result is

$$F = F_1 + \frac{F_2 - 1}{g_{a1}} + \frac{F_3 - 1}{g_{a1}g_{a2}} + \cdots + \frac{F_k - 1}{g_{a1}g_{a2}\cdots g_{a(k-1)}} \tag{14.50}$$

If the two-ports are characterized by equivalent temperatures rather than noise figures, then the equivalent temperature of the cascade, T_e, is related to the equivalent temperatures and available gains of the individual stages by

$$T_e = T_{e1} + \frac{T_{e2}}{g_{a1}} + \frac{T_{e3}}{g_{a1}g_{a2}} + \cdots + \frac{T_{ek}}{g_{a1}g_{a2}\cdots g_{a(k-1)}} \tag{14.51}$$

Equation (14.51) may be established by combining Eq. (14.50) with Eq. (14.49) (see Prob. 14.22).

Suppose that the individual two-ports have comparable noise figures or equivalent temperatures. Then, especially if the gains are large, the contribution to the net output noise of succeeding stages in the cascade becomes progressively smaller. A very effective practice, for the purpose of securing a low-noise receiving system, is to design the first stage of the cascade with a low equivalent temperature and a high gain. A gain of 30 dB is not uncommon. Similarly, equivalent temperatures as low as $T_s = 4°K$ are obtained by cooling the amplifier with liquid nitrogen.

14.5.5 An Example of a Receiving System

The receiver shown in block diagram form in Fig. 14.13 is rather typical of microwave receivers such as are used for *satellite communication*. In such cases, it is certainly justifiable to take considerable pains to keep the noise figure of the receiver as low as possible. For variety, and also to be consistent with practice, we have characterized the noisiness of the first amplifier in terms of a noise temperature, while the other stages have been characterized by a noise figure. We calculate now the overall noise figure of the receiver. The antenna does not enter the calculation, since it is considered the driving source and not part of the receiver. Using Eqs. (14.37) and (14.50), we have

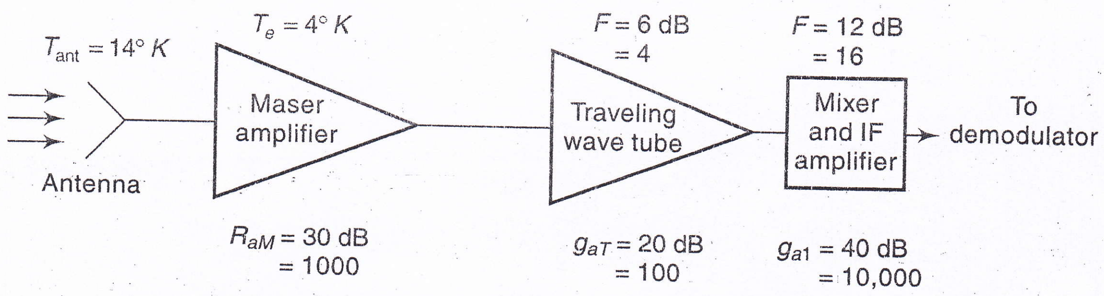


Fig. 14.13 A typical microwave receiver.

Ans- 6(b)

Combining Eqs. (14.39) and (14.40), we have an alternative interpretation of the spot noise figure, that is,

$$F(f) = \frac{G_{ai}^{(n)} / G_{ai}^{(n)}}{G_{ao}^{(s)} / G_{ao}^{(s)}} \tag{14.41}$$

Thus F is a ratio of ratios. The numerator in Eq. (14.41) is the input-signal-to-noise power spectral density ratio, while the denominator is the output-signal-to-noise power spectral density ratio.

Let us assume that in a frequency range from f_1 to f_2 the power spectral densities of signal and noise are uniform. In this case it may be verified (Prob. 14.17) that the average noise figure $\bar{F}$ defined by Eq. (14.38) has the significance

$$\bar{F} = \frac{S_i / N_i}{S_o / N_o} = \frac{S_i}{N_i} \times \frac{N_o}{S_o} \Leftrightarrow \frac{S_i}{N_i} \times \frac{N_o}{g_a S_i} \tag{14.42}$$

where S_i and N_i are, respectively, the total input available signal and noise powers in the frequency range f_1 to f_2, and similarly S_o and N_o are the total output available signal and noise powers.

The noise figure F (or $\bar{F}$) may be expressed in a number of alternative forms which are of interest. If the available gain g_a is constant over the frequency range of interest, so that $F = \bar{F}$, then $S_o = g_a S_i$. In this case Eq. (14.42) may be written as

$$F = \frac{1}{g_a} \frac{N_o}{N_i} \tag{14.43}$$

Further, the output noise N_o is

$$N_o = g_a N_i + N_{tp} \tag{14.44}$$

where $g_a N_i$ is the output noise due to the noise present at the input, and N_{tp} is the additional noise due to the two-port itself. Combining Eqs. (14.43) and (14.44), we have

$$F = 1 + \frac{N_{tp}}{g_a N_i} \tag{14.45}$$

or, the noise due to the two-port itself may be written, from Eq. (14.45), as

$$N_{tp} = g_a (F - 1) N_i \tag{14.46}$$

14.5.4 Noise Figure and Equivalent Noise Temperature of a Cascade

In Fig. 14.12 is shown a cascade of 2 two-ports with a noise source at the input of noise temperature T_0. The individual two-ports have available gains g_{a1} and g_{a2} and noise figures F_1 and F_2. If the input-source noise power is N_i, the output noise due to this source is $g_{a1}\, g_{a2}\, N_i$. The noise output of the first stage due to the noise generated within this first two-port is $g_{a1}(F_1 - 1)N_i$ from Eq. (14.46). The corresponding noise at the output of the second stage is $g_{a1}\, g_{a2}(F_1 - 1)N_i$. Again, using Eq. (14.46), we find that the noise output due to the noise generated within the second two-port is $g_{a2}(F_2 - 1)N_i$. The total noise output is therefore

$$N_o = g_{a1}\, g_{a2}\, N_i + g_{a1}\, g_{a2}(F_1 - 1)N_i + g_{a2}(F_2 - 1)N_i \tag{14.47}$$

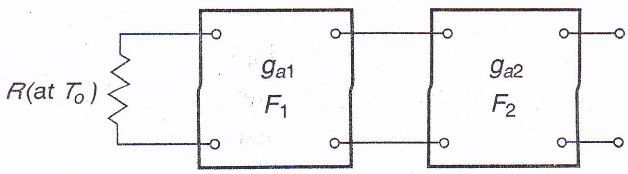


Fig. 14.12 A noise source at temperature T_0 derives a cascade of two-ports.

Ans- 6(c)

these filters on the noise, it is convenient to have a *frequency-domain characterization* of the noise.[1] We shall now establish such a frequency-domain characterization. On the basis of this representation we shall be able to define a *power spectral density* for a noise waveform that has characteristics similar to those of the power spectral density of a deterministic waveform. Our discussion will be somewhat heuristic.

Let us select a particular sample function of the noise, and select from that sample function an interval of duration T extending, say, from $t = -T/2$ to $t = T/2$. Such a noise sample function $n^{(s)}(t)$ is shown in Fig. 7.1a. Let us generate, as in Fig. 7.1b, a periodic waveform in which the waveform in the selected interval is repeated every T sec. This periodic waveform $n_T^{(s)}(t)$ can be expanded in a Fourier series, and such a series will properly represent $n^{(s)}(t)$ in the interval $-T/2$ to $T/2$. The fundamental frequency of the expansion is $\Delta f = 1/T$, and, assuming no dc component, we have

$$n_T^{(s)}(t) = \sum_{k=1}^{\infty} (a_k \cos 2\pi k \,\Delta f t + b_k \sin 2\pi k \,\Delta f t) \tag{7.1}$$

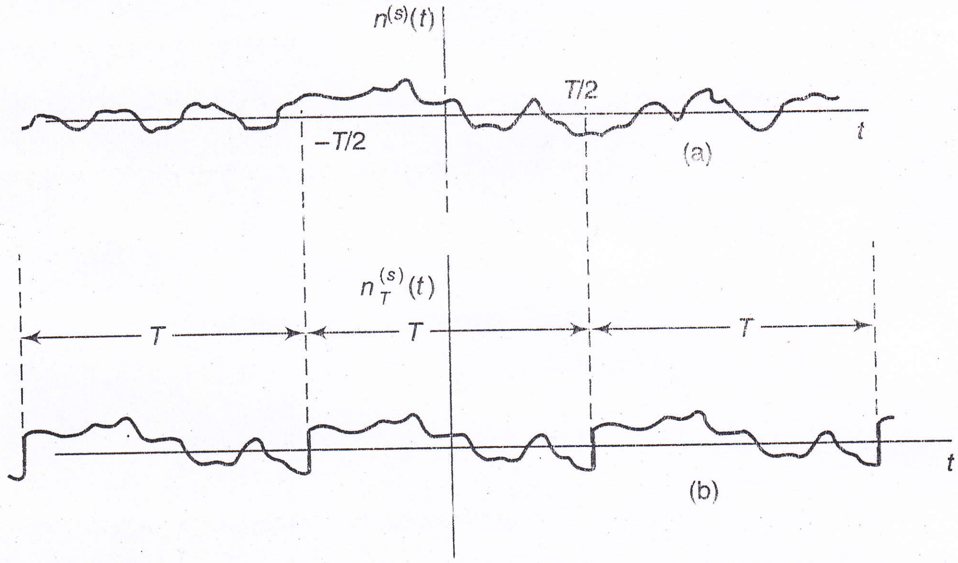


Fig. 7.1 (a) A sample noise waveform. (b) A periodic waveform is generated by repeating the interval in (a) from $-T/2$ to $T/2$.

or alternately

$$n_T^{(s)}(t) = \sum_{k=1}^{\infty} c_k \cos (2\pi k \,\Delta f t + \theta_k) \tag{7.2}$$

in which a_k, b_k and c_k are the constant coefficients of the spectral terms and θ_k is a phase angle. Of course,

$$c_k^2 = a_k^2 + b_k^2 \tag{7.3}$$

and

$$\theta_k = -\tan^{-1} \frac{b_k}{a_k} \tag{7.4}$$

Ans- 6(c)

the volume of the metal. Similarly the molecules of an enclosed gas are in constant motion, colliding with one another and colliding also with the walls of the container. These agitations of molecules are called *thermal* agitations because they increase with temperature.

Let us consider a simple resistor. It is a resistor, or rather a conductor, because there are within it conduction electrons which are free to wander randomly through the entire volume of the resistor. On the average these electrons will be uniformly distributed through the volume, as will positive ions, and the entire structure will be electrically neutral. However, because of the random and erratic wanderings of the electrons, there will be *statistical fluctuations* away from neutrality. Thus at one time or another the distribution of charge may not be uniform, and a voltage difference will appear between the resistor terminals. The random, erratic, unpredictable voltage which so appears is referred to as *thermal resistor noise*. As is to be expected, thermal resistor noise increases with temperature. Resistor noise also increases with the resistance value of the resistor, being zero in a perfect conductor.

A second type of noise results from a phenomenon associated with the flow of current across semiconductor junctions. The charge carriers, electrons or holes, enter the junction region from one side, drift or are accelerated across the junction, and are collected on the other side. The average junction current determines the average interval that elapses between the times when two successive carriers enter the junction. However, the exact interval that elapses is subject to random statistical fluctuations. This randomness gives rise to a type of noise which is referred to as *shot noise*. Shot noise is also encountered as a result of the randomness of emission of electrons from a heated surface and is consequently also associated with thermionic devices.

When a signal reaches a receiver it may well arrive very greatly attenuated. It is therefore necessary to provide amplification. This amplification is accomplished in circuits using active devices (transistors, etc.) and resistors. Hence the signal becomes corrupted by *thermal* and *shot* noise. Even more, the signal may have been contaminated by noise as a result of many types of random disturbances superimposed on the signal during the course of its transfer over the communication channel. The contamination of the signal may take several forms. The noise may be added to the signal, in which case it is called *additive* noise, or the noise may multiply the signal, in which case the effect is called *fading*.

We shall confine our interest, for the most part, albeit not exclusively, to noise which may be described as an ergodic random process. The characteristic of ergodicity of interest here is that an ergodic process is also stationary, that is, statistical averages taken over an ensemble representing the processes yield a result that is independent of the time at which the averages are evaluated. We shall further assume, except where specifically noted, that the probability density of the noise is gaussian. In very many communication systems and in a wide variety of circumstances the assumption of a gaussian density is justifiable. On the other hand, it needs to be noted that such an assumption is hardly universally valid. For example, if gaussian noise is applied to the input of a rectifier circuit, the output is not gaussian. Similarly, it may well be that the noise encountered on a telephone line or on other channels consists of short, pulse-type disturbances whose amplitude distribution is decidedly not gaussian.

7.2 FREQUENCY-DOMAIN REPRESENTATION OF NOISE

In a communication system noise is often passed through filters. These filters are usually described in terms of their characteristics in the frequency domain. Hence, to determine the influence of