NEAR EAST UNIVERSITY

Faculty of Engineering

Department of Electrical and Electronics Engineering

FREQUENCY MODULATION BROADCASTING

Graduation Project
EE 400

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ABSTRACT

Frequency Modulation Broadcasting is of great importance regarding audio transmission and reception by providing good audio performance, as FM is relatively immune to the effects of noise because in Frequency Modulation Broadcasting the message signal is carried in between the carrier signal thus the effect of noise on the carrier signal is removed by using a filter. Whereas in AM noise effects the carrier signal also the message signal as message signal is carried at the top peak amplitude of the carrier signal. Also because generation of FM is a good deal more complex to think about and visualize than those of AM. This is mainly because FM involves minute frequency variations of the carrier, whereas AM results in large-scale amplitude variations of the carrier.

The main objective of this thesis is to provide information about the broadcasting of FM. The factors effecting transmission and reception in the FM broadcasting. A high level mathematical tool, Bessel functions are presented to solve for the frequency components of a FM. The spikes of external noise picked up during transmission are clipped off by a limitter. The FM discriminator extracts the intelligence that has been modulated onto the carrier via frequency variations, for this purpose slope detector is used to receive FM but is not widely used.

INTRODUCTION

The concept of FM was first practically postulated as an alternative to AM in 1931. At that point, commercial AM broadcasting had been in existence for over 10 years, and the superheterodyne receivers were just beginning to supplant the TRF designs. The goal of research into an alternative to AM at that time was to develop a system less susceptible to external noise pickup. Major E. H. Armstrong developed the first working FM system in 1936, and in July 1939. He began the first regularly scheduled FM broadcast in Alpine, New Jersey.

Chapter 1 discusses about the FM system. Here spectrum of an FM signal, Bandwidth of a sinusoidally modulated FM signal , Effect of the modulation index β on bandwidth, Spectrum of constant bandwidth FM, Spectrum of wideband FM and phasor diagram for FM signals are shown and explained.

In Chapter 2 Frequency modulation transmission has been described including the topics of vital importance in FM transmission such as Simple FM Generator and its two major concepts on which FM Generator is based, FM Analysis has also been described with FM mathematical solutions which are shown by the help of Bessel function, the FM Noise Analysis is also being discussed covering also noise suppression which is playing very important role in the FM transmission as FM has superior noise characteristics over AM, capture effect, preemphasis has also been mentioned. Direct FM generation using varactor diode, reactance modulator and the VCO have also been discussed also with the indirect FM generation and wideband deviation. Phase-Locked-Loop FM transmitter has been discussed in details describing its block diagram, and transmitter schematic and its operation on other bands. Stereo FM and its generation are also covered. The final topic discussed is the FM transmissions describing five major categories in which FM is used and some Fm advantages over SSB and AM.

Chapter 3 relates to the FM reception, including the types of elements used for frequency modulation reception, RF amplifiers, FET RF amplifiers, MOST FAT RF amplifiers with the circuit diagram have been discussed. Limiters, discriminators have also been covered in detail with their related covering fields, phase locked loop, stereo demodulation, SCA decoder, LIC stereo decoder are explained with their respective block diagrams and circuit.

CHAPTER 1

FREQUENCY MODULATION SYSTEM

1.1 Angle Modulation

Angle Modulation is defined as modulation where the angle of a sine wave carrier is varied from its reference value. Angle Modulation has two subcategories, Phase Modulation and Frequency Modulation:

Phase Modulation (PM): Angle Modulation where the phase angle of a carrier is caused to depart from its reference value by an amount proportional to the modulating signal amplitude.

Frequency Modulation (FM): Angle modulation where the instantaneous frequency of a carrier is caused to vary by an amount proportional to the modulating signal amplitude.

The key difference between these two similar forms of modulation is that in PM the amount of phase change is proportional to intelligence amplitude. While in FM it is the frequency change that is proportional to intelligence amplitude. As it turns out, PM is not directly used as the transmitted signal in communication systems but does have importance since it is often used to help generate FM, and a knowledge of PM helps us to understand the superior noise characteristics of FM as compared to AM systems. In recent years, it has become fairly common practice to denote angle modulation simply as FM instead of specifically referring to FM and PM.

The concept of FM was first practically postulated as an alternative to AM in 1931. At that point, commercial AM broadcasting had been in existence for over 10 years, and the superheterodyne receivers were just beginning to supplant the TRF designs. The goal of research into an alternative to AM at that time was to develop a system less susceptible to external noise pickup. Major E. H. Armstrong developed the first working FM system in 1936, and in July 1939. He began the first regularly scheduled FM broadcast in Alpine, New Jersey.

All the modulation schemes considered up to the present point have two principal features in common. In the first place, each spectral component of the baseband signal gives rise to one or two spectral components in the modulated signal. These components are separated from the carrier by a frequency difference equal to the frequency of the baseband component. Most importantly, the nature of the modulators is such that the spectral components which they produce depend only on the carrier frequency and the baseband frequencies. The amplitudes of the spectral components of the modulator output may depend on the amplitude of the input signals; however, the frequencies of the spectral components do not. In the second place, all the operations performed on the signal (addition, subtraction, and multiplication) are linear operations so that superposition applies. Thus, if a baseband signal $m_1(t)$ introduces one spectrum of components into the modulated signal and a second signal $m_2(t)$ introduces a second spectrum, the application of the sum $m_1(t) + m_2\{t)$ will introduce a spectrum which is the sum of the spectra separately introduced. All these systems are referred to under the designation " amplitude or linear modulation". This terminology must be taken with some reservation, for we have noted that, at least in the special case of single sideband using modulation with a single sinusoid, there is no amplitude variation at all. And even more generally, when the amplitude of the modulated signal does vary, the carrier envelope need not have the waveform of the baseband signal.

We now turn our attention to a new type of modulation, which is not characterized by the features referred to above. The spectral components in the modulated waveform depend on the amplitude as well as the frequency of the spectral components in the baseband signal. Furthermore, the modulation system is not linear and superposition does not apply. Such a system results when, in connection with a carrier of constant amplitude, the phase angle is made to respond in some way to a baseband signal. Such a signal has the form

$$v(t) = A \cos [\omega_c t + \phi(t)] \tag{1.1}$$

in which A and ω_c are constant but in which the phase angle $\phi(t)$ is a function of the baseband signal. Modulation of this type is called angle modulation for obvious reasons. It is also referred to as phase modulation since $\phi(t)$ is the phase angle of the argument of the cosine function.

1.2 Phase And Frequency Modulation

To review some elementary ideas in connection with sinusoidal waveforms, let us recall that the function A cos $\omega_c t$ can be written as

$$A \cos \omega_c t = \text{real part } (Ae^{j\omega ct}) \tag{1.2}$$

The function $Ae^{j\theta}$ is represented in the complex plane by a phasor of length A and an angle θ measured counterclockwise from the real axis. If $\theta = \omega_c$ then the phasor rotates in the counterclockwise direction with an angular velocity ω_c . With respect to a coordinate system which also rotates in the counterclockwise direction with angular velocity ω_c , the phasor will be stationary. If in Eq. (1.1) ϕ is actually not time-dependent but is a constant, then $v(t)$ is to be represented precisely in the manner just described. But suppose $\phi = \phi(t)$ does change with time and makes positive and negative excursions. Then $v(t)$ would be represented by a phasor of amplitude A which runs ahead of and falls behind the phasor representing A cos $\omega_c t$. We may, therefore, consider that the angle $\omega_c t + \phi(t)$, of $v(t)$, undergoes a modulation around the angle $\theta = \omega_c t$, The waveform of $v(t)$ is, therefore, a representation of a signal which is modulated in phase.

If the phasor of angle $\theta + \phi(t) = \omega_c t + \phi(t)$ alternately runs ahead of and falls behind the phasor $\theta = \omega_c t$, then the first phasor must alternately be rotating more, or less, rapidly than the second phasor. Therefore we may consider that the angular velocity of the phasor of $v(t)$ undergoes a modulation around the nominal angular velocity ω_c. The signal

$v(t)$ is, therefore, an angular-velocity-modulated waveform. The angular velocity associated with the argument of a sinusoidal function is equal to the time rate of change of the argument (i.e., the angle) of the function. Thus we have that the instantaneous radial frequency $\omega = d(\theta + \phi)/dt$, and the corresponding frequency $f = \omega/2\pi$ is

$$f = 1/2\pi \; d/dt \, [\omega_c t + \phi(t)] = \omega_c/2\pi + 1/2\pi \; d/dt \;\; \phi(t) \tag{1.3}$$

The waveform $v\{t)$ is, therefore, modulated in frequency. In initial discussions of the sinusoidal waveform it is customary to consider such a waveform as having a fixed frequency and phase. In the present discussion we have generalized these concepts somewhat. To acknowledge this generalization, it is not uncommon to refer to the frequency f in Eq. (1.3) as the instantaneous frequency and $\phi(t)$ as the instantaneous phase. If the frequency variation about the nominal frequency ω_c is small, that is, if $d\phi(t)/dt << \omega_c$, then the resultant waveform will have an appearance which is readily recognizable as a "sine wave," albeit with a period which changes somewhat from cycle to cycle. Such a waveform is represented in Fig. 1.1 . In this figure the modulating signal is a square wave. The frequency-modulated signal changes frequency whenever the modulation changes level.

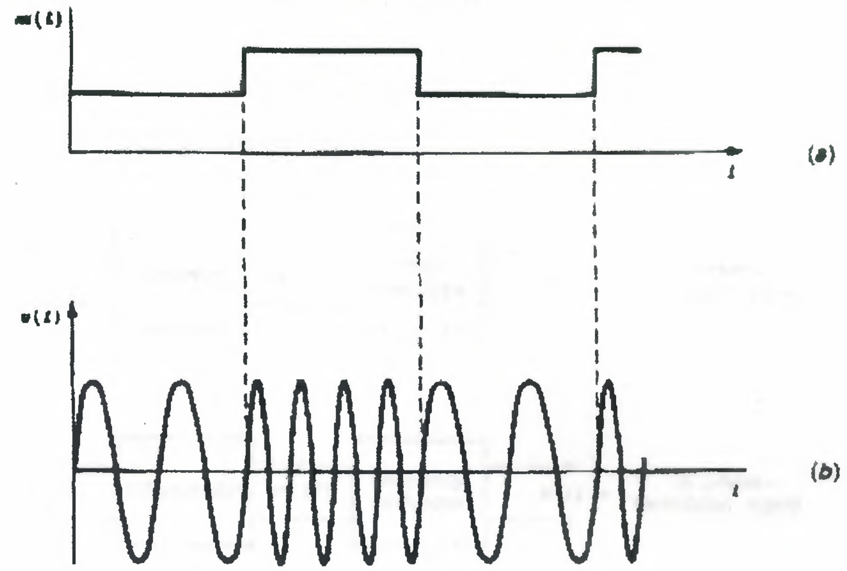

Figure 1.1 An angle-modulated waveform. (a) Modulating signal (b) Frequently-modulated sinusoidal carrier signal.

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. (1.1) 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. (1.3), with $f_c = \omega_c / 2\pi$

$$d\phi(t)/dt = 2\pi(f - f_c) \tag{1.4}$$

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.

1.3 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. 1.2.

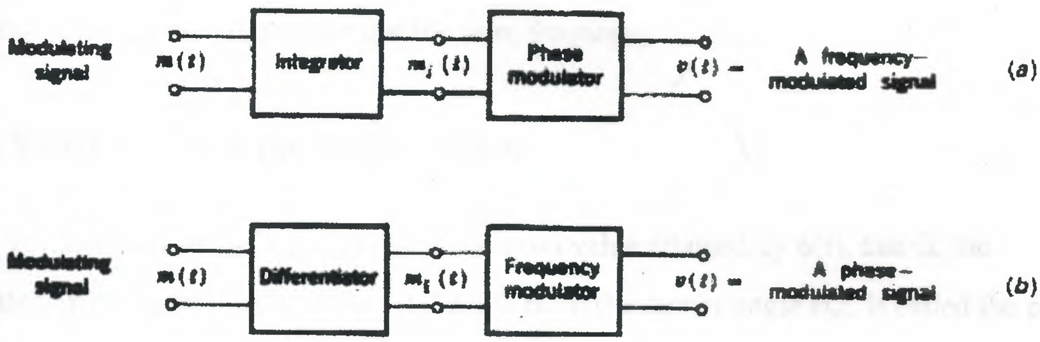

Figure 1.2 Illustrating the relationship between phase and frequency modulation.

In Fig. 1.2a 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

$$v(t) = A \cos [\omega_c t + k m_i(t)] \tag{1.5}$$

k being a constant.

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

$$v \equiv f - f_c = k/2\pi \; m(t) \tag{1.6}$$

Since the deviation of the instantaneous frequency is directly proportional to the modulating signal, the combination of integrator and phase modulator of Fig.1.2a constitutes a device for producing a frequency-modulated output. Similarly, the combination in Fig. 1.2b of the differentiator and frequency modulator generates a phase-modulated output, i.e., and 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. (1.1) 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.

1.4 Phase And Frequency Deviation

In the waveform of Eq. (1.1) 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.

When the angular (and consequently the frequency) variation is sinusoidal with frequency f_m, we have, with $\omega_m = 2\pi f_m$

$$v(t) = A \cos (\omega_c t + \beta \sin \omega_m t) \tag{1.7}$$

Where β is the peak amplitude of $\phi(t)$. In this case β which is the maximum phase deviation, is usually referred to as the modulation index. The instantaneous frequency is

$$f = \omega_c / 2\pi + \beta\omega_m / 2\pi \cos \omega_m t \tag{1.8a}$$

$$= f_c + \beta f_m \cos \omega_m t \tag{1.8b}$$

The maximum frequency deviation is defined as Δf and is given by

$$\Delta f = \beta f_m \tag{1.9}$$

Equation (4.4-1) can, therefore, be written

$$v(t) = A \cos(\omega_c t + \Delta f/f_m \sin \omega_m t) \tag{1.10}$$

While the instantaneous frequency f lies in the range $f_c \pm \Delta f$, it should not be concluded that all spectral components of such a signal lie in this range. We consider next the spectral pattern of such an angle-modulated waveform.

1.5 Spectrum Of An FM Signal Sinusoidal Modulation

In this section we shall look into the frequency spectrum of the signal

$$v(t) = \cos(\omega_c t + \beta \sin \omega_m t) \tag{1.11}$$

which is the signal of Eq. (1.7) with the amplitude arbitrarily set at unity as a matter of convenience. We have

$$\cos(\omega_c t + \beta \sin \omega_m t) = \cos \omega_c t \cos(\beta \sin \omega_m t) - \sin \omega_c t \sin(\beta \sin \omega_m t) \tag{1.12}$$

Consider now the expression $\cos(\beta \sin \omega_m t)$ which appears as a factor on the right-hand side of Eq. (1.12). It is an even, periodic function having an angular frequency ω_m. Therefore it is possible to expand this expression in a Fourier series in which $\omega_m/2\pi$ is the fundamental frequency. We shall not undertake the evaluation of the coefficients in the Fourier expansion of $\cos(\beta \sin \omega_m t)$ but shall instead simply write out the results. The coefficients are, of course, functions of β, and, since the function is even, the coefficients of the odd harmonics are zero. The result is

$$\begin{aligned}\cos(\beta \sin \omega_m t) = {} & J_0(\beta) + 2J_2(\beta) \cos 2\omega_m t + 2J_4(\beta) \cos 4\omega_m t \\ & + \cdots + 2J_{2n}(\beta) \cos 2n\omega_m t + \cdots \end{aligned} \tag{1.13}$$

while for $\sin(\beta \sin \omega_m t)$, which is an odd function, we find the expansion contains only odd harmonics and is given by

$$\sin(\beta \sin \omega_m t) = 2J_1(\beta) \sin\omega_m t + 2J_3(\beta) \sin 3\omega_m t + \cdots + 2J_{2n-1}(\beta) \sin(2n-1)\omega_m t + \cdots \quad (1.14)$$

The functions $J_n(\beta)$ occur often in the solution of engineering problems. They are known as Bessel functions of the first kind and of order n.

Putting the results given in Eqs. (1.13) and (1.14) back into Eq. (1.12) and using the identities

$$\cos A \cos B = \tfrac{1}{2} \cos(A - B) + \tfrac{1}{2} \cos(A + B) \quad (1.15)$$

$$\sin A \sin B = \tfrac{1}{2} \cos(A - B) - \tfrac{1}{2} \cos(A+B) \quad (1.16)$$

We find that v(t) in Eq. (1.11) becomes

$$v(t) = J_0(\beta) \cos \omega_c t - J_1(\beta)[\cos(\omega_c - \omega_m)t - \cos(\omega_c + \omega_m)t] + J_2(\beta)[\cos(\omega_c - 2\omega_m)t + \cos(\omega_c + 2\omega_m)t] - J_3(\beta)[\cos(\omega_c - 3\omega_m)t - \cos(\omega_c + 3\omega_m)t] + \cdots \quad (1.17)$$

Observe that the spectrum is composed of a carrier with an amplitude $J_0(\beta)$ and a set of sidebands spaced symmetrically on either side of the carrier at frequency separations of ω_m ,$2\omega_m$, $3\omega_m$, etc. In this respected the result is unlike that which prevails in the amplitude-modulation systems discussed earlier, since in AM a sinusoidal modulating signal gives rise to only one sideband or one pair of sidebands. A second difference is that the present modulation system is nonlinear.

1.6 Bandwidth Of A Sinusoidally Modulated FM Signal

In principle, when a FM signal is modulated, the number of sidebands is infinite and the bandwidth required to encompass such a signal is similarly infinite in extent. As a matter of practice, it turns out that for any β, so large a fraction of the total power is confined to the sidebands which lie within some finite bandwidth that no serious distortion of the signal results if the sidebands outside this bandwidth are lost.

It is found experimentally that the distortion resulting from bandlimiting a FM signal is tolerable as long as 98 percent or more of the power is passed by the bandlimiting filter. This definition of the bandwidth of a filter is, admittedly, somewhat vague, especially since the term " tolerable " means different things in different applications. However, using this definition for bandwidth, one can proceed with an initial tentative design of a system. When the system is built, the bandwidth may thereafter be readjusted, if necessary. For sinusoidal modulation the bandwidth required to transmit or receive the FM signal is

$$B = 2(\beta + 1)f_m \tag{1.18}$$

It is verified on a numerical basis that the rule given in Eq. (1.18) holds up to $\beta = 29$, which is the largest value of β .

Using Eq. (1.9), we may put Eq. (1.18) in a form, which is more immediately significant. We find

$$B = 2\,(\Delta f + f_m) \tag{1.19}$$

Expressed in words, the bandwidth is twice the sum of the maximum frequency deviation and the modulating frequency. This rule for bandwidth is called Carson's rule. The spectra of several FM signals with sinusoidal modulation are shown in Fig. 1.3 for various values of β . The spectral lines have, in every case, been drawn upward even when the corresponding entry is negative. Hence, the lines represent the magnitudes only of the spectral components. Not all spectral components have been drawn. Those far removed from the carrier, which are too small to be drawn conveniently to scale, have been omitted.

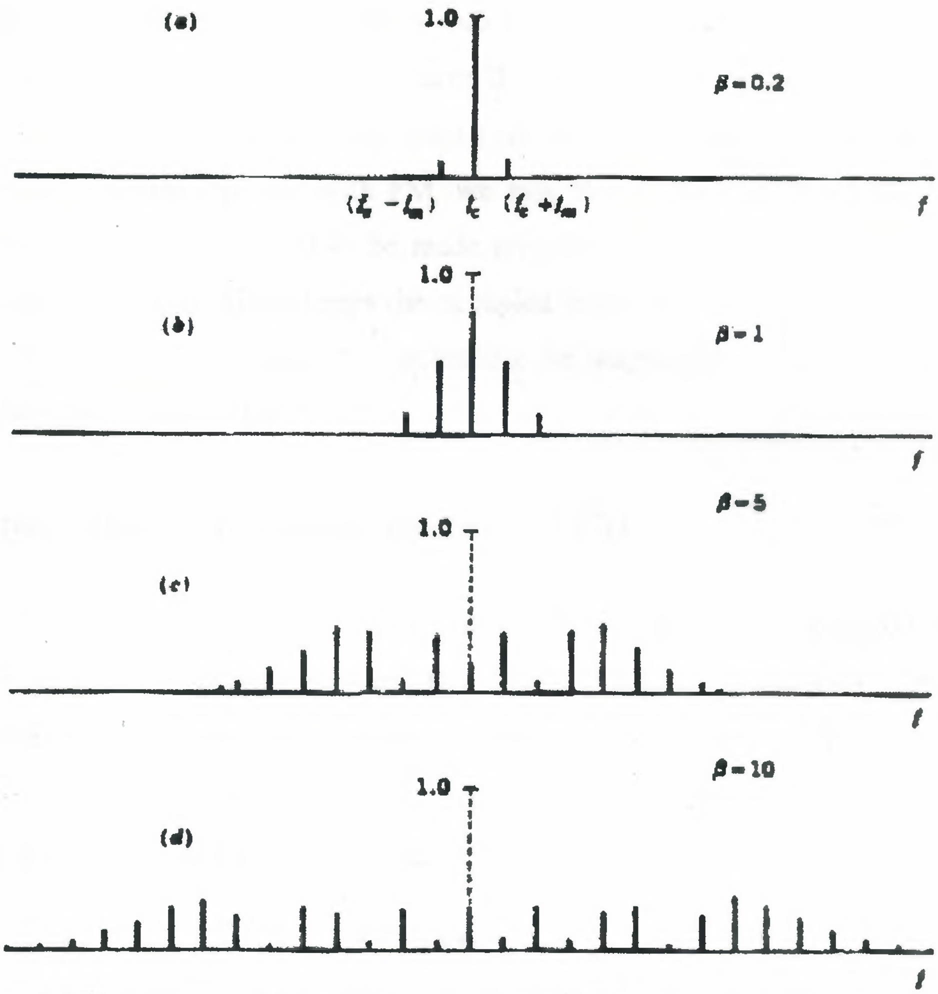

Figure 1.3 The spectra of sinusoidally modulated FM signals for various values of β.

1.7 Effect Of The Modulation Index β On Bandwidth

The modulation index β plays a role in FM, which is not unlike the role played by the parameter m in connection with AM. In the AM case, and for sinusoidal modulation, we established that to avoid distortion we must observe m = 1 as an upper limit. It was also apparent that when it is feasible to do so, it is advantageous to adjust m to be close to unity, that is, 100 percent modulation; by so doing, we keep the magnitude of the recovered baseband signal at a maximum. On this same basis we expect the advantage to lie with keeping ft as large as possible. For, again, the larger is β , the stronger will be the recovered signal. While in AM the constraint that m ≤ 1 is imposed by the necessity to avoid distortion, there is no similar absolute constraint on β.

There is, however, a constraint, which needs to be imposed on β for a different reason. From Eq. (1.18) for β>>1 we have $B \cong 2\beta f_m$ Therefore the maximum value we may allow for ft is determined by the maximum allowable bandwidth and the modulation frequency. In comparing AM with FM, we may then note, in review, that in AM the recovered modulating signal may be made progressively larger subject to the onset of distortion in a manner which keeps the occupied bandwidth constant. In FM there is no similar limit on the modulation, but increasing the magnitude of the recovered signal is achieved at the expense of bandwidth.

1.8 Spectrum Of "Constant Bandwidth" FM

The Fig. 1.4 shows the spectrum for three values of β for the condition that βf_m is kept constant. The nominal bandwidth $B \cong 2\,\Delta f = 2\,\beta f_m$, is consequently constant. The amplitude of the unmodulated carrier at f_c is shown by a dashed line. Note that the extent to which the actual bandwidth extends beyond the nominal bandwidth is greatest for small β and large f_m and is least for large β and small f_m.

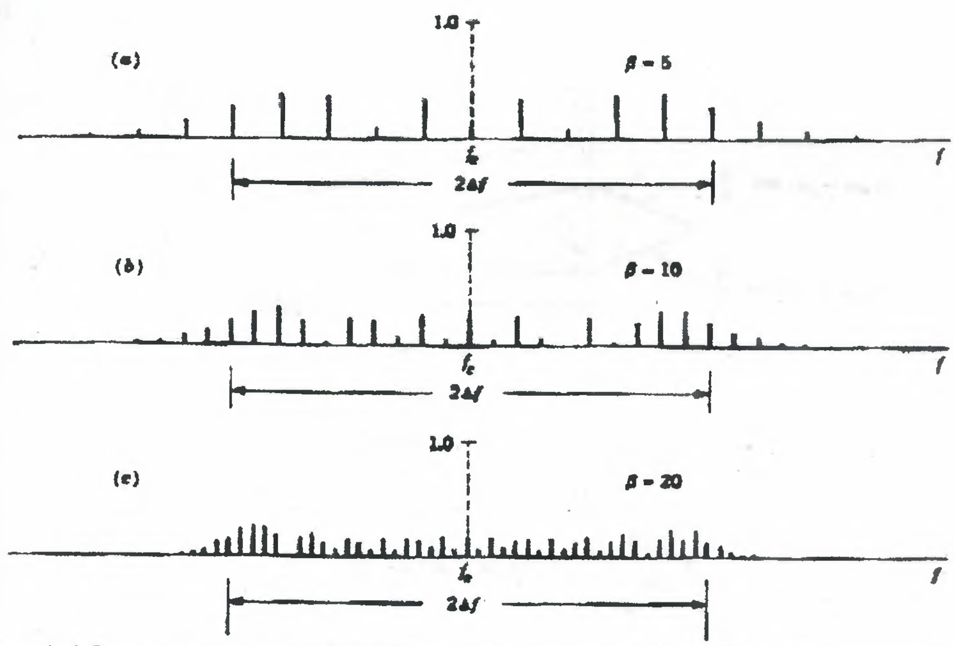

Figure 1.4 Spectra of sinusoidally modulated FM signals. The nominal bandwidth β ≈ $2\beta f_m = 2\Delta f$ is kept fixed.

In commercial FM broadcasting, the Federal Communications Commission allows a frequency deviation Δf= 75 kHz . If we assume that the highest audio frequency to be transmitted is 15 kHz , then at this frequency $\beta = \Delta f/f_m = 75/15 = 5$. For all other modulation frequencies β is larger than 5. When $\beta = 5$, there are $\beta + 1 = 6$ significant sideband pairs so that at f_m = 15 kHz the bandwidth required is B = 2×6×15 =180 kHz, which is to be compared with 2 Δf = 150 kHz. When β =20, there are 21 significant sideband pairs, and B = 2 × 21 × 15 / 4 = 157.5 kHz. In the limiting case of very large β and correspondingly very small f_m, the actual bandwidth becomes equal to the nominal bandwidth 2 Δf.

1.9 Phasor Diagram For FM Signals

With the aid of a phasor diagram we shall be able to arrive at a rather physically intuitive understanding of how so odd an assortment of sidebands as in Eq. (1.17) yields an FM signal of constant amplitude. The diagram will also make clear the difference between AM and narrowband FM (NBFM). In both of these cases there is only a single pair of sideband components.

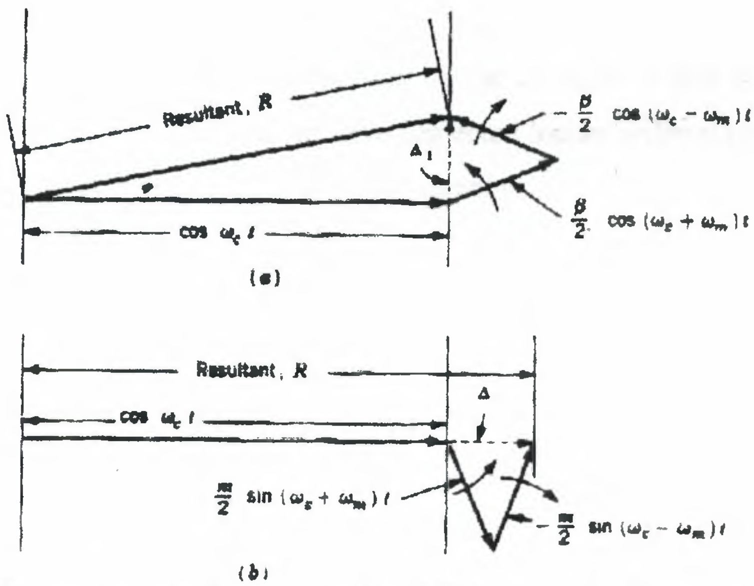

Figure 1.5 (a) Phasor diagram for a narrowband FM signal. (b) Phasor diagram for an AM signal.

Refer to Fig. 1.5(a) Assuming a coordinate system which rotates counter-clockwise at an angular velocity ω_c. In the same coordinate system, the phasor for the term $(\beta/2)\cos(\omega_c + \omega_m)t$ rotates in a counter clockwise direction at an angular velocity ω_m ,while the phasor for the term $-(\beta/2)\cos(\omega_c - \omega_m)t$ rotates in a clockwise direction, also at the angular velocity ω_m. At the time t =0, both phasors, which represent the sideband components, have maximum projections in the horizontal direction. At this time one is parallel to, and one is antiparallel to, the phasor representing the carrier, so that the two cancel. The situation depicted in Fig. 1.5(a) corresponds to a time shortly after t = 0. At this time, the rotation of the sideband phasors which are in opposite directions, as indicated by the curved arrows, have given rise to a sum phasor Δ_1. In the coordinate system in which the carrier phasor is stationary, the phasor Δ_1 always stands perpendicularly to the carrier phasor and has the magnitude.

$$\Delta_1 = \beta \sin \omega_m t \tag{1.20}$$

The carrier now slightly reduced in amplitude, and Δ_1 combine to give rise to a resultant R. The angular departure of R from the carrier phasor is ϕ. It is readily seen from Fig. 1.5(a) that since $\beta \ll 1$, the maximum value of $\phi \cong \tan\phi = \beta$, as is to be expected. The small variation in the amplitude of the resultant which appears in Fig. 1.5(a) is only the result of the fact that we have neglected higher order sidebands.Now let us consider the phasor diagram for AM. The AM signal is

$$(1+ m \sin \omega_m t) \cos \omega_c t = \cos\omega_c t + m/2 \sin (\omega_c+\omega_m)t - m/2 \sin (\omega_c-\omega_m)t \tag{1.21}$$

And the individual terms are represented as phasors in Fig. 1.5(b). We see that there is a 90° phase shift in the phases of the sidebands between the FM and AM cases. In Fig. 4.10-1b the sum Δ of the sideband phasors is given by

$$\Delta = m \sin \omega_m t \tag{1.22}$$

The important difference between the FM and AM cases is that in the former the sum Δ_1 is always perpendicular to the carrier phasor, while in the latter the sum Δ is always parallel to the carrier phasor. Hence in the AM case, the resultant R does not rotate with respect to the carrier phasor but instead varies in amplitude between $1 + m$ and $1 - m$.

To return now to the FM case and to Fig. 1.5(a) the following point is worth noting. When the angle ϕ completes a full cycle, that is, Δ_1 varies from $+\beta$ to $-\beta$ and back again to $+\beta$, the magnitude of the resultant R will have executed two full cycles. For R is a maximum at $\Delta_1 = \beta$, a minimum at $\Delta_1 = 0$, a maximum again when $\Delta_1 = -\beta$, and so on. On this basis, it may well be expected that if an additional sideband pair is to be added to the first to make R more nearly constant, this new pair must give rise to a resultant Δ_2 which varies at the frequency $2\omega_m$. Thus, we are not surprised to find that as the phase deviation β increases, a sideband pair comes into existence at the frequencies $\omega_c \pm 2\omega_m$.

As long as we depend on the first-order sideband pair only, we see from Fig. 1.5(a) that ϕ cannot exceed 90°. A deviation of such magnitude is hardly adequate. For consider, as above, that $\Delta f = 75$ kHz and that $f_m = 50$ Hz. Then $\omega_m = 75{,}000/50 = 1500$ rad, and the resultant R must, in this case, spin completely about $1500/2\pi$ or about 240 times. Such wild whirling is made possible through the effect of the higher-order sidebands. As noted, the first-order sideband pair gives rise to a phasor $\Delta_1 = J_1(\beta) \sin \omega_m t$, which phasor is perpendicular to the carrier phasor. It may also be established by inspection of Eq. (1.17) that the second-order sideband pair gives rise to a phasor $\Delta_2 = J_2(\beta) \cos \omega_m t$ and that this phasor is parallel to the carrier phasor. Continuing, we easily establish that all odd numbered sideband pairs give rise to phasors.

$$\Delta_n = J_n(\beta) \sin n\omega_m t \qquad n \text{ odd} \tag{1.23}$$

which are perpendicular to the carrier phasor, while all even-numbered sideband pairs give rise to phasors.

$$\Delta_n = J_n(\beta) \cos n\omega_m t \qquad n \text{ even} \tag{1.24}$$

Which are parallel to the carrier phasor. Thus, phasors $\Delta_1, \Delta_2, \Delta_3$ etc., alternately perpendicular and parallel to the carrier phasor, are added to carry the end point of the resultant phasor R completely around as many times as may be required, while maintaining R at constant magnitude.

1.10 Spectrum Of WideBand FM (WBFM) Arbitrary Modulation

In this section we engage in a heuristic discussion of the spectrum of a wideband FM signal. As a matter of fact, we shall be able to do no more than to deduce a means of expressing approximately the power spectral density of a WBFM signal. But this result is important and useful.

Previously, to characterize a FM signal as being narrowband or wideband, we had used the parameter $\beta \equiv \Delta f / f_m$, where Δf is the frequency deviation and f_m the frequency of the sinusoidal modulating signal. The signal was then NBFM or WBFM depending on whether $\beta << 1$ or $\beta >> 1$. Alternatively we distinguished one from the other on the basis of whether one or very many sidebands were produced by each spectral component of the modulating signal, and on the basis of whether or not superposition applies. We consider now still another alternative.

Let the symbol $\nu \equiv f - f_c$ represent the frequency difference between the instantaneous frequency f and the carrier frequency f_c that is, $\nu(t) = (k/2\pi)m(t)$ see Eq. (1.6). The period corresponding to ν is $T = 1/\nu$. As f varies, so also will ν and T. The frequency ν is the frequency with which the resultant phasor R in Fig.1.5 rotates in the coordinate system in which the carrier phasor is fixed. In WBFM this resultant phasor rotates through many complete revolutions, and its speed of rotation does not change radically from revolution to revolution. Since, the resultant R is constant, then if we were 0to examine the plot as a function of time of the projection of R in, say, the horizontal direction, we would recognize it as a sinusoidal waveform because its frequency would be changing very slowly. No appreciable change in frequency would take place during the course of a cycle. Even a long succession of cycles would give the appearance of being of rather constant frequency. In NBFM, on the other hand, the phasor R simply oscillates about the position of the carrier phasor. Even though, in this case, we may still formally calculate a frequency ν, there is no corresponding time interval during which the phasor makes complete revolutions at approximately a constant rate ν.

Now let us consider that a carrier is wideband FM-modulated by a signal m(t) such as, say, an audio signal. Let the modulation m(t) be characterized by the probability density function f(m). Then the fraction of the time that m(t) spends in the range between m_1 and m_1 + dm is the probability that m(t) lies between m_1, and m_1 + dm, that is, $f(m_1)$ dm. Corresponding to each value of m(t), the value of the frequency deviation $\nu(t) = (k/2\pi)m(t)$. Hence, during the time m(t) is in the range m_1 and m_1 + dm, ν is in the range ν_1 and $\nu_1 + d\nu$. As we have seen in WBFM, the frequency ν changes only relatively slowly. Thus the assignment of a Frequency ν to a waveform during the interval when m(t) has a value corresponding to ν has a physical as well as a purely mathematical significance. On this basis, it is reasonable to say that, of the total power in the FM waveform, the fraction of the power in the frequency range between ν_1 and $\nu_1 + d\nu$ is proportional to the time m(t} spends in the range m_1 to m_1 + dm. With $G(\nu)$, the power spectral density of the FM waveform, we have the result that $G(\nu_1)\, d\nu$ is proportional to $f(m_1)$ dm. Finally, since $d\nu$ is proportional to dm, we have the most important result that $G(\nu)$ is proportional to f\{m). Expressed in words, the power spectral density $G(\nu)$ of a WBFM waveform is determined by, and has the same form as, the density function f(m) of the modulating waveform.

1.11 Bandwidth Required For A Gaussian Modulated WBFM Signal

Earlier, we found that when a carrier was sinusoidally modulated, Carson's rule $B = 2(\Delta f+f_m)$ given in Eq. (1.19) specifies the bandwidth required to transmit enough of the power (98 percent) so that the modulation may be recovered without distortion- We now make a similar calculation for the case where the modulation has a gaussian distribution. The result is extremely important since many physically encountered signals, while not precisely gaussian, are reasonably approximated as gaussian. We note that if m(t} is gaussian, so also is G(f). Therefore the two-sided spectral density G(f) has the form

$$G(f) = A^2 / 4(\sqrt{2\pi})\Delta f_{rms} [e^{-(f-fc)2/2(\Delta frms)2} + e^{-(f+fc)2/2(\Delta frms)2}] \quad (1.25)$$

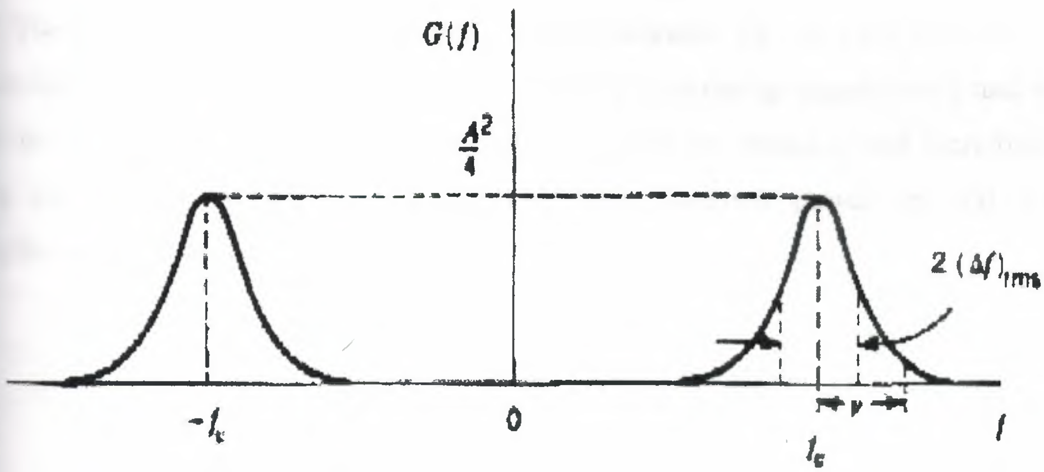

Figure 1.6 The power spectral density of a carrier f_c frequency-modulated by a baseband signal with a gaussian amplitude distribution. The variable ν measures the departure of frequency from the carrier frequency.

As shown in Fig. 1.6 where A is the amplitude of the FM waveform and Δf_{rms} is the variance of the gaussian power spectrum density. An FM waveform of amplitude A has a power $A^2/2$.

We now ask estimate the bandwidth B of a rectangular bandpass filter centered at f_c, which will pass 98 percent of the power of the FM waveform. Recognizing that each of the terms in Eq. (1.25) makes equal contributions to the power and using the variable $v \equiv f \pm f_c$, Bandwidth B is determined as

$$B = 2\,(\sqrt{2})(1.645)\,\Delta f_{rms}$$

$$= 4.6\,\Delta f_{rms} \qquad (1.26)$$

To recapitulate, we find that a modulating signal with a gaussian amplitude distribution gives rise to an FM waveform with a gaussian power spectral density. If the variance of the spectral density is $(\Delta f_{rms})^2$, the bandwidth required to pass 98 percent of the power of the waveform is given by Eq. (1.26).

There is another useful result that may be deduced for the case of a gaussian-modulating signal. Suppose we have two gaussian modulating signals $m_1(t)$ and $m_2(t)$. The probability density functions of these two signals are identical and therefore, will give rise to WBFM waveforms with the same gaussian power spectral density distribution and hence bandwidth B.

CHAPTER 2

FM TRANSMISSION

2.1 A Simple FM Generator

To gain an intuitive understanding of FM. the system illustrated in Fig. 2-1 should be considered. This is actually a very simple, yet highly instructive, FM transmitting system. It consists of an LC tank circuit, which, in conjunction with an oscillator circuit, generates a sine-wave output. The capacitance section of the LC tank is not a standard capacitor but is a capacitor microphone. This popular type of microphone is often referred to as a condenser mike and is, in fact, a variable capacitor. When no sound waves reach its plates, it presents a constant value of capacitance at its two output terminals. However, when sound waves reach the mike, they alternately cause its plates to move in and out.

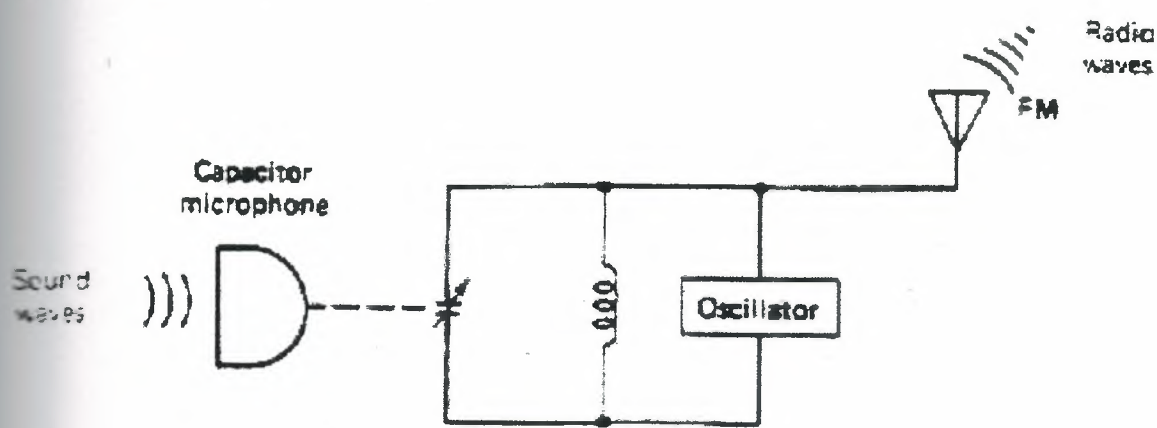

Figure 2.1 Capacitor microphone FM generator.

This causes its capacitance to go up and down around its center value. The rate of this capacitance change is equal to the frequency of the sound waves striking the mike and the amount of capacitance change is proportional to the amplitude of the sound waves.

Since this capacitance value has a direct effect on the oscillator's frequency, the following two important conclusions can be made concerning the system's output frequency:

1. The frequency of impinging sound waves determines the rate of frequency change.
2. The amplitude of impinging sound waves determines the amount of frequency change.

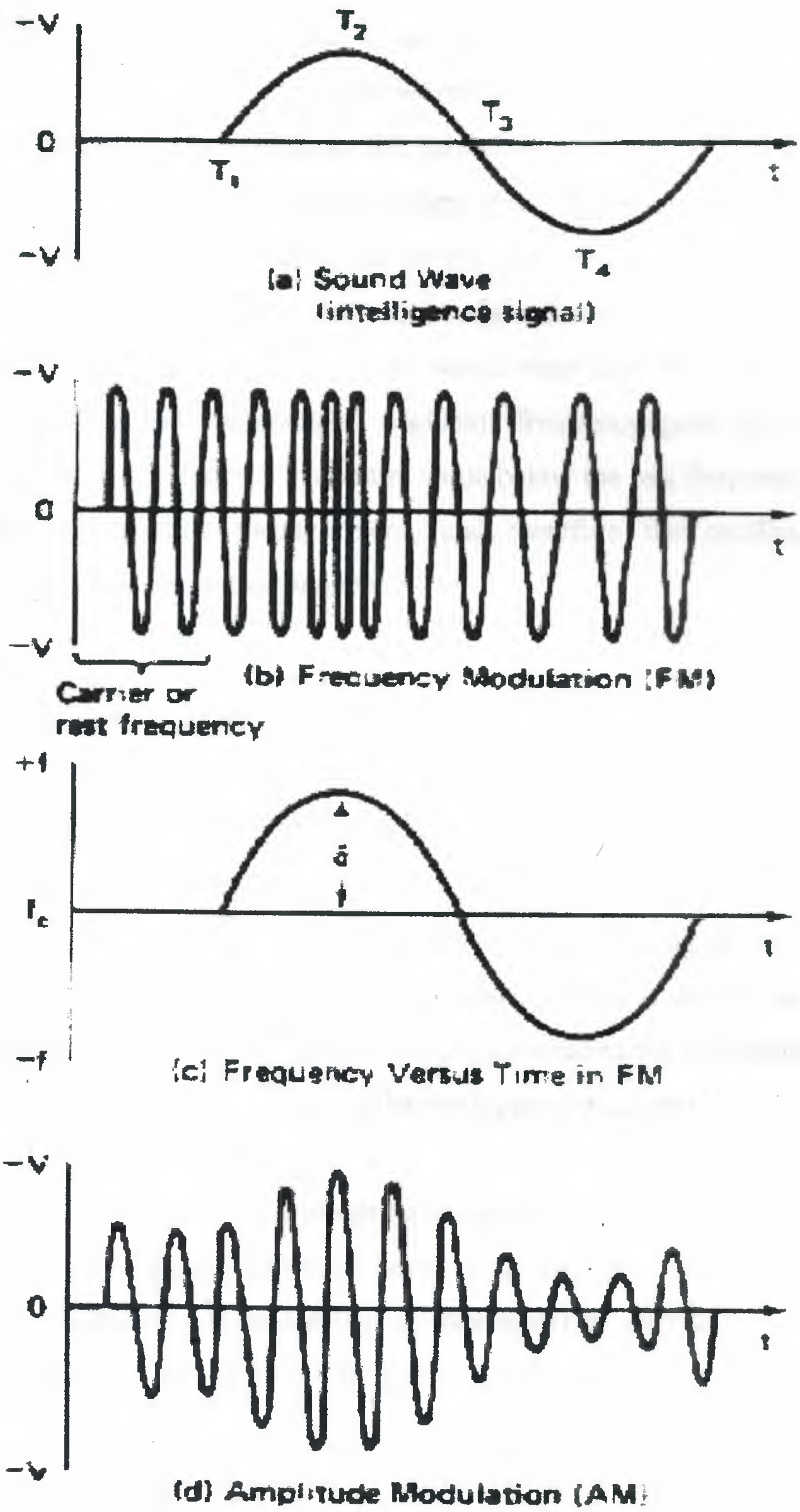

Figure 2.2 FM representation

Consider the case of the sinusoidal sound wave (the intelligence signal) shown in Fig. 2.2(a). Up until time T1 the oscillator's waveform at (b) is a constant frequency with constant amplitude. This corresponds to the carrier frequency (f_c) or rest frequency in FM systems. At T1 the sound wave at (a) starts increasing sinusoidally and reaches a maximum positive value at T2. During this period, the oscillator frequency is gradually increasing and reaches its highest frequency when the sound wave has maximum amplitude at time T2. From time T2 to T4 the sound wave goes from maximum positive to maximum negative and the resulting oscillator frequency goes from a maximum frequency above the rest value to a maximum value below the rest frequency, At time T3 the sound wave is passing through zero, and therefore the oscillator output is instantaneously equal to the carrier frequency.

2.1.1 The Two Major Concepts

The amount of oscillator frequency increase and decrease around f_c is called the frequency deviation, δ. This deviation is shown in Fig 2.2(c) as a function of time. It is ideally shown as a sine-wave replica of the original intelligence signal. It shows that the oscillator output is indeed an FM waveform. Recall that FM is denned as a sine-wave carrier that changes in frequency by an amount proportional to the instantaneous value of the intelligence wave and at a rate equal to the intelligence frequency.

Fig. 2.2(d) shows the AM wave resulting from the intelligence signal shown at (a). This should help you to see the difference between an AM and FM signal. In the case of AM, the carrier's amplitude is varied (by its sidebands) in step with the intelligence, while in FM, the carrier's frequency is varied in step with the intelligence.

The capacitor microphone FM generation system is seldom used in practical applications; its importance is derived from its relative ease of providing an understanding of FM basics. If the sound-wave intelligence striking the microphone were doubled in frequency from 1 kHz to 2 kHz with constant amplitude, the rate at which the FM output swings above and below the center (f_c) frequency would change from 1 kHz to 2 kHz. However, since the intelligence amplitude was not changed, the amount of frequency deviation (δ) above and below f_c will remain the same. On the other hand, if the 1-kHz intelligence frequency were kept the same but its amplitude were doubled, the rote of deviation above and below f_c would remain at 1 kHz, but the amount of frequency deviation would double.

It will often be helpful to review the capacitor mike FM generator:

1. The intelligence amplitude determines the amount of carrier frequency deviation.

2. The intelligence frequency (f_I) determines the rate of carrier frequency deviation.

2.2 FM Analysis

The complete mathematical analysis of angle modulation requires the use of high-level mathematics. For our purposes, it will suffice to simply give the solutions and discuss them. For phase modulation (PM), the equation for the instantaneous voltage is

$$e = A \sin(\omega_c t + m_p \sin\omega_I t) \tag{2.1}$$

where e = instantaneous voltage

A = peak value of original carrier wave

ω_c = carrier angular velocity ($2\pi f_c$)

m_p = maximum phase shift caused by the intelligence signal (radians)

ω_I = modulating (intelligence) signal angular velocity ($2\pi f_I$)

The maximum phase shift caused by the intelligence signal, m_p, is defined as the modulation index for PM.

The following equation provides the equivalent formula for FM:

$$e = A\sin(\omega_c t + m_f \sin\omega_I t) \qquad (2.2)$$

All the terms in Eq. (2.2) are defined as they were for Eq. (2.1), with the exception of the new term, m_f. In fact, the two equations are identical except for that term. It is defined as the modulation index for FM, m_f. It is equal to

$$m_f = \text{FM modulation index} = \delta / f_I \qquad (2.3)$$

where δ = maximum frequency shift caused by the intelligence signal (deviation)

f_I = frequency of the intelligence (modulating) signal

Comparison of Eqs. (2.1) and (2.2) points out the only difference between PM and FM. The equation for PM shows that the phase of the carrier varies with the modulating signal amplitude (since m_p is determined by this),and in FM the carrier phase is determined by the ratio of intelligence signal amplitude (which determines δ) to the intelligence frequency (f_I). Thus, FM is not sensitive to the modulating signal frequency but PM is. The difference between them is subtle—in fact, if the intelligence signal is integrated and then allowed to phase-modulate the carrier, an FM signal is created. This is the method used in the Armstrong indirect FM system . In FM the amount of deviation produced is not dependent on the intelligence frequency as it is for PM. The amount of deviation is proportional to the intelligence signal amplitude for both PM and FM. These conditions are shown in Fig. 2.3.

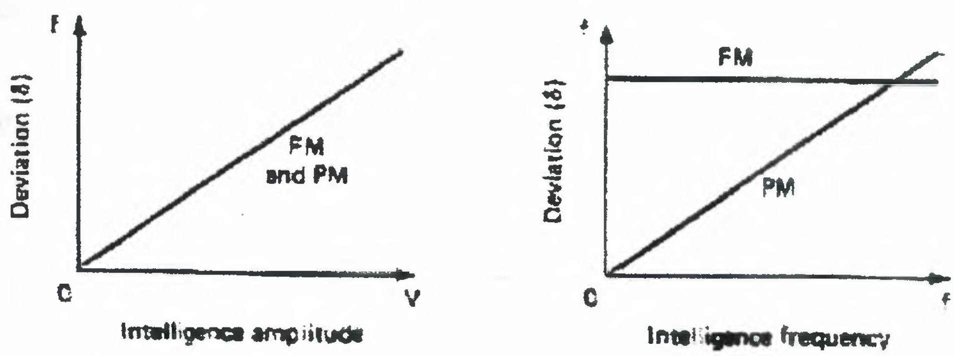

Figure 2.3 Deviation effects on FM/PM by intelligence parameters.

2.2.1 FM Mathematical Solution

The FM formula Eq. (2.2) is really more complex than it looks because it contains the sine of a sine. To solve for the frequency components of an FM wave requires the use of a high-level mathematical tool, Bessel functions. They show that frequency-modulating a carrier with a pure sine wave actually generates an infinite number of sidebands (components) spaced at multiples of the intelligence frequency, f_I, above and below the carrier! Fortunately, the amplitude of these sidebands approaches a negligible level the farther away they are from the carrier, which allows FM transmission within finite bandwidths.

The Bessel function solution to the FM equation is

$$f_c(t) = J_0(m_f)\cos\omega_c t - J_1(m_f)[\cos(\omega_c - \omega_I)t - \cos(\omega_c + \omega_I)t + J_2(m_f)[\cos(\omega_c - 2\omega_I)t + \cos(\omega_c + 2\omega_I)t] + \cdots \quad (2.4)$$

where

$f_c(t) =$ FM frequency components

$J_0(m_f)\cos\omega_c t$ = carrier component

$J_1(m_f)\cos(\omega_c - \omega_I)t + \cos(\omega_c + \omega_I)t$ = component at $\pm f_I$ around the carrier

$J_2(m_f)\cos(\omega_c - 2\omega_I)t + \cos(\omega_c + 2\omega_I)t$ = component at $\pm 2f_I$ around the carrier , etc.

To solve for the amplitude of any side-frequency component, J_n , the following equation should be applied:

$$J_N(m_f) = (m_f / 2)\, n\, [\, 1/n! - (m_f / 2)^2 / 1!(n+1)! + (m_f / 2)^4 / 2!(n+2)! + \cdots] \quad (2.5)$$

Table 2.1 FM side frequencies from Bessel functions

	n OR ORDER																
x (m_f)	(CARRIER) J_0	J_1	J_2	J_3	J_4	J_5	J_6	J_7	J_8	J_9	J_{10}	J_{11}	J_{12}	J_{13}	J_{14}	J_{15}	J_{16}
0.00	1.00	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.25	0.98	0.12	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
0.5	0.94	0.24	0.03	—	—	—	—	—	—	—	—	—	—	—	—	—	—
1.0	0.77	0.44	0.11	0.02	—	—	—	—	—	—	—	—	—	—	—	—	—
1.5	0.51	0.56	0.23	0.06	0.01	—	—	—	—	—	—	—	—	—	—	—	—
2.0	0.22	0.58	0.35	0.13	0.03	—	—	—	—	—	—	—	—	—	—	—	—
2.5	−0.05	0.50	0.45	0.22	0.07	0.02	—	—	—	—	—	—	—	—	—	—	—
3.0	−0.26	0.34	0.49	0.31	0.13	0.04	0.01	—	—	—	—	—	—	—	—	—	—
4.0	−0.40	−0.07	0.36	0.43	0.28	0.13	0.05	0.02	—	—	—	—	—	—	—	—	—
5.0	−0.18	−0.33	0.05	0.36	0.39	0.26	0.13	0.05	0.02	—	—	—	—	—	—	—	—
6.0	0.15	−0.28	−0.24	0.11	0.36	0.36	0.25	0.13	0.06	0.02	—	—	—	—	—	—	—
7.0	0.30	0.00	−0.30	−0.17	0.16	0.35	0.34	0.23	0.13	0.06	0.02	—	—	—	—	—	—
8.0	0.17	0.23	−0.11	−0.29	−0.10	0.19	0.34	0.32	0.22	0.13	0.06	0.03	—	—	—	—	—
9.0	−0.09	0.24	0.14	−0.18	−0.27	−0.06	0.20	0.33	0.30	0.21	0.12	0.06	0.03	0.01	—	—	—
10.0	−0.25	0.04	0.25	0.06	−0.22	−0.23	−0.01	0.22	0.31	0.29	0.20	0.12	0.06	0.03	0.01	—	—
12.0	0.05	−0.22	−0.08	0.20	0.18	−0.07	−0.24	−0.17	0.05	0.23	0.30	0.27	0.20	0.12	0.07	0.03	0.01
15.0	−0.01	0.21	0.04	−0.19	−0.12	0.13	0.21	0.03	−0.17	−0.22	−0.09	0.10	0.24	0.28	0.25	0.18	0.12

Thus, solving for these amplitudes is a very tedious process and strictly dependent on the modulation index, m_f. Table 2.1 gives the solution for a number of modulation indexes. Notice that for no modulation (m_f = 0), the carrier (Jo) is the only frequency present and exists at its full value of 1.However, as the carrier becomes modulated, energy is shifted from the carrier and into the sidebands. For m_f = 0.25, the carrier amplitude has dropped to 0.98 and the first side frequencies at $\pm f_I$ around the carrier (J_1) have an amplitude of 0.12. As indicated previously, FM generates an infinite number of sidebands but in this case, J_2, J_3, J_4, . . . all have negligible value. Thus, an FM transmission with m_f = 0.25 requires the same bandwidth ($2f_I$) as an AM broadcast. Computer programs are also used for Bessel function.

2.2.2 Zero Carrier Amplitude

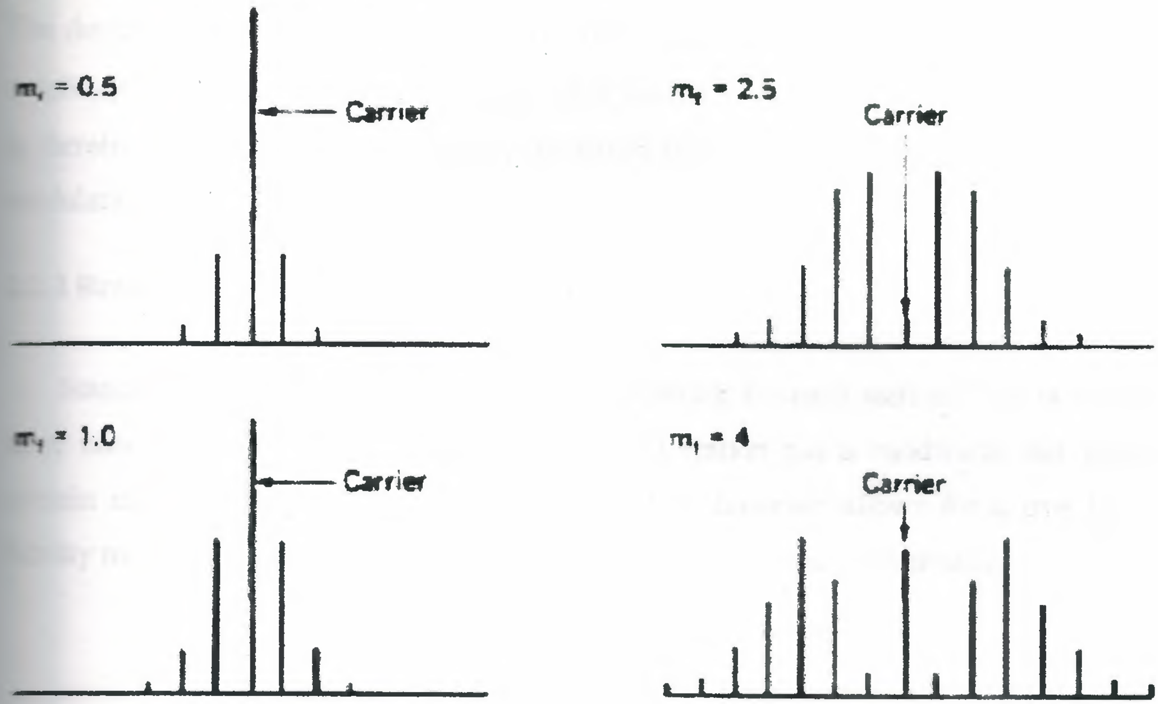

Figure 2.4 Frequency spectrum for FM (constant modulating frequency ,variable deviation).

Figure 2.4 shows the FM frequency spectrum for various levels of modulation while keeping the modulation frequency constant. The relative amplitude of all components is obtained from Table 2.1. Notice from the table that between $m_f = 1$ and $m_f = 2.5$. the carrier goes from a plus to a minus value. The minus sign simply indicates a phase reversal, but when $m_f = 2.4$, the carrier component has zero amplitude and all the energy is contained in the side frequencies. This also occurs when $m_f = 5.5$, 8.65, and between 10 and 12, and 12 and 15.

The zero-carrier condition suggests a convenient means of determining the deviation produced in an FM modulator. A carrier is modulated by a single sine wave at a known frequency. The modulating signal's amplitude is varied while observing the generated FM on a spectrum analyzer. At the point where the carrier amplitude goes to zero. The modulation index m_f is determined based on the number of sidebands displayed. If four or five sidebands appear on both sides of the nulled carrier, you can assume that $m_f = 2.4$. The deviation, δ, is then equal to $2.4 \times f_I$. The modulating signal could be increased in amplitude and the next carrier null should he at $m_f = 5.5$. A check on modulator linearity is thereby possible since the frequency deviation should be directly proportional to the modulating signal's amplitude.

2.2.3 Broadcast FM

Standard broadcast FM uses a 200-kHz bandwidth for each station. This is a very large allocation when one considers that one FM station has a bandwidth that could contain many standard AM stations. Broadcast FM, however allows for a true high-fidelity modulating signal up to 15 kHz and offers superior noise performance.

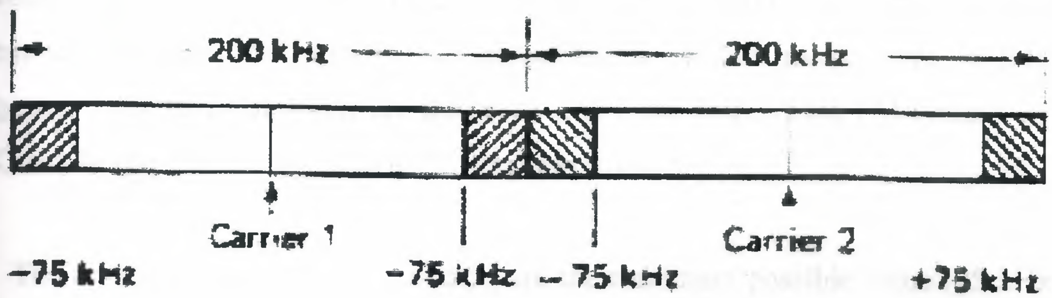

Figure 2.5 Commercial FM bandwidth allocations for two adjacent stations.

Figure 2.5 shows the FCC allocation for commercial FM stations. The maximum allowed deviation around the carrier is ±75 kHz, and 25-kHz guard bands at the upper and lower ends are also provided. The carrier is required to maintain a ±2-kHz stability. Recall that an infinite number of side frequencies are generated during frequency modulation, but their amplitude is gradually decreasing as you move away from the carrier, In other words, the significant side frequencies exist up to ±75 kHz around the carrier, and the guard bands ensure that adjacent channel interference will not be a problem.

Since full deviation (δ) is 75 kHz, that is 100% modulation. By definition, 100% modulation in FM is when the deviation is the full permissible amount. Recall that the modulation index, m_f, is

$$m_f = \delta / f_I \tag{2.3}$$

so that the actual modulation index at 100% modulation varies inversely with the intelligence frequency, f_I. This is in contrast with AM, where full or 100% modulation means a modulation index of 1 regardless of intelligence frequency.

2.2.4 Narrowband FM

Frequency modulation is also widely used in communication (i.e., not to entertain) systems such as police, aircraft, taxicabs, weather service, and private industry networks. These applications are often voice transmissions, which means that intelligence frequency maximums of 3 kHz are the norm. These are narrowband FM systems in that FCC bandwidth allocations of 10 to 30 kHz are provided.

The modulation index when δ and f_I, are the maximum possible value (75 kHz and 15 kHz, respectively, for broadcast FM) is often called the deviation ratio. It has a value of 75 kHz/15 kHz, or 5, for broadcast FM.

In FM, the transmitted waveform never varies in amplitude, just frequency. Therefore, the total transmitted power must remain constant regardless of the level of modulation. It is thus seen that whatever energy is contained in the side frequencies has been obtained from the carrier. No additional energy is added during the modulation process. The carrier in FM is not redundant as in AM, since its (the carrier) amplitude is dependent on the intelligence signal.

2.3 Noise Suppression

The most important advantage of FM over AM is the superior noise characteristics. The static noise is rarely heard on FM, although it is quite common in AM reception. . The addition of noise to a received signal causes a change in its amplitude. Since the amplitude changes in AM contain the intelligence, any attempt to get rid of the noise adversely affects the received signal. However, in FM, the intelligence is not carried by amplitude changes but instead by frequency changes. The spikes of external noise picked up during transmission are "clipped" off by a limiter circuit and/or through the use of detector circuits that are insensitive to amplitude changes.

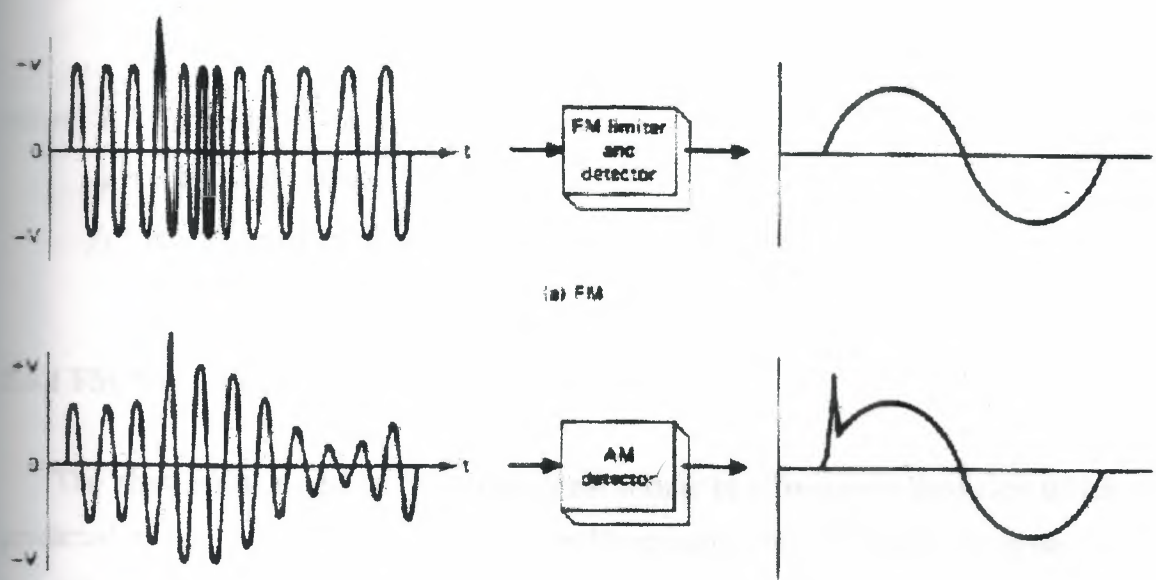

Figure 2.6 FM , AM noise comparison.

Figure 2.6(a) shows the noise removal action of an FM limiter circuit, while at Fig. 2.6(b) the noise spike feeds right through to the speaker in an AM system. The advantage for FM is clearly evident; in fact, you may think that the limiter removes all the effects of this noise spike. Unfortunately, while it is possible to clip the noise spike off, it still causes an undesired phase shift and thus frequency shift of the FM signal, and this frequency shift cannot be removed.

The noise signal frequency will be close to the frequency of the desired FM signal due to the selective effect of the tuned circuits in a receiver. In other words, if you are tuned to an FM station at 96 MHz, the receiver's selectivity provides gain only for frequencies near 96 MHz. The noise that will affect this reception must, therefore, also be around 96 MHz, since all other frequencies will be greatly attenuated. The effect of adding the desired and noise signals will give a resultant signal with a different phase angle than the desired FM signal alone. Therefore, the noise signal, even though it is clipped off in amplitude, will cause phase modulation (PM), which indirectly causes undesired FM. The amount of frequency deviation (FM) caused by PM is

$$\delta = \phi \times f_I \tag{2.7}$$

where δ = frequency deviation

ϕ = phase shift (radians)

f_I = frequency of intelligence signal

2.3.1 FM Noise Analysis

The phase shift caused by the noise signal results in a frequency deviation which is predicted by Eq. (2.7). Consider the situation illustrated in Fig. 2.7. Here the noise signal is one-half the desired signal amplitude, which provides a voltage S/N ratio of 2:1. This is an intolerable situation in AM but as the following analysis will show, not so bad in FM.

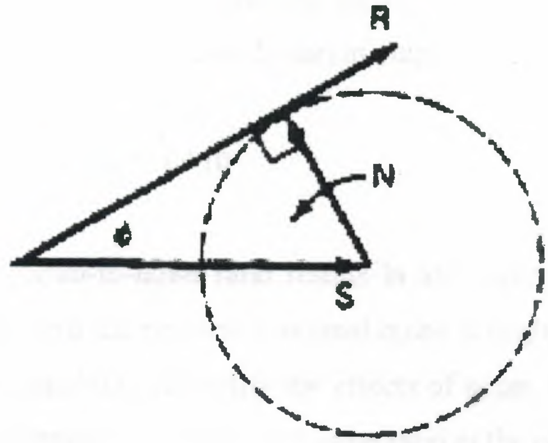

Figure 2.7 Phase shift (ϕ) as a result of noise

Since the noise (N) and desired signal (S) are at different frequencies (but in the same range, as dictated by a receiver's tuned circuits), the noise is shown as a rotating vector using the S signal as a reference. The phase shift of the resultant (R) is maximum when R and N are at right angles to one another.
At this worst-case condition

$$\phi = \sin^{-1} N/S = \sin^{-1} 1/2 = 30^\circ$$

or 30° / (57.3° per radian) = 0.52 rad or about ½ rad .

If the intelligence frequency, f_I, were known, then the deviation (δ) caused by this severe noise condition could now be calculated using Eq. (2.7).Since $\delta = \phi \times f_I$ the worst-case deviation occurs for the maximum intelligence frequency. Assuming f_I maximum of 15 kHz, the absolute worst case δ due to this severe noise signal is

$$\delta = \phi \times f_I = 0.5 \times 15 \text{ kHz} = 7.5 \text{ kHz}$$

In standard broadcast FM, the maximum modulating frequency is 15 kHz and the maximum allowed deviation is 75 kHz above and below the carrier. Thus, a 75-kHz deviation corresponds to maximum modulating signal amplitude and full volume at the receiver's output. The 7.5-kHz worst-case deviation output due to the S/N = 2 condition is

$$7.5\text{kHz} / 75\text{kHz} = 1 / 10$$

and, therefore, the 2:1 signal-to-noise ratio results in an output signal-to-noise ratio of 10:1. This result assumes that the receiver's internal noise is negligible. Thus, FM is seen to exhibit a very strong capability to nullify the effects of noise. In AM, a 2:1 signal-to-noise ratio at the input essentially results in the same ratio at the output. Thus, FM is seen to have an inherent noise reduction capability not possible with AM.

An increase in allowed deviation means that increased bandwidth for each station would be necessary, however. In fact, many FM systems utilized as communication links operate with decreased bandwidths—narrowband FM systems. It is typical for them to operate with a 10-kHz maximum deviation. The inherent noise reduction of these systems is reduced by the lower allowed δ but is somewhat offset by the lower maximum modulating frequency of 3 kHz usually used for voice transmissions.

2.3.2 Capture Effect

This inherent ability of FM to minimize the effect of undesired signals (noise) also applies to the reception of an undesired station operating at the same or nearly the same frequency as the desired station. This is known as the capture effect. You may have noticed when riding in a car that an FM station is suddenly replaced by a different one. You may even find that the receiver alternates abruptly back and forth between the two. This occurs because the two stations are presenting a variable signal as you drive. The capture effect causes the receiver to lock on the stronger signal by suppressing the weaker but can fluctuate back and forth when the two are nearly equal. However, when they are not nearly equal, the inherent FM noise suppression action is very effective in preventing the interference of an unwanted (but weaker) station. The weaker station is suppressed . FM receivers typically have a capture ratio of 1 dB—this means suppression of a 1-dB (or more) weaker station is accomplished. In AM, it is not uncommon to hear two separate broadcasts at the same time, but this is certainly a rare occurrence with FM.

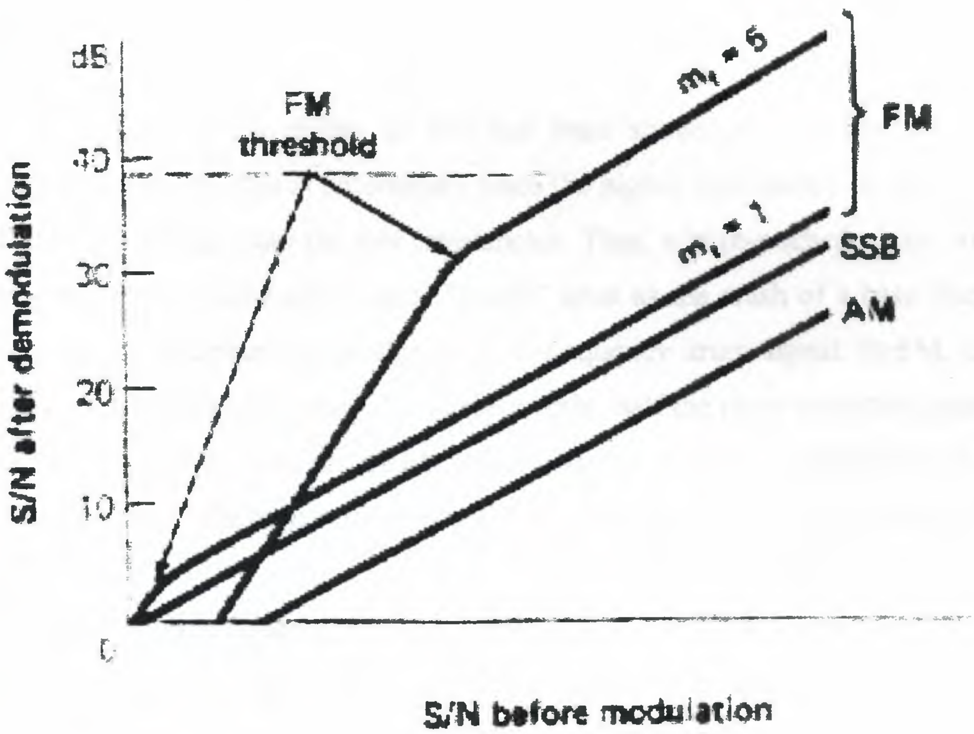

Figure 2.8 S/N for basic modulation schemes.

The capture effect can also be illustrated by referring to Fig. 2.8. Notice that the S/N before and after demodulation for SSB and AM is linear. Assuming noiseless demodulation schemes, SSB (and DSB) has the same S/N at the detector's input and output. The degradation shown for AM is due to so much of the signal's power being wasted in the redundant carrier. FM systems with m_f greater than 1 show an actual improvement in S/N. For example, consider m_f = 5 in Fig. 2.8. When S/N before demodulation is 20, the S/N after demodulation is about 38,a significant improvement.

Insight into the capture effect is provided by consideration of the inflection point (often termed threshold) shown in Fig. 2.8. Notice that a rapid degradation in S/N after demodulation results when the noise is approaching the same level as the desired signal. This threshold situation is noticeable when driving in a large city. The "fluttering" noise often heard is caused when the FM signal is reflecting off various structures. The signal strength fluctuates widely due to the additive or subtractive effects on the total received signal. The effect can cause the output to totally blank out and resume at a rapid rate, as the S/N before demodulation moves back and forth through the threshold level.

2.3.3 Preemphasis

The noise suppression ability of FM has been shown to decrease with higher intelligence frequencies. This is unfortunate since the higher intelligence frequencies tend to be of lower amplitude than the low frequencies. Thus, a high-pitched violin note that the human ear may perceive as the same "sound" level as the crash of a base drum may have only half the electrical amplitudes the low-frequency drum signal. In FM, half the amplitude means half the deviation and, subsequently, half the noise reduction capability. To counter-act this effect, virtually all FM transmissions provide an artificial boost to the electrical amplitude of the higher frequencies. This process is termed preemphasis.

By definition, preemphasis involves increasing the relative strength of the high-frequency components of the audio signal before it is fed to the modulator. Thus. the relationship between the high-frequency intelligence components and the noise is altered. While the noise remains the same, the desired signal strength is increased.

A potential disadvantage, however, is that the natural balance between high- and low-frequency tones at the receiver would be altered. A deemphasis circuit in the receiver, however, corrects this defect, by reducing the high-frequency audio the same amount as the preemphasis circuit increased it, thus regaining the original tonal balance. In addition the deemphasis network operates on both the high-frequency signal and the high-frequency noise; therefore, there is no change in the improved signal-to-noise ratio. The main reason for the preemphasis network, then, is to prevent the high-frequency components of the transmitted intelligence from being degraded by noise that would otherwise have more effect on the higher than on the lower intelligence frequencies.

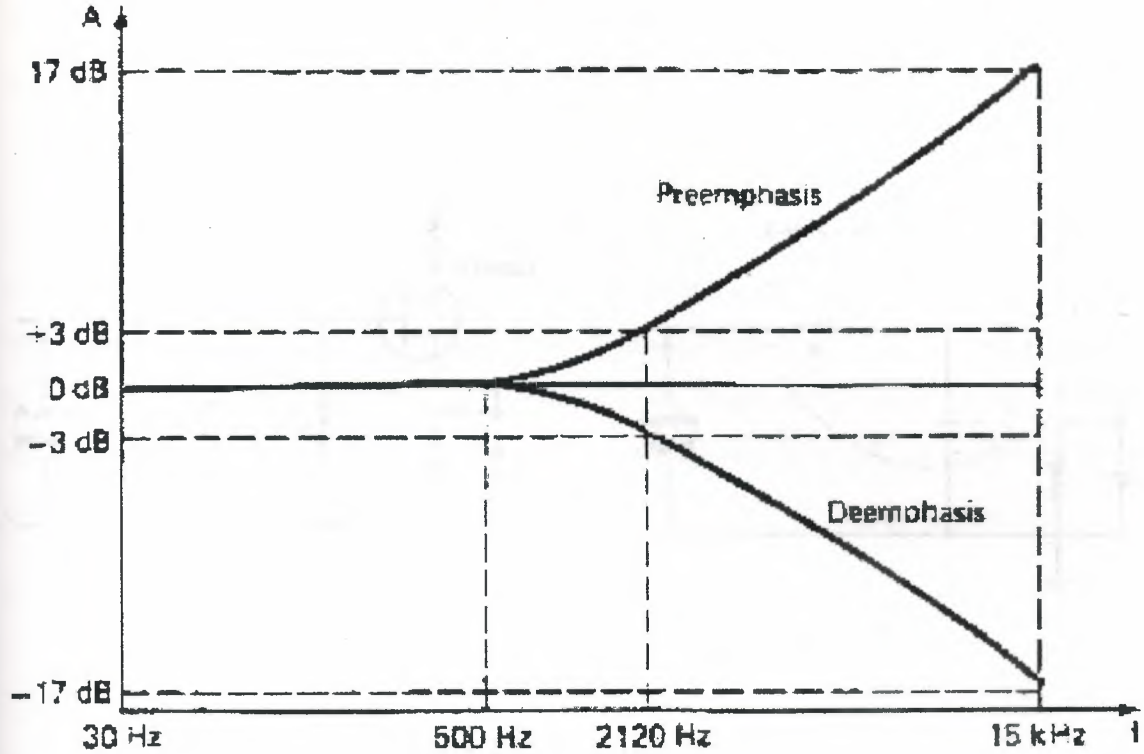

Figure 2.9 Emphasis curves ($\tau = 75\ \mu s$).

The deemphasis network is normally inserted between the detector and the audio amplifier in the receiver. This ensures that the audio frequencies are returned to their original relative level before amplification. The preemphasis characteristic curve is flat up to 500 Hz, as shown in Fig. 2.9. From 500 to15,000 Hz there is a sharp increase in gain up to approximately 17 dB. The gain at these frequencies is necessary to maintain the signal-to-noise ratio at high audio frequencies. The frequency characteristic of the deemphasis network is directly opposite to that of the preemphasis network. The high-frequency response decreases in proportion to its increase in the preemphasis network. The characteristic curve of the deemphasis circuit should be a mirror image of the preemphasis characteristic curve. Figure 2.9 shows the pre- and deemphasis curves as used by standard FM broadcast in the United States. The 3-dB points occur at 2120 Hz as predicted by the RC time constant (τ) of 75 μs used to generate them.

$$f = 1/2\pi RC = 1/2\pi \times 75\mu s = 2120\ Hz$$

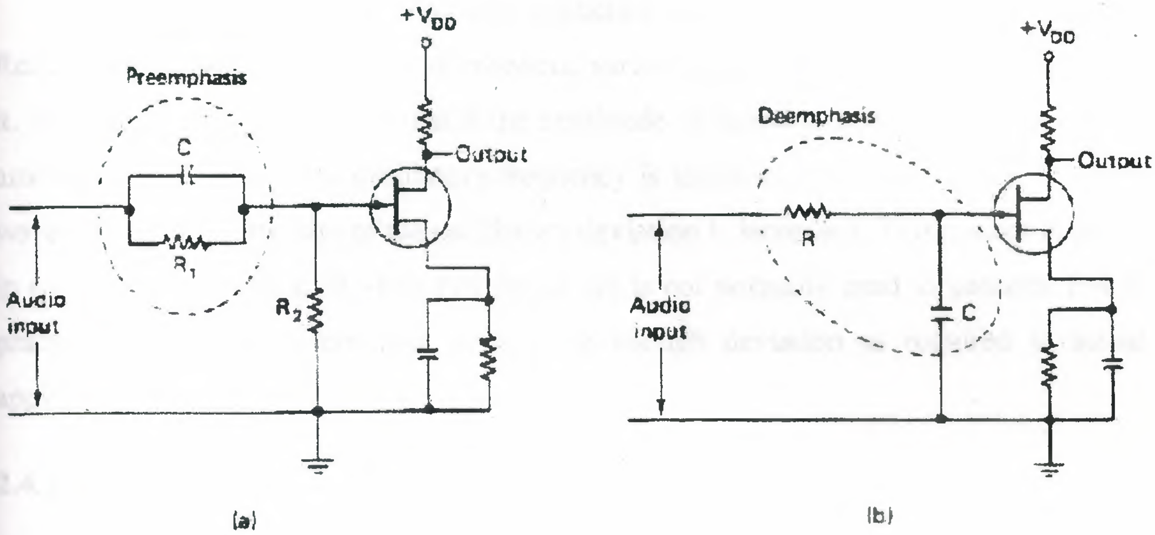

Figure 2.10 Emphasis circuits.

Figure 2.10(a) shows a typical preemphasis circuit. The impedance to the audio voltage is mainly that of the parallel circuit of C and R_1, as the effect of R_2 is small in comparison to that of either C or R_1. Since capacitive reactance is inversely proportional to frequency, audio frequency increases cause the reactance of C to decrease. This decrease of X_c provides an easier path for high frequencies as compared to R. Thus. with an increase of audio frequency, there is an increase in signal voltage. The result is a larger voltage drop across R_2 (the amplifier's input) at the higher frequencies and thus greater output.

Figure 2.10(b) depicts a typical deemphasis network. Note the physical position of R and C in relation to the base of the transistor. As the frequency of the audio signal increases, the reactance of capacitor C decreases. The voltage division between R and C now provides a smaller drop across C. The audio voltage applied to the base decreases; therefore, a reverse of the preemphasis circuit is accomplished. For the signal to be exactly the same as before preemphasis and deemphasis, the time constants of the two circuits must be equal to each other.

2.4 Direct FM Generation

The capacitance microphone system explained can be used to generate FM directly. Recall that the capacitance of the microphone varies in step with the sound wave striking it. Reference to Fig. 2.1 shows that if the amplitude of sound striking it is increased, the amount of deviation of the oscillator's frequency is increased. If the frequency of sound waves is increased, the rate of the oscillator's deviation is increased. This system is useful in explaining the principles of an FM signal but is not normally used to generate FM in practical systems. It is not able to produce enough deviation as required in actual applications.

2.4.1 Varactor Diode

A varactor diode may be used to generate FM directly. All reverse-biased diodes exhibit a junction capacitance that varies inversely with the amount of reverse bias. A diode that is physically constructed so as to enhance this characteristic is termed a varactor diode.

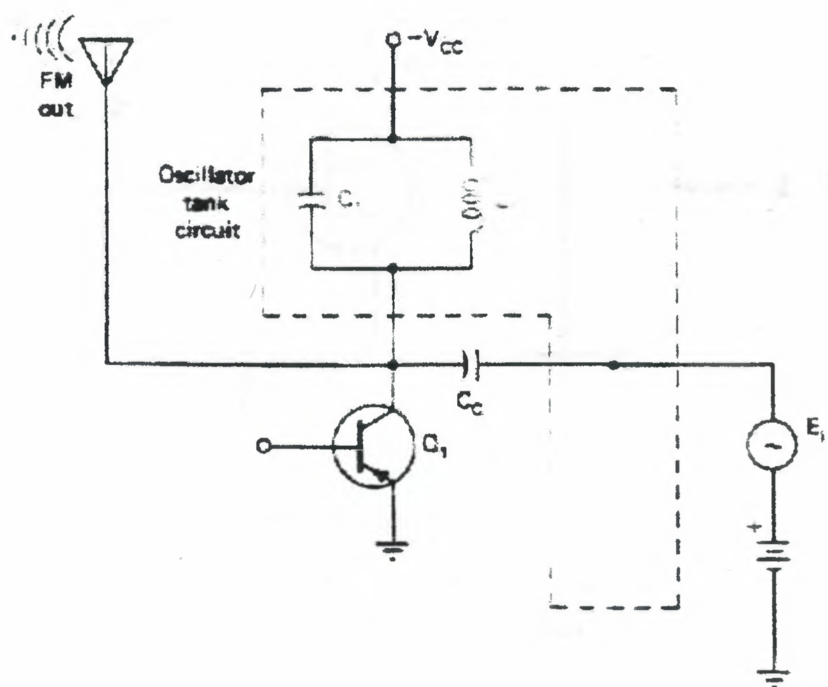

Figure 2.11 Varactor diode modulator.

Figure 2.11 shows a schematic of a varactor diode modulator. With no intelligence signal (E_i) applied, the parallel combination of C_1, L_1 and D_1's capacitance forms the resonant carrier frequency. The coupling capacitor C_c isolates the dc levels and intelligence signal while looking like a short to the high-frequency carrier. When the intelligence signal, E_i, is applied to the varactor diode, its reverse bias is varied, which causes the diode's Junction capacitance to vary in step with E_i. The oscillator frequency is subsequently varied as required for FM, and the FM signal is available at Q_1's collector.

2.4.2 Reactance Modulator

While the varactor diode modulator can be called a reactance modulator, the term is usually applied to those in which an active device is made to look like a variable reactance. The reactance modulator is a very popular means of FM generation. In order to determine how an active device can be made to look like a reactance, consider the JFET in Fig. 2.12.

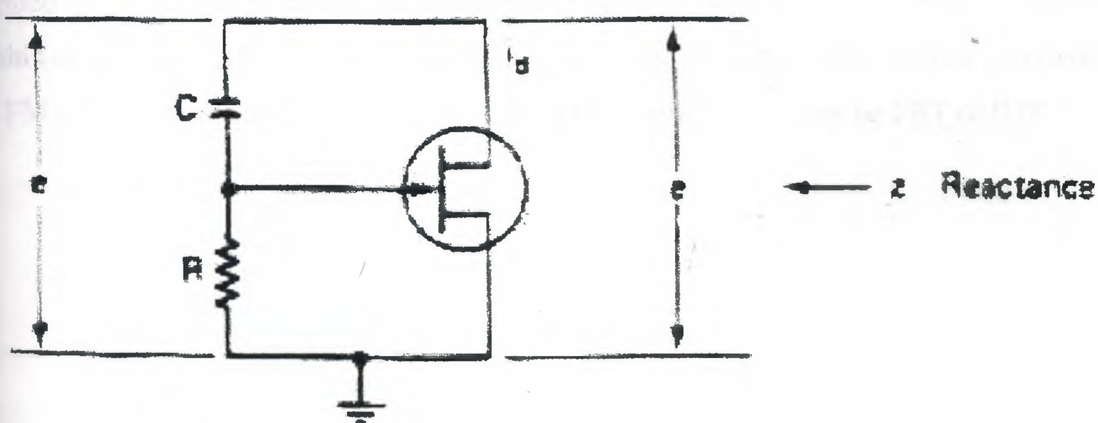

Figure 2.12 Reactance circuit

Assuming the JFET's gate current to be nearly zero, we obtain

$$E_G = I_1 R \qquad (2.8)$$

But I_1 is

$$I_1 = E/R - jX_c \tag{2.9}$$

And substituting Eq. (2.9) into Eq. (2.8),we have

$$E_G = (R \times E)/(R - jX_c) \tag{2.10}$$

The JFET drain current , I_D, is

$$I_D = G_M E_G \tag{2.11}$$

where G_M is the JFET's transconductance

If a modulating signal is applied to the JFET gate in Fig. 2.12, the amount of capacitance will vary in step because the JFET transconductance,G_M, is varied by an applied gate voltage. All that is necessary to generate FM, then. is to connect a JFET's drain (or BJT's collector) to ground terminals across an oscillator's tank circuit to provide an FM generator as shown in Fig. 2.13. The active device can either be FET or BJT.

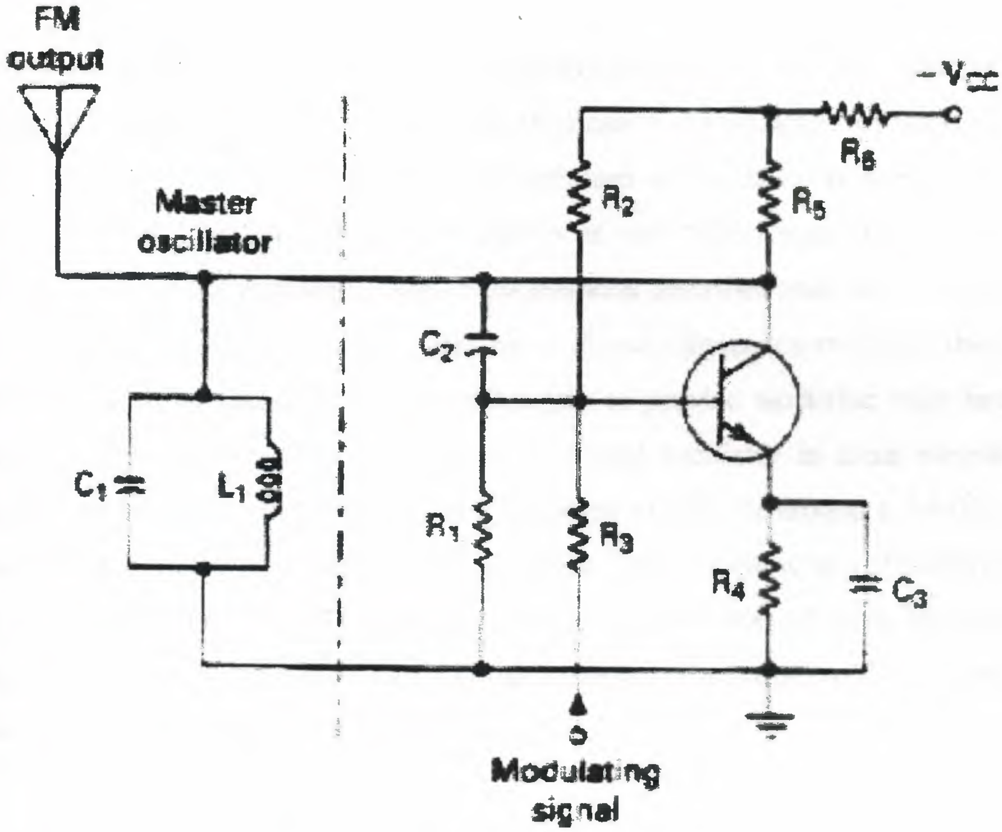

Figure 2.13 Reactance modulator

2.4.3 LIC VCO FM Generation

A voltage-controlled oscillator (VCO) produces an output frequency that is directly proportional to a control voltage level. The circuitry necessary to produce such an oscillator with a high degree of linearity between control voltage and frequency was formerly prohibitive on a discrete component basis, but now that low-cost monolithic LIC VCOs are available, they make FM generation extremely simple.

2.4.4 Crosby Modulator

Now that three practical methods of FM generation have been shown varactor diode, reactance modulator, and the VCO.it is time to consider the weakness of these methods. Notice that in no case was a crystal oscillator used as the basic reference or carrier frequency. The stability of the carrier frequency is very tightly controlled by the FCC, and that stability is not attained by any of the methods described thus far. Because of the high Q of crystal oscillators, it is not possible to directly frequency-modulate them, their frequency cannot be made to deviate sufficiently to provide workable wide band FM systems. It is possible to directly modulate a crystal oscillator in some narrow band applications. If a crystal is modulated to a deviation of ±50 Hz around a 5-MHz center frequency and both are multiplied by 100, a narrow band system with a 500-MHz carrier ±5 kHz deviation results. One method of circumventing this dilemma for wideband systems is to provide some means of automatic frequency control (AFC) to correct any carrier drift by comparing it to a reference crystal oscillator.

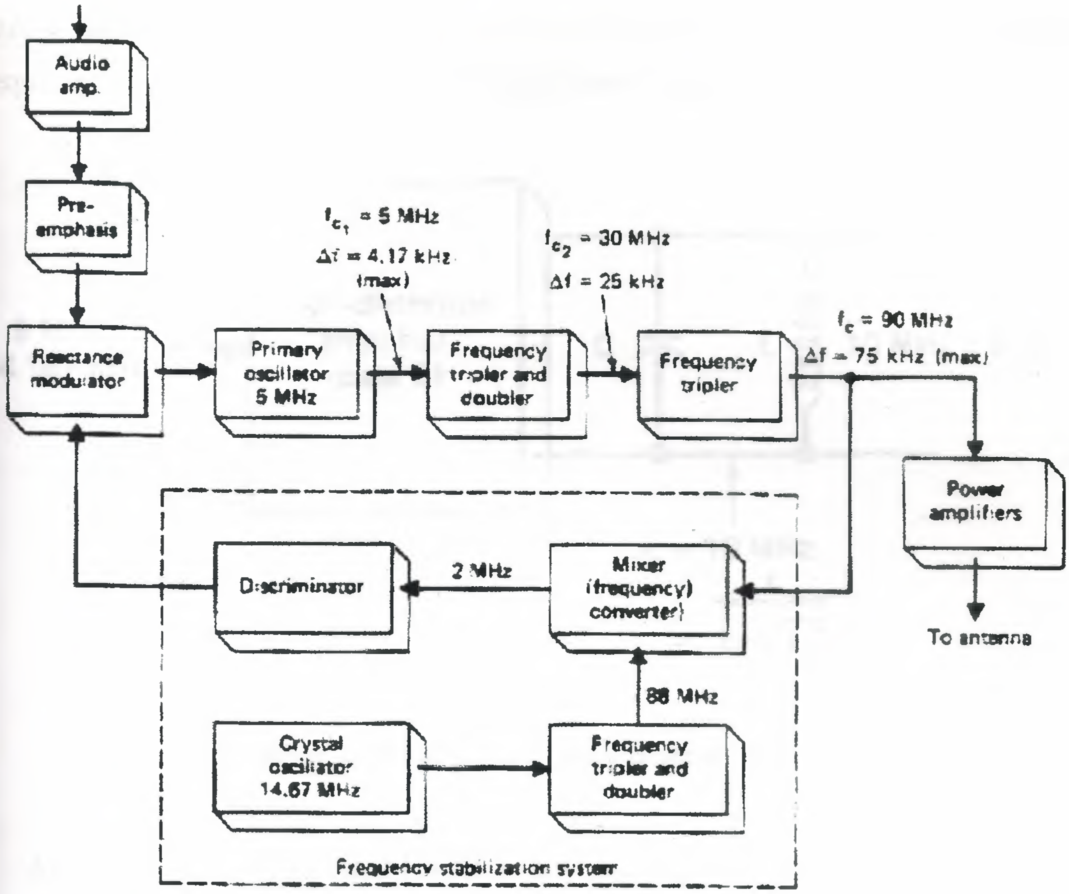

Figure 2.15 Crosby direct FM transmitter.

FM systems utilizing direct generation with AFC are called Crosby systems. A Crosby direct FM transmitter for a standard broadcast station at 90 MHz is shown in Fig. 2.15. Notice that the reactance modulator starts at an initial center frequency of 5 MHz and has a maximum deviation of ±4.167 kHz. This is a typical situation in that reactance modulators cannot provide deviations exceeding about ±5 kHz and still offer a high degree of linearity (i.e., delta f directly proportional to the modulating voltage amplitude). Consequently, frequency multipliers are utilized to provide a × 18 multiplication up to a carrier frequency of 90 MHz (18 × 5 MHz) with a ±75-kHz (18 x 4.167 kHz) deviation. Notice that both the carrier and deviation are multiplied by the multiplier.

Frequency multiplication is normally obtained in steps of ×2 or ×3 (doublers or triplers). The principle involved is to feed a frequency rich in harmonic distortion (i.e., from a class C amplifier) into an LC tank circuit tuned to two or three times the input frequency. The harmonic is then the only significant output, as illustrated in Fig, 2.16.

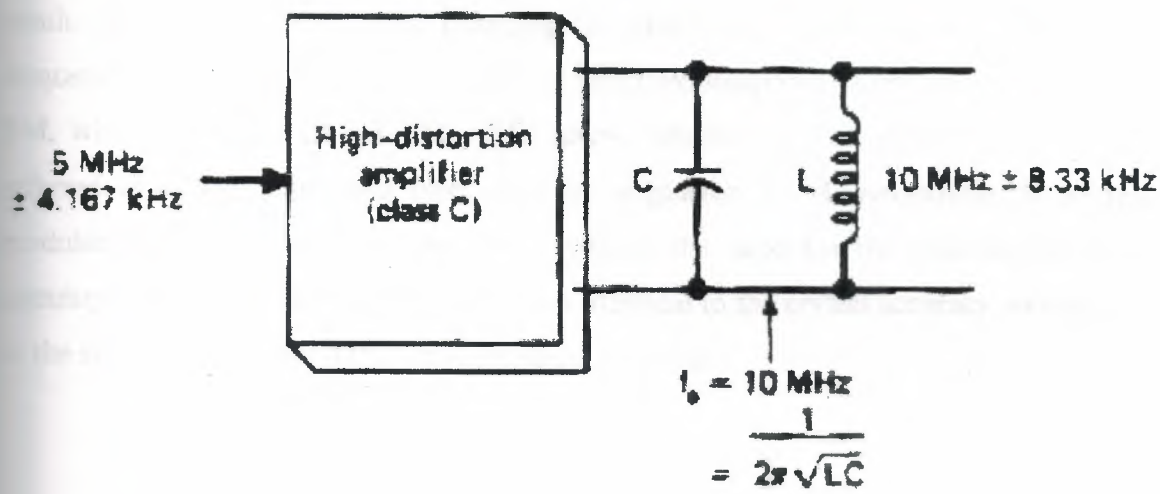

Figure 2.16 Frequency multiplication (doubler)

After the × 18 multiplication (3×2×3) shown in Fig. 2.15, the FM exciter function is complete. The term exciter is often used to denote the circuitry that generates the modulated signal. The excited output goes to the power amplifiers for transmission and to the frequency stabilization system. The purpose of this system is to provide a control voltage to the reactance modulator whenever it drifts from its desired 5-MHz value. The control (AFC) voltage then varies the reactance of the primary 5-MHz oscillator slightly to bring it back on frequency.

The mixer in Fig. 2.15 has the 90-MHz carrier and 88-MHz crystal oscillator signal as inputs. The mixer output only accepts the difference component of 2 MHz, which is fed to the discriminator. A discriminator is the opposite of a VCO, in that it provides a dc level output based on the frequency input. The discriminator output in Fig. 2.15 will be zero if it has an input of exactly 2 MHz, which occurs when the transmitter is at precisely 90 MHz. Any carrier drift up or down causes the discriminator output to go positive or negative, resulting in the appropriate primary oscillator readjustment.

2.5 Indirect FM Generation

If the phase of a crystal oscillator's output is varied, phase modulation (PM) will result. As discussed previously, changing the phase of a signal indirectly causes its frequency to be changed. We thus find that direct modulation of a crystal is possible via PM, which indirectly creates FM. This indirect method of FM generation is usually referred to as the Armstrong type, after its originator, E. H. Armstrong. It permits modulation of a stable crystal oscillator without the need for the cumbersome AFC circuitry and also provides carrier accuracies identical to the crystal accuracy, as opposed to the slight degradation of the Crosby system's accuracy.

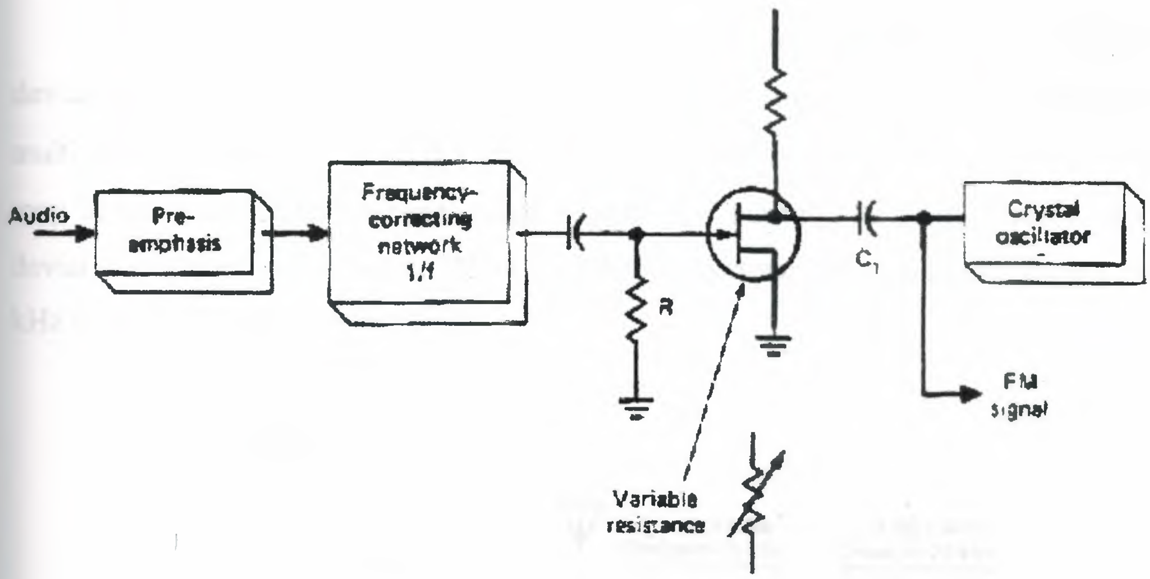

Figure 2.17 Indirect FM via PM

A simple Armstrong modulator is depicted in Fig. 2.17. The JFET is biased in the ohmic region by keeping V_{DS} low. In that way it presents a resistance from drain to source that is made variable by the gate voltage (the modulating signal). Notice that the modulating signal is first given the standard preemphasis and then applied to a frequency-correcting network. This network is a low-pass RC circuit (an integrator) that makes the audio output amplitude inversely proportional to its frequency. This is necessary because in phase modulation, the frequency deviation created is not only proportional to modulating signal amplitude (as desired for FM) but also to the modulating signal frequency (undesired for FM). Thus, in PM if a 1-V, 1-kHz modulating signal caused a 100-Hz deviation, a 1-V, 2-kHz signal would cause a 200-Hz deviation instead of the same deviation of 100 Hz if that signal were applied to the $1/f$ network.

The Armstrong modulator of Fig. 2.17 indirectly generates FM by changing the phase of a crystal oscillator's output. That phase change is accomplished by varying the phase angle of an RC network (C_1 and the JFET's resistance), in step with the frequency-corrected modulating signal.

2.5.1 Obtaining Wideband Deviation

The indirectly created FM is not capable of much frequency deviation. A typical deviation is 50 Hz out of 1 MHz (50 ppm). Thus, even with a ×90 frequency multiplication, a 90-MHz station would have a deviation of 90×50 Hz = 4.5 kHz. This may be adequate for narrowband communication FM but falls far short of the 75-kHz deviation required for broadcast FM. A complete Armstrong FM system providing a 75-kHz deviation is shown in Fig. 2.18.

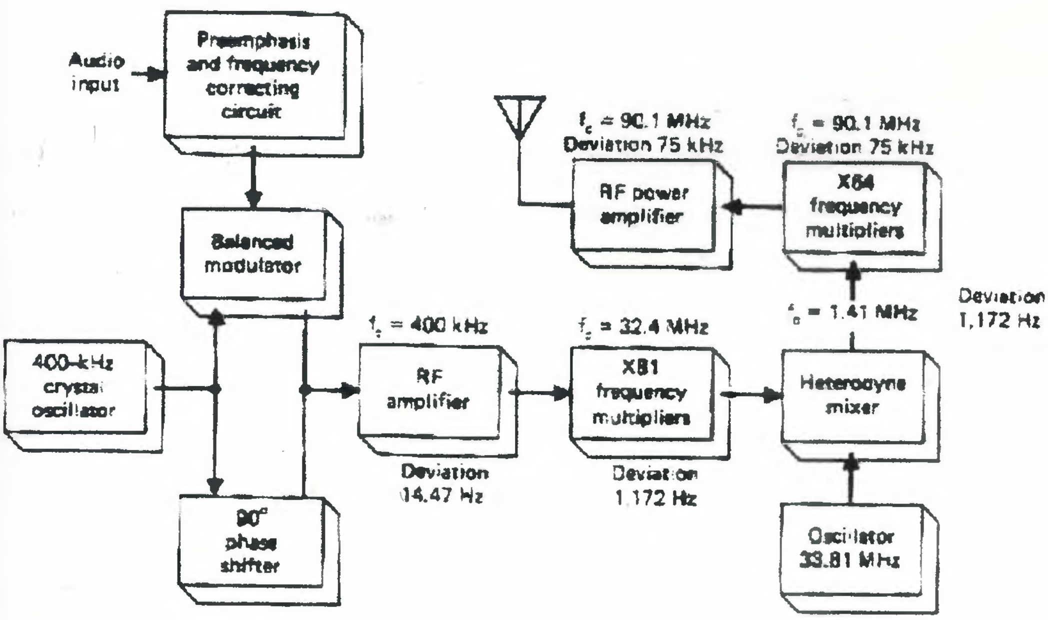

Figure 2.18 PLL FM transmitter block diagram

It uses a balanced modulator and 90° phase shifter to phase-modulate a crystal oscillator. Sufficient deviation is obtained by a combination of multipliers and mixing. The ×81 multipliers (3×3×3×3) raise the initial 400 kHz ± 14.47 Hz signal to 32.4 MHz ± 1172 Hz. The carrier and deviation are multiplied by 81. Applying this signal to the mixer, which also has a crystal oscillator signal input of 33.81 MHz, provides an output component (among others) of 33.81 MHz - (32.4 MHz ± 1172 Hz) or 1.41 MHz ± 1172 Hz. Notice that the miser output changes the center frequency without changing the deviation. Following the mixer, the ×64 multipliers accept only the mixer difference output component of 1.41 MHz ± 1172 Hz and raise that to (64 ×1.41 MHz) ± (64 × 1172 Hz) or the desired 90 MHz ± 75 kHz.

2.6 Phase-Locked-Loop FM Transmitter

Block Diagram

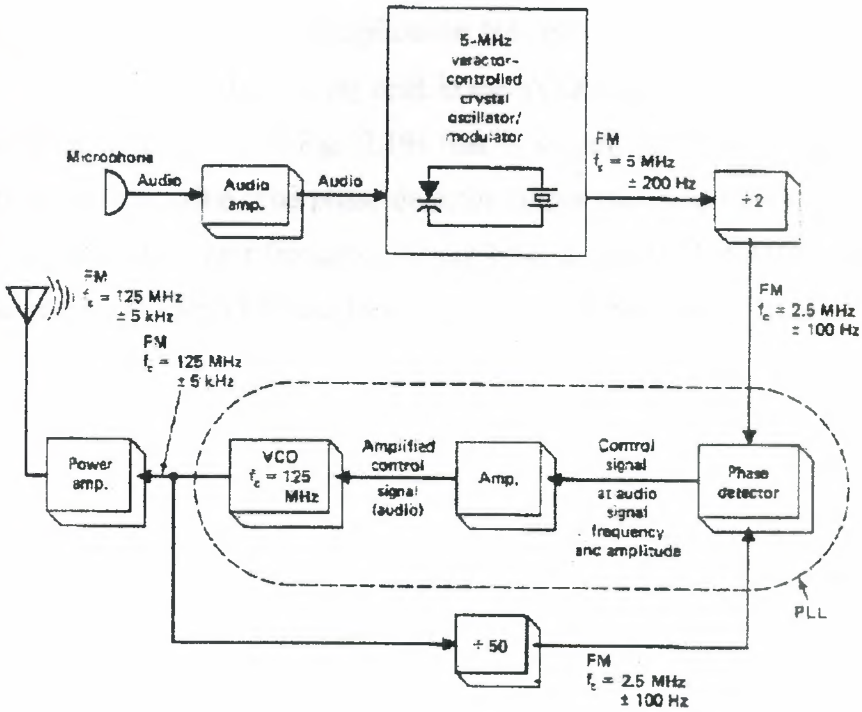

Figure 2.19 PLL FM transmitter block diagram.

The block diagram shown in Fig. 2-19 provides a very practical way to fabricate an FM transmitter. The amplified audio signal is used to frequency modulate a crystal oscillator. The crystal frequency is "pulled" slightly by the variable capacitance exhibited by the varactor diode. The approximate ±200-Hz deviation possible in this fashion is adequate for narrowband systems. The FM output from the crystal oscillator is then divided by 2 and applied as one of the inputs to the phase detector of a phase-locked-loop (PLL) system. As indicated in Fig. 5.19, the other input to the phase detector is the same and its output is therefore (in this case) the original audio signal. Th input control signal to the VCO is therefore the same audio signal and its output will be its free-running value of 125 MHz ± 5 kHz. which is set up to be exactly 50 times the 2.5-MHz value of the divided-by-2 crystal frequency of 5 MHz.

The FM output signal from the VCO is given power amplification and then driven into the transmitting antenna. This output is also sampled by a ÷50 network, which provides the other FM signal input to the phase detector. The PLL system effectively provides the required ×50 multiplication but, more important, provides the transmitter's required frequency stability. Any drift in the VCO center frequency causes an input to the phase detector (input 2 in Fig. 2.19) that is slightly different from the exact 2.5-MHz crystal reference value. The phase detector output therefore develops an error signal that corrects the VCO center frequency output back to exactly 125 MHz, This dynamic action of the phase detector/VCO and feedback path is the basis of a PLL.

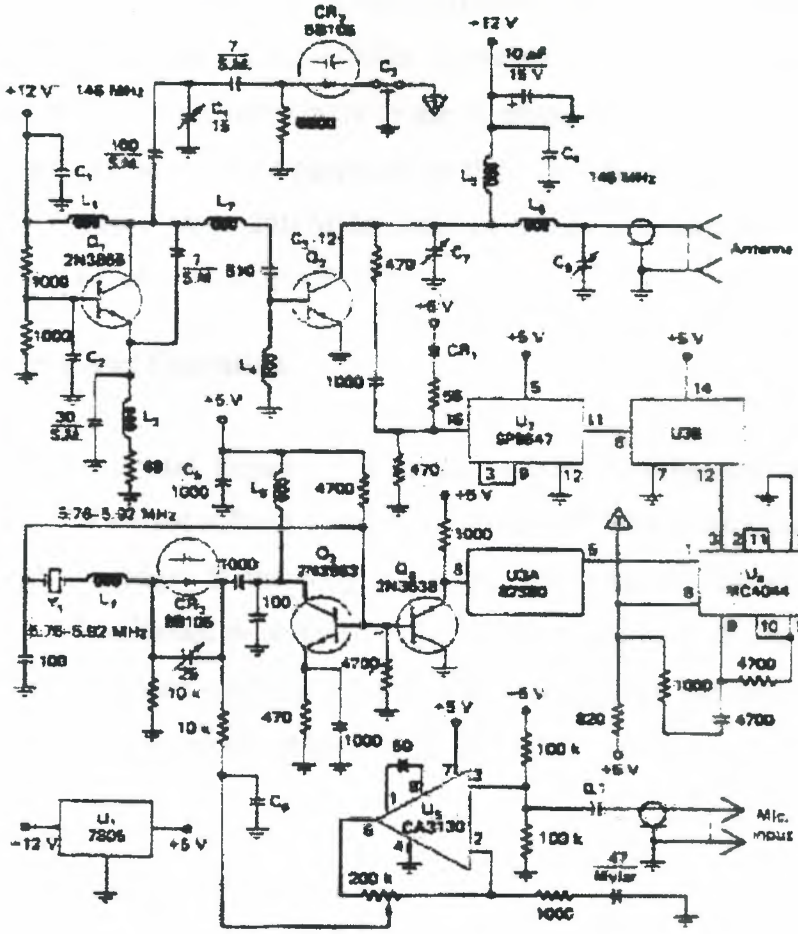

Figure 2.20 PLL FM transmitter schematic.

2.6.1 Circuit Description

The circuit schematic shown in Fig. 2.20 is a practical working system for the block diagram of Fig. 2.19. This circuitry approach eliminates the cumber-some oscillator-multiplier chain approach and allows for a very compact transmitter. The microphone input is amplified by U_5, which is an RCA CA3130 IC. Its output is used to "pull" the varactor-controlled crystal oscillator package comprised of Y_1 , CR_3, and Q_3. Its output is amplified by Q_4 and then divided by 2 by the U_{3A} IC. Refer back to the block diagram in Fig. 2.19 as an aid in this circuit description. The output of the U_{3A} is applied as one of the inputs to the U_4 phase detector amplifier IC. The U_4 output at pin 8 is applied to the VCO made up of varactor diode CR_2 and Q_1. Its output (about 200 mW) is applied to the power amplifier stage {Q_2), which provides about 2 W into the antenna. Its output is sampled by the U_2 IC, which is an emitter-coupled logic (ECL) device that provides a $\div 10$ function at frequencies up to250 MHz. Its output is then $\div 5$ by the TTL U_{3B} IC. Its output is then applied as the other input to the U_4 phase detector/amplifier IC. The values in this schematic are selected for operation on the 146-MHz amateur band, but operation can be accomplished at up to 250 MHz, with the U_7; ECL divider IC being the limiting upper frequency factor.

2.6.2 Alignment And Operation

The reference crystal frequency is determined by dividing the desired operating frequency by 25. Varactor voltage is monitored with a VTVM or oscilloscope while C_1 is varied through its range. If the loop is locked, the varactor voltage will vary with adjustment of C_1 and should be adjusted to 2.5 V. The transmitter should be terminated in a non reactive 50-Ω load, and the RF amplifier adjusted for maximum power output. Some means of determining deviation will be necessary, and the transmitter will then be ready for use.

2.6.3 Operation On Other Bands

A transmitter may be constructed for use on the 200-MHz band by redesigning the oscillator and RF amplifier tuned circuits to resonate in that band Q_1 and Q_2 will operate efficiently at frequencies up to 400 MHz. Crystal frequency is determined in the manner indicated previously. If separate oscillator-amplifier modules are constructed for 144 and 200 MHz, or perhaps even 50 MHz and switched electronically with ECL gates, it is possible to operate on several bands with the same phase-locked-loop components, at a considerable cost savings. It is also possible to select a low-power oscillator and an unmodulated crystal oscillator to generate the LO signal for a receiver. ECL divider; are available which allow application of this circuit at higher frequencies, but a frequency division of more than 50 is required in order that the maximum operating frequency of U_4 not be exceeded.

2.7 Stereo FM

The advent of stereo records and tapes and the associated high-fidelity play-back equipment in the 1950s led to the development of stereo FM transmissions as authorized by the FCC in 1961. Stereo systems involve generating two separate signals, as from the left and right sides of a concert hall performance. When played back on left and right speakers, the listener gains greater spatial dimension or directivity.

A stereo radio broadcast requires that two separate 30-Hz to 15-kHz signals be used to modulate the carrier in such a way that the receiver can extract the "left" and "right" channel information and separately amplify them into their respective speakers. In essence, then, the amount of information to be transmitted is doubled in a stereo broadcast. Hartley's law tells us that either the bandwidth or time of transmission must therefore be doubled, but this is not practical. The problem was solved by making more efficient use of the available bandwidth (200 kHz) by frequency *multiplexing the* two required modulating signals, Multiplex operation is the simultaneous transmission of two or more signals on one carrier.

2.7.1 Modulating Signal

The system approved by the FCC is compatible in that a stereo broadcast received by a normal FM receiver will provide an output equal to the sum of the left plus right channel (L + R), while a stereo receiver can provide separate left and right channel signals.

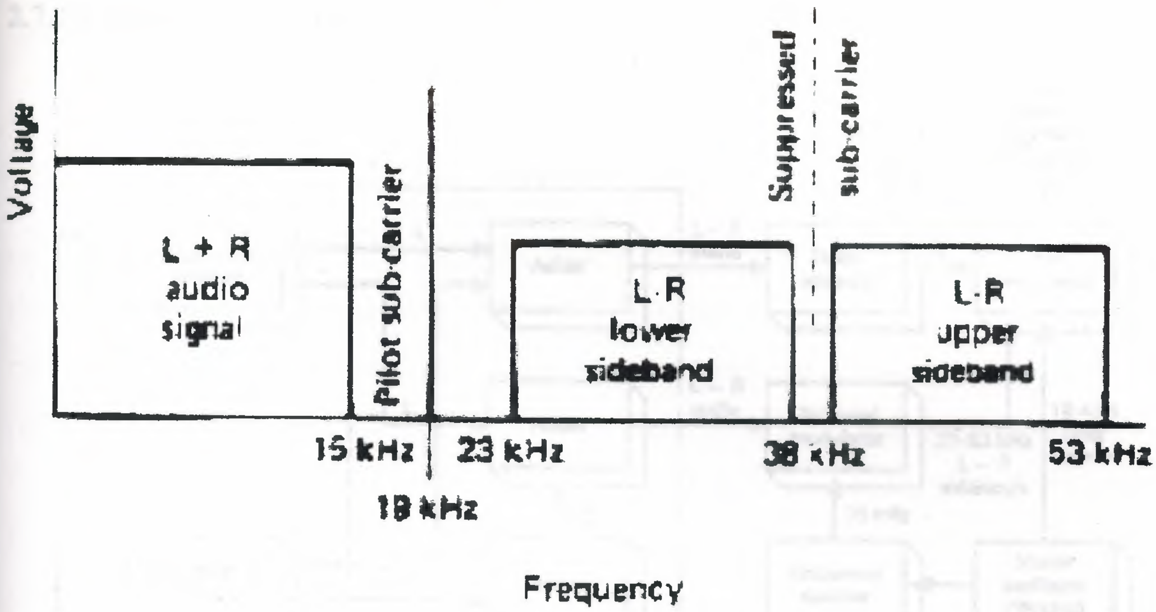

Figure 2.21 Composite modulating signals.

The stereo transmitter has a modulating signal, as shown in Fig. 2.21. Notice that the sum L + R modulating signal extends from 30 Hz to 15 kHz just as does the full audio signal used to modulate the carrier in standard FM broadcasts. However, a signal corresponding to the left channel minus right channel (L - R) extends from 23 to 53 kHz. In addition, a 19-kHz pilot subcarrier is included in the composite stereo modulating signal.

The two different signals (L +R and L - R) are used to modulate the carrier. The signal is an example of frequency-division multiplexing, in that two different signals are multiplexed together by having them exist in two different frequency ranges.

2.7.2 FM Stereo Generation

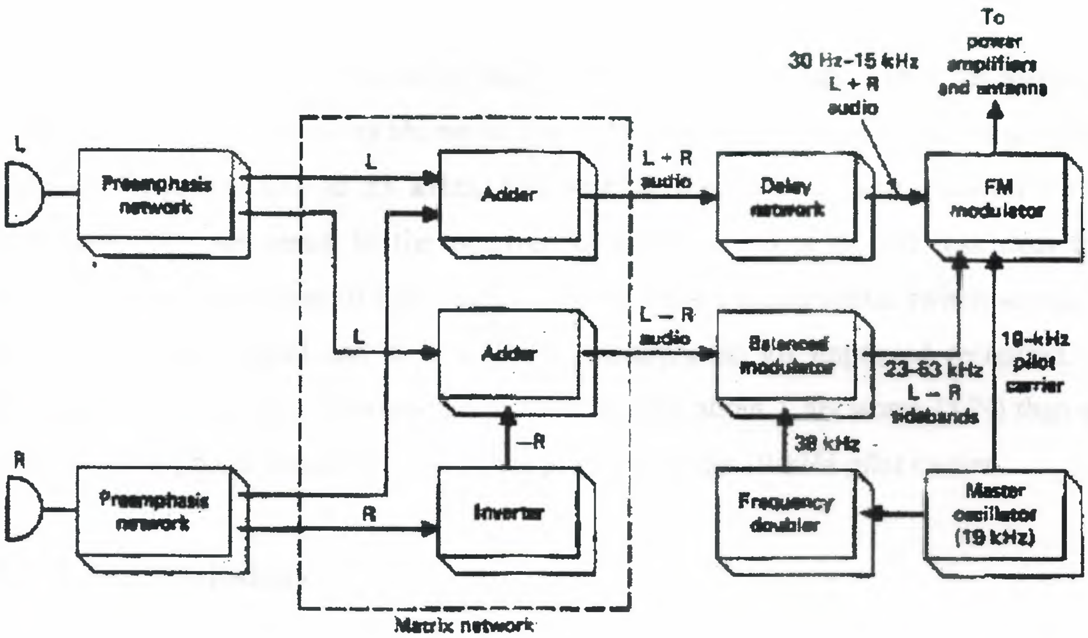

Figure 2.22 Stereo FM transmitter.

The block diagram in Fig. 2.22 shows the method whereby the composite-modulating signal is generated and applied to the FM modulator for subsequent transmission. The left and right channels are picked up by their respective microphones and individually preemphasized. They are then applied to a matrix network that inverts the right channel, giving a -R signal, and then combines (adds) L and R to provide an (L + R) signal and also combines L and -R to provide the (L - R) signal. The two outputs are still 30-Hz to 15-kHz audio signals at this point. The (L - R) signal and a 38-kHz carrier signal are then applied to a balanced modulator that suppresses the carrier but provides a double-sideband (DSB) signal at its output. The upper and lower sidebands extend from 30 Hz to 15 kHz above and below the suppressed 38 kHz carrier and therefore range from 23 kHz (38 kHz - 15 kHz) up to 53 kHz (38 kHz + 15 kHz). Thus, the L - R signal has been translated from audio upto a higher frequency so as to keep it separate from the 30 Hz to 15 kHz (L + R) signal. The (L + R) signal is given a slight delay so that both signals are applied to the FM modulator in time phase due to the slight delay encountered by the (L - R) signal in the balanced modulator. The 19-kHz master oscillator in Fig. 2.22

is applied directly to the FM modulator and also doubled in frequency to 38 kHz for the balanced modulator carrier input.

Stereo FM is more prone to noise than monophonic broadcasts. The L -R signal is weaker than the L + R signal, as shown in Fig. 2.21. The L – R signal is also at a higher modulating frequency (23 to 53 kHz), and both of these effects cause poorer noise performance. The net result to the receiver is a S/N of about 20 dB less than the monophonic signal. Because of this, some receivers have a mono/stereo switch so that a noisy (weak) stereo signal can be changed to monophonic for improved reception. A stereo signal received by a monophonic receiver is only about 1 dB worse (S/N) than an equivalent monophonic broadcast, due to the presence of the 19-kHz pilot carrier.

2.8 FM Transmissions

There are five major categories in which FM is used :

1. Noncommercial broadcast at 88 to 90 MHz .

2. Commercial broadcast with 200-kHz channel bandwidths from 90 to 108 MHz .

3. Television audio signals with 50-kHz channel bandwidths at 54 to 88 MHz, 174 to 216 MHz, and 470 to 806 MHz .

4. Narrowband public service channels from 108 to 174 MHz and in excess of 806 MHz .

5. Narrowband amateur radio channels at 29.6 MHz, 52 to 53 MHz, 144 to 147.99 MHz. 440 lo 450 MHz, and in excess of 902 MHz

The output powers range from milliwatt levels for the amateurs up to 100 kW for broadcast FM. You will note that FM is not used at frequencies below about 30 MHz. This is due to the phase distortion introduced to FM signals by the earth's ionosphere at frequencies below approximately 30 MHz, Frequencies above 30 MHz are transmitted "line-of-sight" and are not significantly affected by the ionosphere. This situation explains the limited range (normally, 70 to 80 mi) of FM transmission due to the earth's curvature.

Another advantage that FM has over SSB and AM, other than superior noise performance, is the fact that low-level modulation can be used with subsequent highly efficient class C power amplifiers. Since the FM waveform does not vary in amplitude, the intelligence is not lost by class C power amplification as it is for AM and SSB. Recall that a class C amplifier tends to provide a constant output amplitude due to the LC tank circuit flywheel effect. Thus, there is no need for high-power audio amplifiers in an FM transmitter and, more important, all the power amplification takes place at about 90% efficiency (class C) as compared to a maximum of about 70% for linear power amplifiers.

CHAPTER 3

FM RECEPTION

3.1 FM Receiver

The basic FM receiver uses the superheterodyne principle. In block diagram form, it has many similarities to the receivers. In Fig. 3.1, the only apparent differences are the use of the word discriminator in place of detector, the addition of a deemphasis network, and the fact that AGC may or may not be used as indicated by the dashed lines.

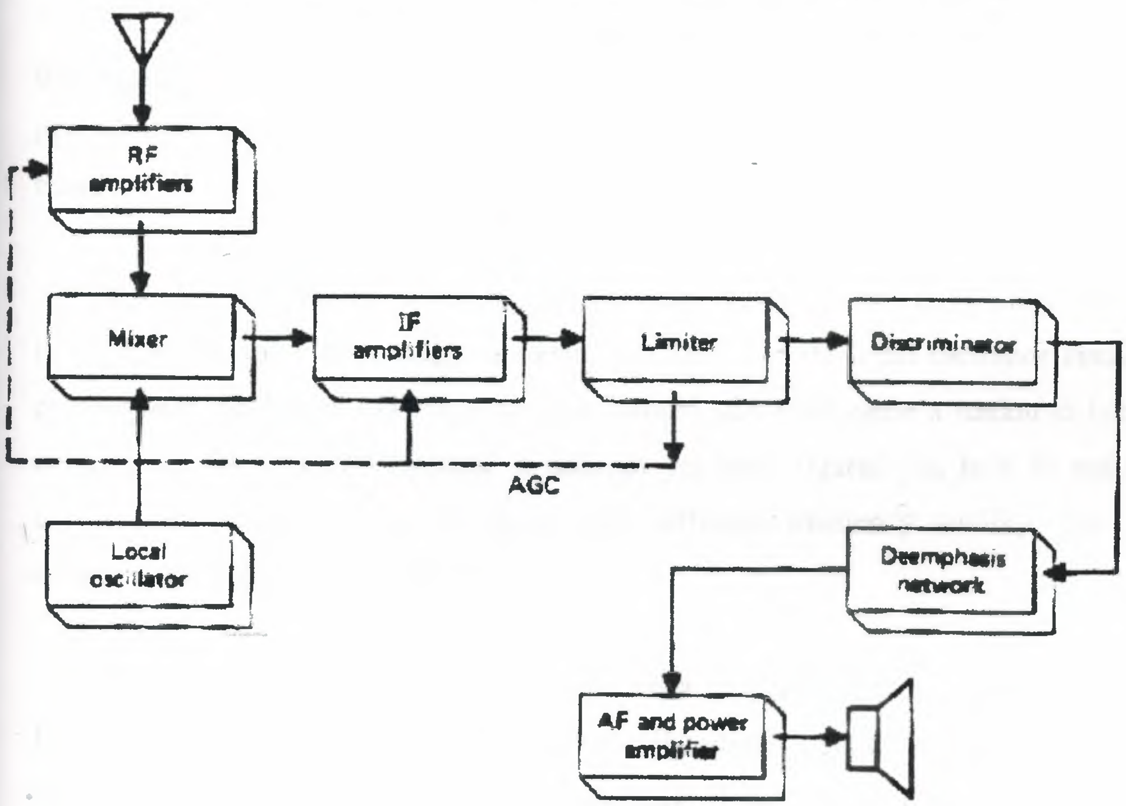

Figure 3.1 FM receiver block diagram.

The discriminator extracts the intelligence from the high-frequency carrier and can also be called the detector, as in AM receivers. By definition, however, a discriminator is a device in which amplitude variations are derived in response to frequency or phase variations and it is the preferred term for describing an FM demodulator.

The deemphasis network following demodulation is required to bring the high-frequency intelligence back to the proper amplitude relationship with the lower frequencies. Recall that the high frequencies were preemphasized at the transmitter to provide them with greater noise immunity.

The fact that AGC is optional in an FM receiver may be surprising to you. From your understanding of AM receivers, you know that AGC is essential to their satisfactory operation. However, the use of limiters in FM receivers essentially provides an AGC function.

Many older FM receivers also included an automatic frequency (AFC) function. This is a circuit that provides a slight automatic control over the local oscillator circuit. It compensates for drift in LO frequency that would otherwise cause a station to become detuned. It was necessary because it had not yet been figured out how to make an economical LC oscillator at 100 MHz with sufficient frequency stability. The AFC system is not needed in new designs.

The mixer, local oscillator, and IF amplifiers are basically similar to those discussed for AM receivers and do not require further elaboration. It should be noted that higher frequencies are usually involved, however, because of the fact that FM systems generally function at higher frequencies. The universally standard IF frequency for FM is 10.7 MHz, as opposed to 455 kHz for AM. Because of significant differences in all the other portions of the block diagram shown in Fig. 3.1, they are discussed in the following sections.

3.2 RF Amplifiers

Broadcast AM receivers normally operate quite satisfactorily without any RF amplifier. This is rarely the case with FM receivers, however, except for frequencies in excess of 1000 MHz (I GHz) when it becomes preferable to omit it. The essence of the problem is that FM receivers can function with smaller received signals than AM or SSB receivers because of their inherent noise reduction capability. This means that FM receivers can function with a lower sensitivity, and are called upon to deal with input signals of 1 μV or less as compared with perhaps a 30 μV minimum input for AM. If a 1-μV signal is fed directly into a mixer, the inherently high noise factor of an active mixer stage destroys the intelligibility of the 1-.μV signal. It is. therefore, necessary to amplify the 1-μV level in an RF stage to get the signal up to at least 10 to 20 μV before mixing occurs. The FM system can tolerate 1 μV of noise from a mixer on a 20-μV signal but obviously cannot cope with 1μ V of noise with a 1 μV signal .

This reasoning also explains the abandonment of RF stages for the ever-increasing FM systems at the 1-GHz-and-above region. At these frequencies, transistor noise is increasing while gain is decreasing. The frequency is reached where it is advantageous to feed the incoming FM signal directly into a diode mixer so as to immediately step it down to a lower frequency for subsequent amplification. Diode (passive) mixers are less noisy than active mixers.

Of course, the use of an RF amplifier reduces the image frequency problem. Another benefit is the reduction in local oscillator reradiation effects. Without an RF amp, the local oscillator signal can more easily get coupled back into the receiving antenna and transmit interference.

3.2.1 FET RF Amplifiers

Virtually all RF amps used in quality FM receivers utilize FETs as the active element. In fact, their input impedance at the high frequency of FM signals is greatly reduced because of their input capacitance. The fact that FETs do not offer any significant impedance advantage over other devices at high frequencies is not a deterrent, however, since the impedance that an RF stage works from (the antenna) is only several hundred ohms or less anyway.

The major advantage is that FETs have an input/output square-law relationship while vacuum tubes have a 3/2-power relationship and BJTs have a diode-type exponential characteristic. A square-law device has an output signal at the input frequency and a smaller distortion component at two times the input frequency, whereas the other devices mentioned have many more distortion components, with some of them occurring at frequencies close to the desired signal. The use of a FET at the critical small signal level in a receiver means that the device distortion components are easily filtered out by its tuned circuits, since the closest distortion component is two times the frequency of the desired signal. This becomes an extreme factor when you tune to a weak station that has a very strong adjacent signal. If the high-level adjacent signal gets through the input tuned circuit, even though greatly attenuated, it would probably generate distortion components at the desired signal frequency by a non-square-law device, and the result is audible noise in the speaker output. This form of receiver noise is called cross-modulation. This is similar to inter modulation distortion, which is characterized by the mixing of two undesired signals, resulting in an output component that is equal to the desired signal's frequency. The possibility of inter modulation distortion is also greatly minimized by use of FET RF amplifiers.

3.2.2 MOSFET RF Amplifiers

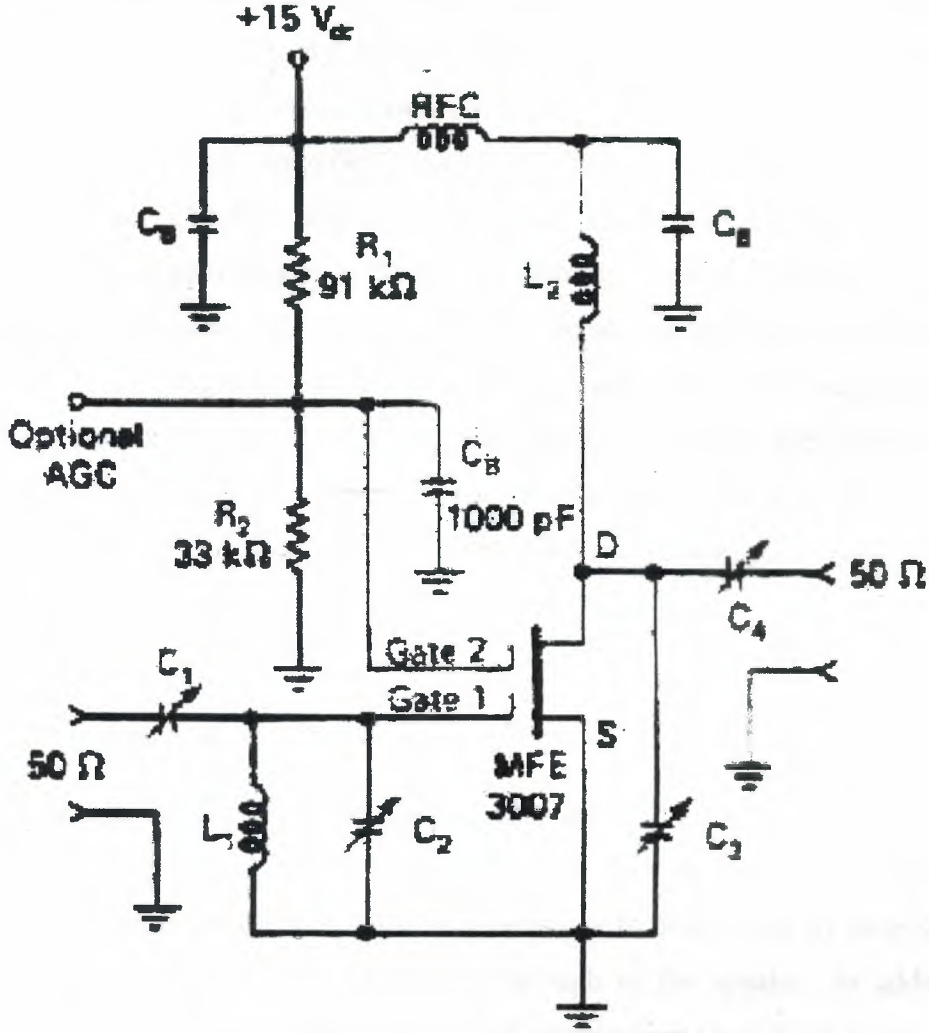

Figure 3.2 MOSFET RF amplifier.

A dual-gate, common-source MOSFET RF amplifier is shown in Fig. 3.2.The use of a dual-gate device allows a convenient isolated input for an AGC (Automatic Gain Control) level to control device gain, The MOSFETs also offer the advantage of increased dynamic range over JFETs. That is, a wider range of input signal can be tolerated by the MOSFET while still offering the desired square-law input/output relationship. A similar arrangement is often utilized in mixers since the extra gate allows for a convenient injection point for the local oscillator signal. The accompanying chart in Fig. 3.2 provides component values for operation at 100 MHz and 400-MHz center frequencies. The antenna input signal is coupled into gate 1 via the coupling/tuning network comprised of C_1, L_1, and C_2. The output signal is taken at the drain, which is coupled to the next stage by the L_2-C_3-C_4 combination. The bypass capacitor C_B next to L_2 and the radio-frequency choke (RFC) ensure that the signal frequency is not applied to the dc power supply. The RFC acts as an open to the signal while appearing as a short to dc, and the bypass capacitor acts in the inverse fashion. These precautions are necessary to RF frequencies because while power supply impedance is very low at low frequencies and dc, it looks like a high impedance to RF and can cause appreciable signal power loss. The bypass capacitor from gate 2 to ground provides a short to any high-frequency signal that may get to that point. It is necessary to maintain the bias stability set up by R_1 and R_2. The MFE 3007 MOSFET used in this circuit provides a minimum power gain of 18 dB at 200 MHz.

3.3 Limiters

A limiter is a circuit whose output is a constant amplitude for all inputs above a critical value. Its function in an FM receiver is to remove any residual (unwanted) amplitude modulation and the amplitude variations due to noise. Both of these variations would have an undesirable effect if carried through to the speaker. In addition, the limiting function also provides AGC action since signals from the critical minimum value up to some maximum value provide a constant input level to the detector. By definition, the discriminator (detector) ideally would not respond to amplitude variations anyway, since the information is contained in the amount of frequency deviation and the rate at which it deviates back and forth around its center frequency.

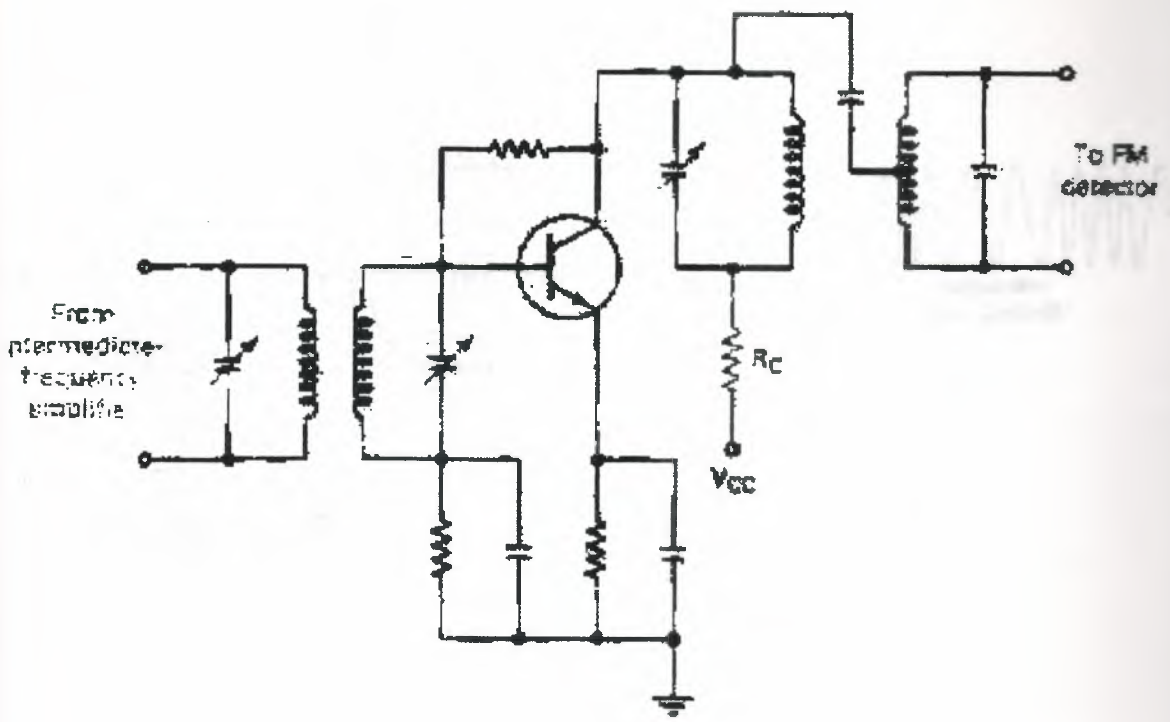

Figure 3.3 Transistor limiting circuit.

A transistor limiter is shown in Fig. 3.3. Notice the dropping resistor, R_c, which limits the dc collector supply voltage. This provides a low dc collector voltage which makes this stage very easily overdriven. This is the desired result. As soon as the input is large enough to cause clipping at both extremes of collector current, the critical limiting voltage has been attained and limiting action has started.

The input/output characteristic for the limiter is shown in Fig 3.4, and it shows the desired clipping action and the effects of feeding the limited (clipped) signal into an LC tank circuit tuned to the signal's center frequency. The natural flywheel effect of the tank removes all frequencies not near the center frequency and thus provides a sinusoidal output signal as shown. The omission of an LC circuit at the limiter output is desirable for some demodulator circuits. The quadrature detector uses the square-wave like waveform that results.

3.3.1 Limiting and Sensitivity

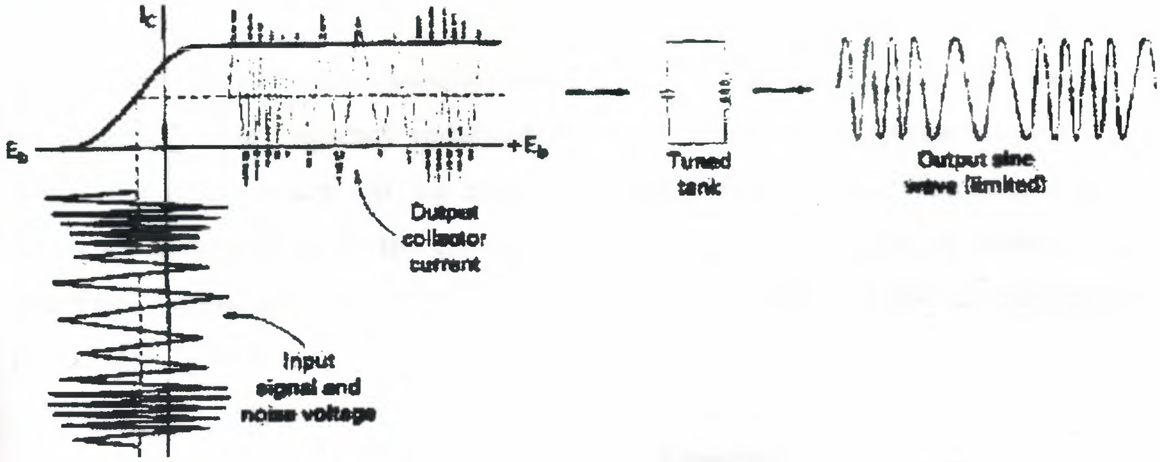

Figure 3.4 Limiter input/output and flywheel effects.

A limiter requires about 1 V of signal to begin limiting. Much amplification of the received signal is therefore needed prior to limiting, which explains its position following the IF stages. When enough signal arrives at the receiver to start limiting action, the set quiets, which means that background noise disappears. The sensitivity of an FM receiver is defined in terms of how much input signal is required to produce a specific level of quieting, normally 30 dB. This means that a good-quality receiver with a rated 1.5-µV sensitivity will have background noise 30 dB down from the desired input signal that has a !.5-µV level.

The minimum required voltage for limiting is called the quieting, threshold, or limiting knee voltage. The limiter then provides a constant-amplitude output up to some maximum value which prescribes the limiting range. Going above the maximum value results either in a reduced and/or distorted output. It is possible that a single-stage limiter will not allow for adequate range, thereby requiring a double limiter or the development of AGC control on the RF and IF amplifiers to minimize the possible limiter input range.

It is most common for today's FM receivers to use IC IF amplification.In these cases, the ICs have a built-in limiting action of very high quality (i.e.,wide dynamic range).

3.4 Discriminators

The FM discriminator (detector) extracts the intelligence that has been modulated onto the carrier via frequency variations. It should provide an intelligence signal whose amplitude is dependent on instantaneous carrier frequency deviation and whose frequency is dependent on the carrier's rate of frequency deviation. A desired output amplitude versus input frequency characteristic for a broadcast FM discriminator is provided in Fig. 3.5.

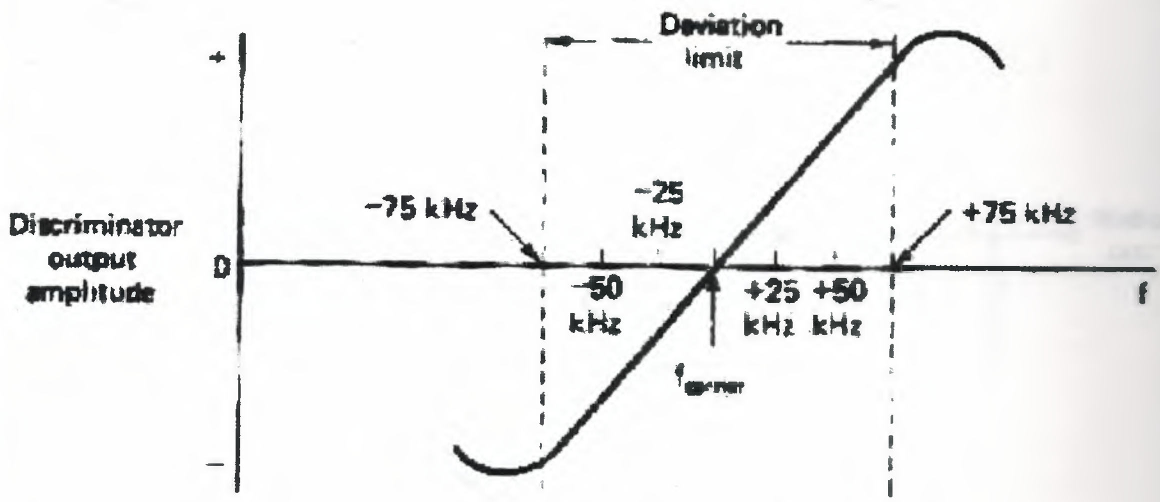

Figure 3.5 FM discriminator characteristics.

Notice that the response is linear in the allowed area of frequency deviation and that the output amplitude is directly proportional to carrier frequency deviation. Keep in mind, however, that FM detection takes place following the IF amplifiers, which means that the ±75-kHz deviation is intact but that carrier frequency translation (usually to 10.7 MHz) has occurred.

3.4.1 Slope Detector

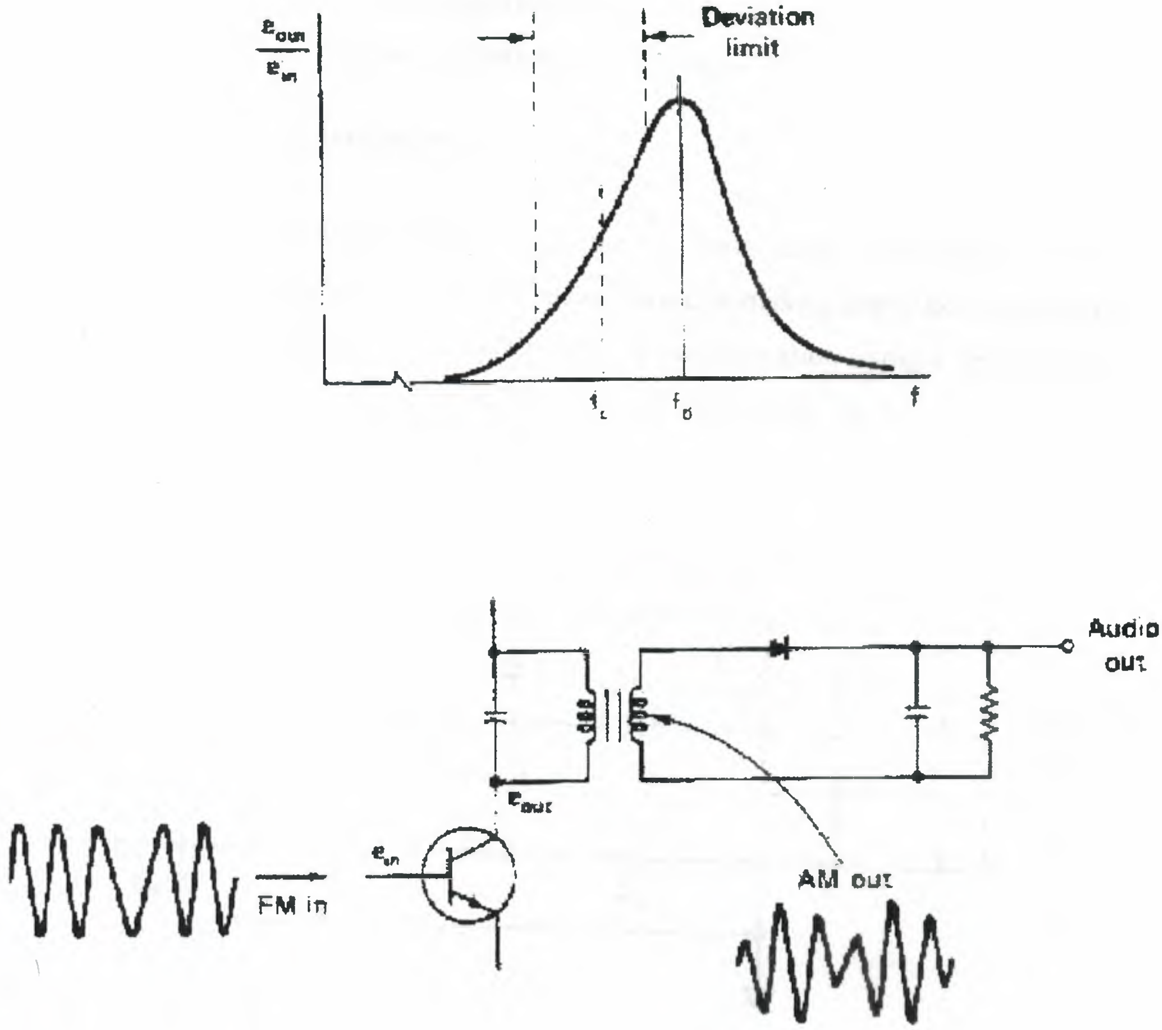

Figure 3.6 Slope detection.

The easiest FM discriminator to understand is the slope detector in Fig. 3.6. The LC tank circuit which follows the IF amplifiers and limiter is detuned from the carrier frequency so that f_c falls in the middle of the most linear region of the response curve. When the FM signal rises in frequency above f_c, the output amplitude increases while deviations below f_c cause a smaller output. The slope detector thereby changes FM into AM, and a simple diode detector then recovers the intelligence contained in the AM waveform's envelope . In an emergency, an AM receiver can be used to receive FM by detuning the tank circuit feeding the diode detector. Slope detection is not widely used in FM receivers because the slope characteristic of a tank circuit is not very linear, especially for the large-frequency deviations of wideband FM.

3.4.2 Foster-Seely Discriminator

The two classical means of FM detection are the Foster-Seely discriminator and the ratio detector. While their once widespread use is now diminishing due to new techniques afforded by ICs, they remain a popular means of discrimination using a minimum of circuitry. A typical Foster-Seely discriminator circuit is shown in Fig. 3.7.

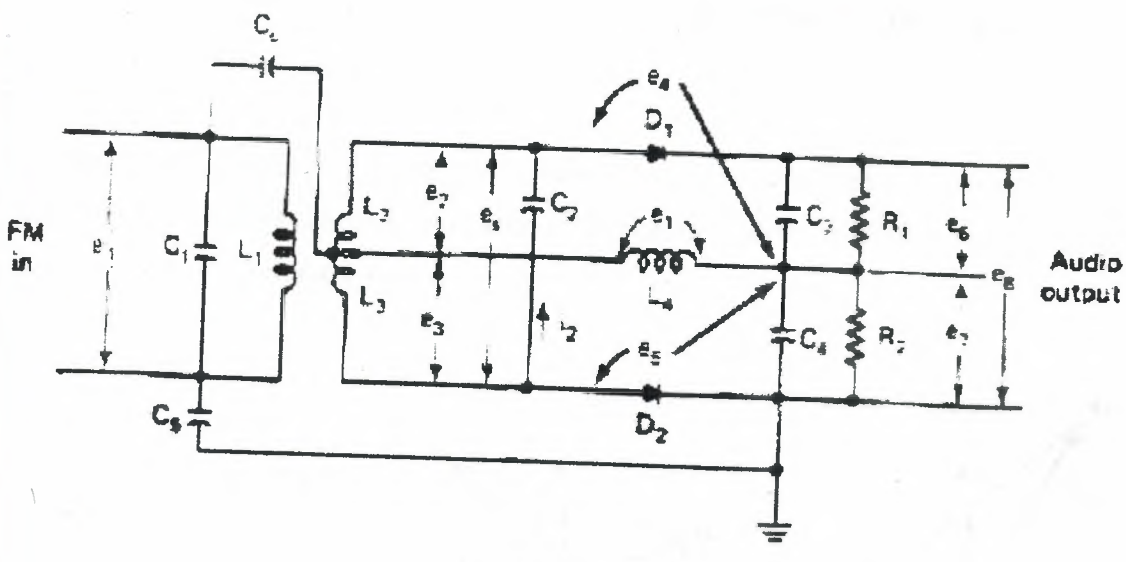

Figure 3.7 Foster-Seely discrinator.

In it, the two tank circuits [L_1C_1 and $(L_2+L_3)C_2$] are tuned exactly to the carrier frequency. Capacitors C_c, C_4, and C_5 are shorts to the carrier frequency. The following analysis applies to an unmodulated carrier input:

1. The carrier voltage e_1 appears directly across L_4 because C_c and C_4 are shorts to the carrier frequency.

2. The voltage e_s across the transformer secondary (L_2 in series with L_3) is 180° out of phase with e_1 by transformer action, as shown in Fig. 6.8(a). The circulating $L_2L_3C_2$ tank current, i_s, is in phase with e_s since the tank is resonant.

3. The current i_s, flowing through inductance L_2L_3 produces a voltage drop that lags the current, i_s, by 90°. The individual components of this voltage, e_2 and e_3, are thus displaced by 90° from i_s, as shown in Fig. 3.8(a) and are 180° out of phase with each other because they are the voltage from the ends of a center-tapped winding.

4. The voltage e_4 applied to the diode D_1, C_3, and R_1 network will be vector sum of e_1 and e_2 [Fig. 3.8(a)]. Similarly, the voltage e_5 is the sum of e_1 and e_3. The magnitude of e_6 is proportional to e_4 while e_7 is proportional to e_5.

5. The output voltage, e_8, is equal to the sum of e_6 and e_7 and is zero since the diodes D_1 and D_2 will be conducting current equally (since $e_4 = e_5$) but in opposite directions through the R_1C_3 and R_2C_4 networks.

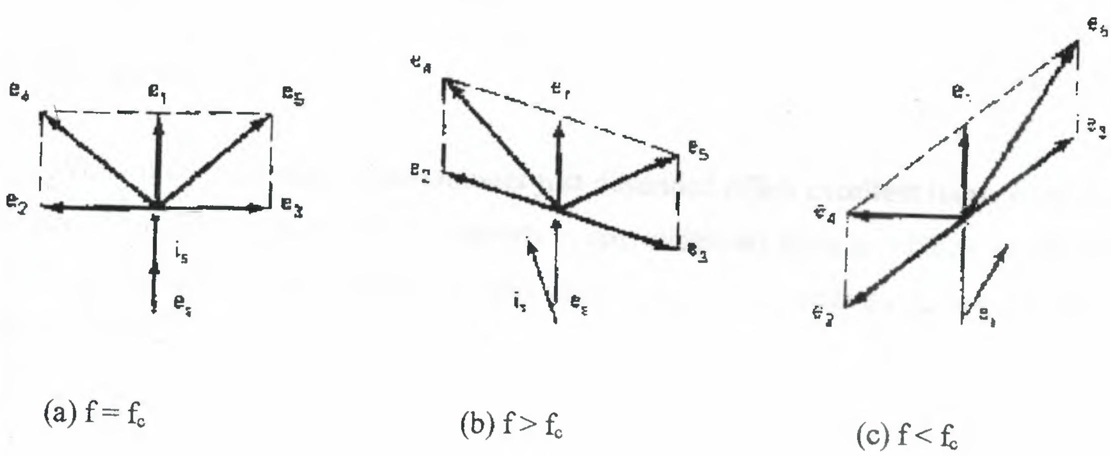

Figure 3.8 Discriminator phase relations.

The discriminator output is zero with no modulation (frequency deviation) as is desired. The following discussion now considers circuit action at some instant when the input signal e_1 is above the carrier frequency. The phasor diagram of Fig. 3.8(b) is used to illustrate this condition:

1. Voltages e_1 and e_s are as before, but e_s , now sees an inductive reactance, because the tank circuit is above resonance, Therefore, the circulating tank current, i_5 , lags e_s.

2. The voltages e_2 and e_3 must remain 90° out of phase with i_s as shown in Fig. 3.8(b). The new vector sums of $e_2 + e_1$ and $e_3 + e_1$ are no longer equal, so e_4 causes a heavier conduction of D_1 than exists for D_2 .

3. The output, e_8, which is the sum of e_6 and e_7 will go positive since the current down through R_1C_3 is greater than the current up through R_2 C_4 (e_4 is greater than e_5).

The output for frequencies above resonance (f_c) is therefore positive, while the phasor diagram in Fig. 3.8(c) shows that at frequencies below resonance the output goes negative. The amount of output is determined by the amount of frequency deviation, while the frequency of the output is determined by the rate at which the FM input signal varies around its carrier or center value.

3.4.3 Ratio Detector

While the Foster-Seely discriminator just described offers excellent linear response to wide band FM signals, it also responds to any undesired input amplitude variations. The raw detector does not respond to input variations and minimizes the required limiting before detection.

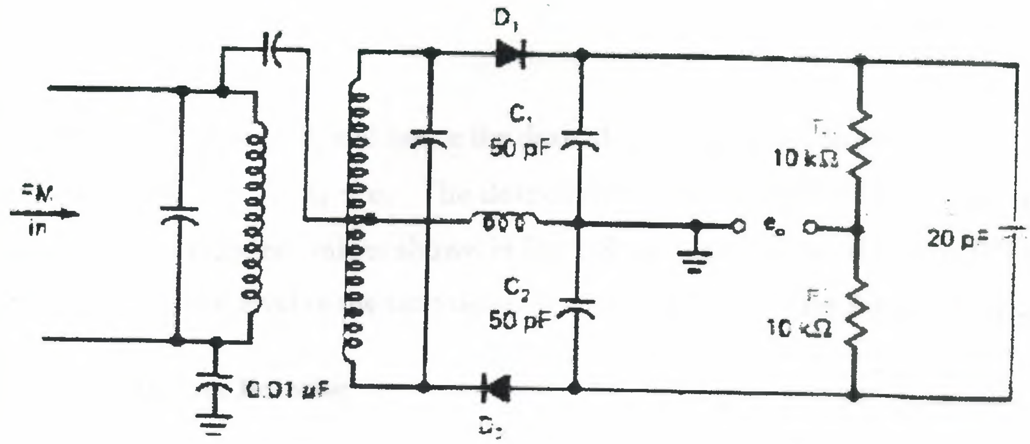

Figure 3.9 Ratio detector.

The ratio detector, shown in Fig. 3.9, is a circuit designed to respond only to frequency changes of the input signal. Amplitude changes in the input have no effect upon the output. The input circuit of the ratio detector is identical to that of the Foster-Seely discriminator circuit, The most immediately obvious difference is the reversal of one of the diodes.

The ratio detector circuit operation is similar to the Foster-Seely. A detailed analysis will therefore not be given. Notice the large electrolytic capacitor, C_5 across the R_1-R_2 combination. This maintains a constant voltage that is equal to the peak voltage across the diode input. This feature eliminates variations in the FM signal thus providing amplitude limiting. The sudden changes in the input signal's amplitude are suppressed by the large capacitor. The Foster-Seely discriminator does not provide amplitude limiting. The voltage E_s is

$$E_s = e_1 + e_2$$

And

$$E_0 = E_s / 2 - e_2 = (e_1 + e_2) / 2 - e_2 = (e_1 - e_2) / 2$$

When $f_{in} = f_c$, $e_1 = e_2$ and hence the desired zero output occurs. When $f_{in} > f_c$, $e_1 > e_2$, and when $f_{in} < f_c$, $e_1 < e_2$. The desired frequency dependent output characteristic results. The component values shown in Fig. 3.9 are typical for a 10.7-MHz IF FM input signal. The output level of the ratio detector is one-half that for the Foster-Seely circuit.

3.4.4 Quadrature Detector

The Foster-Seely and ratio detector circuits do not lend themselves to integration on a single chip due to the transformer required. This has led to increased usage of the quadrature detector and phase-locked loop (PLL). The PLL is introduced in the next section . Quadrature detectors derive their name from use of the FM signal in phase and 90° out of phase. The two signals are said to be in quadrature at a 90° angle.

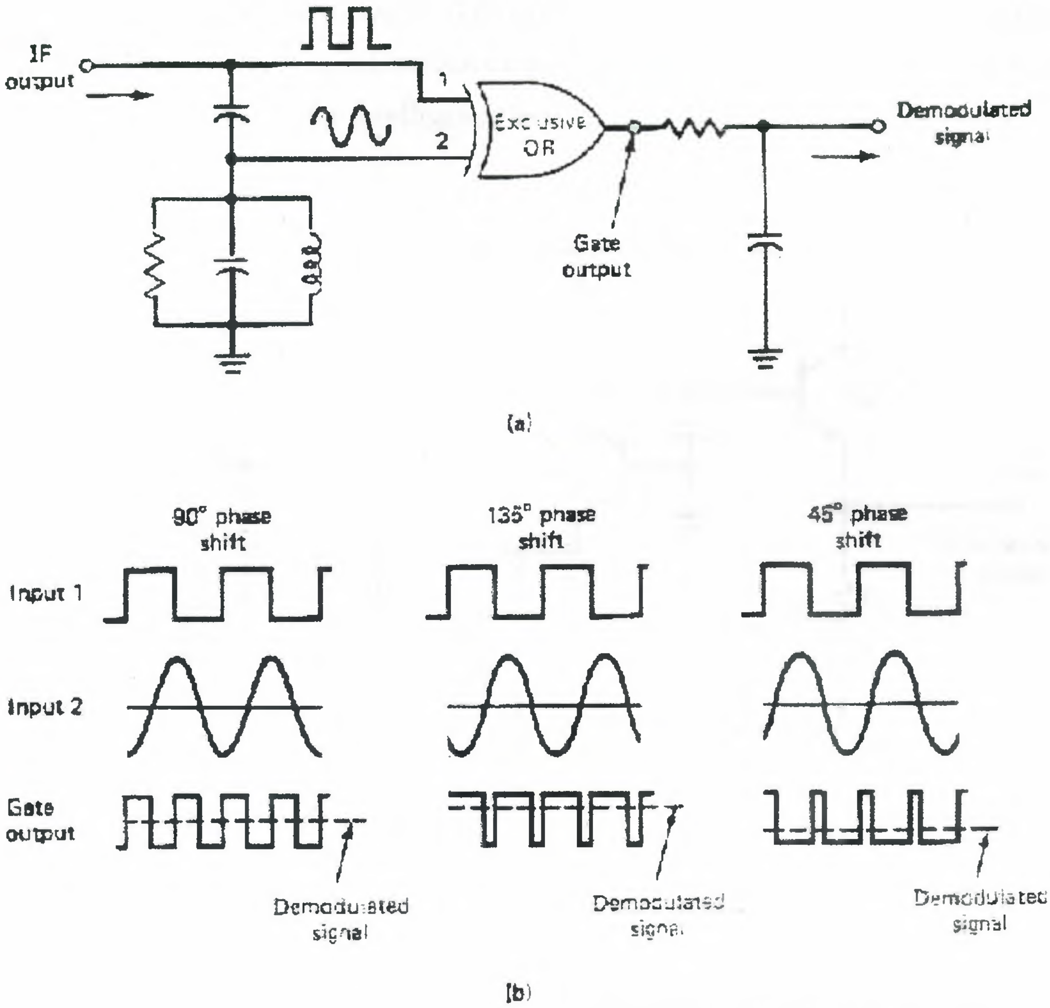

Figure 3.10 Quadrature detection.

The circuit in Fig. 3-10 (a) shows an FM quadrature detector using an Exclusive-OR gate. The limited IF output is applied directly to one input and the phase-shifted signal to the other. Notice that this circuit uses the limited signal that has not been changed back to a sine wave. The L, C, and R values used at the circuit's input are chosen to provide a 90° phase shift at the carrier frequency to the signal 2 input. The signal 2 input is a sine wave due to the LC circuit effects. The upward and downward frequency deviation of the FM signal results in a corresponding higher or lower phase shift. With one input to the gate shifted, the gate output will be a series of pulses with a width proportional to the phase difference. The low-pass RC filter at the gate output "sums" the output, giving an average value which is the intelligence signal. The gate output for three different phase conditions is shown at Fig. 3.10(b). The RC circuit output level for each case is shown with dashed lines. This corresponds to the intelligence level at those particular conditions.

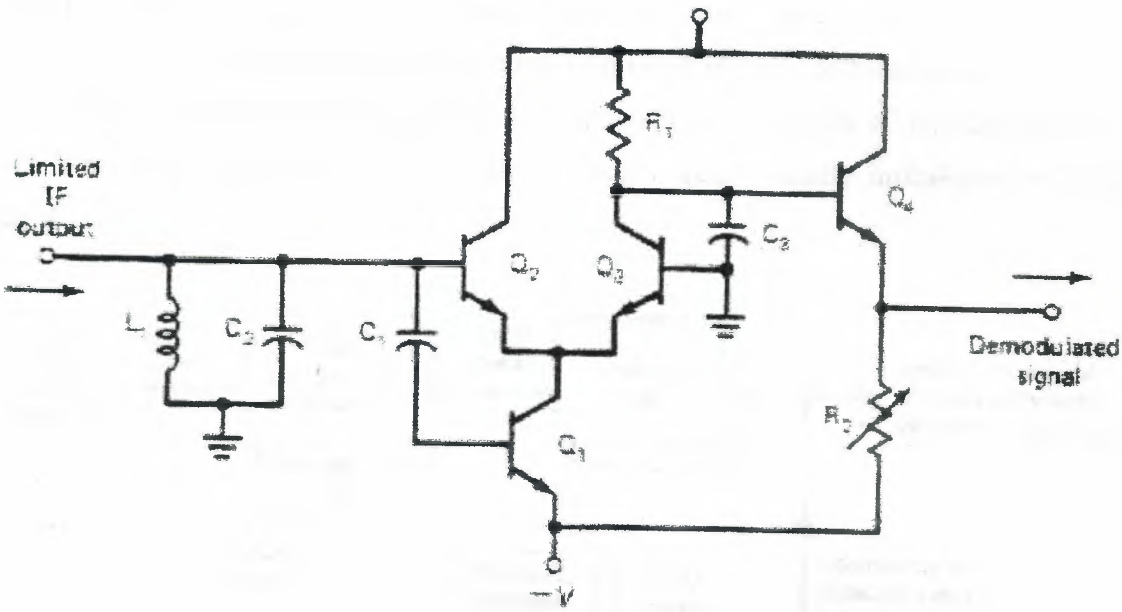

Figure 3.11 Analog quadrature detector.

An analog quadrature detector is possible using a differential amplifier configuration as shown in Fig. 3.11. A limited FM signal switches the transistor current source (Q_1) of the differential pair $Q_2 + Q_3$. L_1 and C_2 should be resonant at the IF frequency. The L_1-C_2-C_1 combination causes the desired frequency-dependent phase shift between the two signals applied to Q_2 and Q_1. The conduction through Q_3 depends on the coincident phase relationships of these two signals. The pulses generated at Q_3 collector are "summed" by the $R_1 - C_3$ low-pass filter, and the resulting intelligence signal is taken at Q_4's emitter. R_2 is adjusted to yield the desired zero-volts output when an undeviating FM carrier is the circuit's input signal.

3.5 Phase-Locked Loop

The phase-locked loop (PLL) has become increasingly popular as a means of FM demodulation in recent years. It eliminates the need for the intricate coil adjustments of the previously discussed discriminators and has many other uses in the field of electronics. It is an example of an old idea, originated in 1932 that was given a new life by integrated circuit technology. Prior to its availability in a single IC package in 1970, its complexity in discrete circuitry form made it economically unfeasible for most applications.

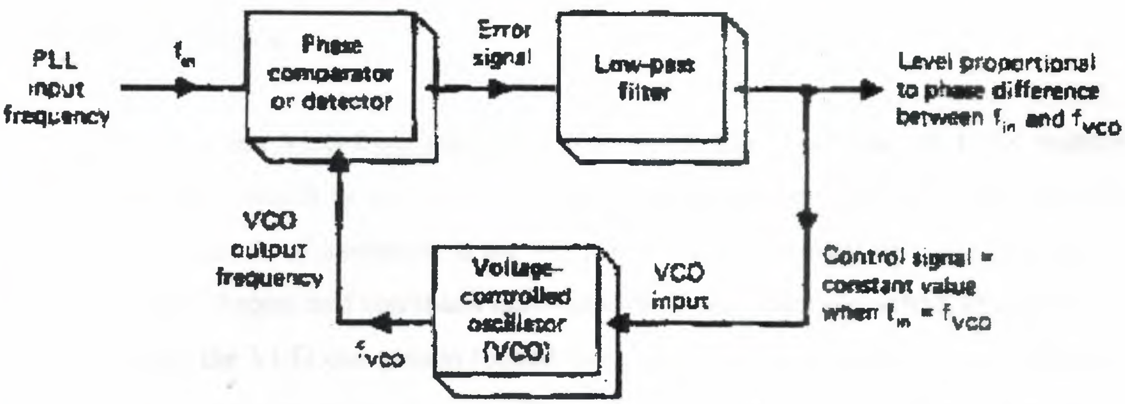

Figure 3.12 PLL block diagram.

The PLL is an electronic feedback control system as represented by the block diagram in Fig. 3.12. The input is to the phase comparator or phase detector, as it is also called. The VCO within the PLL generates the other signal applied to the comparator. The comparator compares the two signals and develops an output that is constant if the two input frequencies are identical. If the two are not identical, then the comparator output, when passed through the low-pass filter, is a level that is applied to the VCO input. This action "closes" the feedback loop since the level applied to the VCO input changes the VCO frequency in an attempt to make it exactly match the PLL input frequency. If the VCO frequency equals the input frequency the PLL has achieved "lock" and the control voltage will be constant for as long as the PLL input frequency remains constant.

3.5.1 PLL Capture and Lock

If the VCO starts to change frequency, it is in the capture state. It then continues to change frequency until its output is the same frequency as the input. At that point, the PLL is locked. The PLL has three possible stales of operation:

1. Free-running

2. Capture

3. Locked or tracking

If the input and VCO frequency are too far apart, the PLL free-runs at the nominal VCO frequency, which is determined by an external timing capacitor. This is not a normally used mode of operation. If the VCO and input frequency are close enough, the capture process begins and continues until the locked condition is reached, Once tracking (lock) begins, the VCO can remain locked over a wider input-frequency-range variation than was necessary to achieve capture. The tracking and capture range are a function of external resistors and/or capacitors selected by the user.

3.5.2 PLL FM Demodulator

If the PLL input is an FM signal, the low-pass filter output (VCO input) is the demodulated signal. The VCO input control signal (demodulated FM) causes the VCO output to match the FM signal applied to the PLL (comparator input). If the FM carrier (center) frequency drifts because of local oscillator drift, the PLL readjusts itself and no realignment is necessary. In a conventional FM discriminator, any shift in the FM carrier frequency results in a distorted output, since the LC detector circuits are then untuned. The PLL FM discriminator requires no tuned circuits and their associated adjustments and "adjusts" itself to any carrier frequency drifts caused by LO or transmitted carrier drift. In addition, the PLL normally has large amounts of internal amplification, which allows the input signal to vary from the micro volt region up to several volts, Since the phase comparator responds only to phase changes and not to amplitudes, the PLL is seen to provide a limiting function of extremely wide range. The use of PLL FM detectors is widespread in current designs.

3.6 Stereo Demodulation

FM stereo receivers are identical to standard receivers up to the discriminator output. At this point, however, the discriminator output contains the 30 Hz to 15 kHz (L + R) signal and the 19-kHz subcarrier and the 23 to 53 kHz (L - R) signal. If a non stereo receiver is tuned to a stereo station, its discriminator output may contain the additional frequencies, but even the 19-kHz subcarrier is above the normal audible range, and its audio amplifiers and speaker would probably not pass it anyway. Thus, the non stereo receiver reproduces the 30 Hz to 15 kHz (L + R) signal (a full monophonic broadcast) and is not affected by the other frequencies. This effect is illustrated in Fig. 3.14.

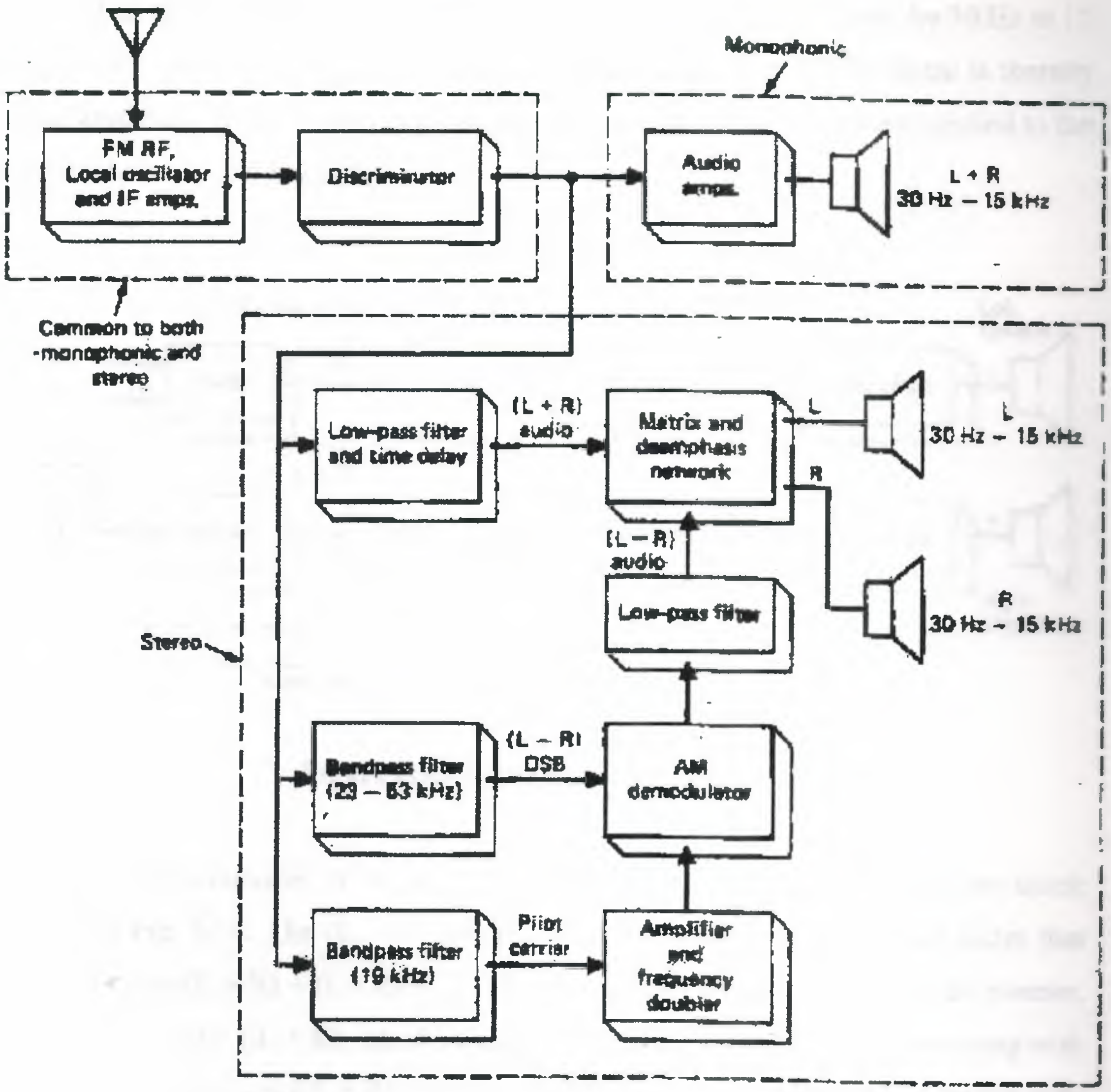

Figure 3.14 Monophonic and stereo receivers.

The stereo receiver block diagram becomes more complex after the discriminator. At this point, the three signals are separated by filtering action. The (L + R) signal is obtained through a low-pass filter and given a delay so that it reaches the matrix network in step with the (L - R) signal. A 23- to 53-kHz bandpass filter "selects" the (L - R) double-side band signal. A 19-kHz bandpass filter takes the pilot carrier and is multiplied by 2 to 38 kHz. which is the precise carrier frequency of the DSB suppressed carrier 23 to 53 kHz (L - R) signal. Combining the 38-kHz and (L - R) signals through the nonlinear device of an AM detector generates sum and difference outputs of which the 30 Hz to 15 kHz (L - R) components are selected by a low-pass filter. The (L - R) signal is thereby retranslated back down to the audio range and it and the (L + R) signal are applied to the matrix network for further processing.

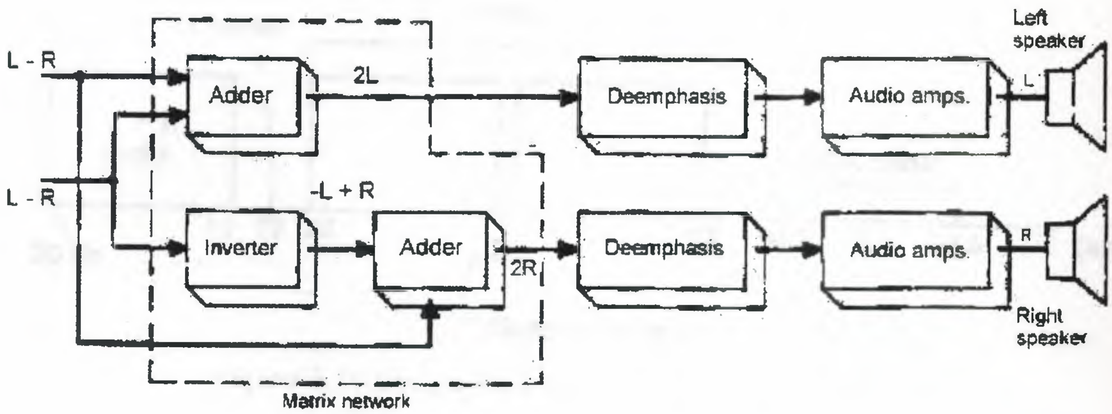

Figure 3.15 Stereo signal processing.

Figure 3.15 illustrates the matrix function and completes the stereo receiver block diagram of Fig. 3.14. The (L + R) and (L - R) signals are combined in an adder that cancels R since (L + R) + (L - R) = 2L. The (L - R) signal is also applied to an inverter, providing -(L - R) = (-L + R). which is subsequently applied to another adder along with (L + R), which produces (-L + R) + (L + R) = 2R. The two individual signals for the right and left channels are then deemphasized and individually amplified to then own speaker. The process of FM stereo is ingenious in its relative simplicity and effectiveness in providing complete compatibility and doubling the amount of transmitted information through the use of multiplexing.

3.6.1 SCA Decoder

The FCC has also authorized FM stations to broadcast an additional signal on their carrier. It may be a voice communication or other signal for any non broadcast-type use. It is often used to transmit music programming which is usually commercial-free but paid or by subscription of department stores supermarkets, and the like. It is termed the subsidiary communication authorization (SCA). It is frequency-multiplexed on the FM modulating signal, usually with a 67-kHz carrier and ±7.5-kHz (narrowband) deviation, as shown in Fig, 3.16.

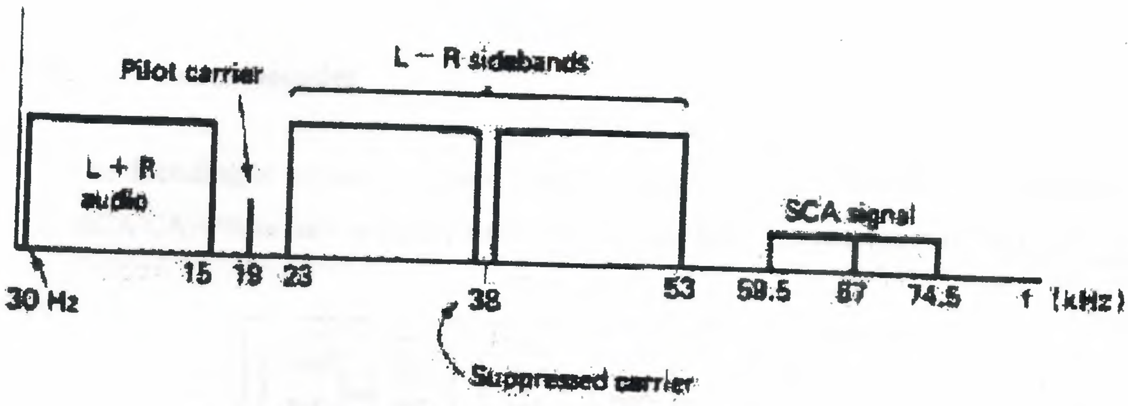

Figure 3.16 Composite stereo and SCA modulating signal.

An SCA decoder circuit using the 565 PLL is provided in Fig. 3.17.

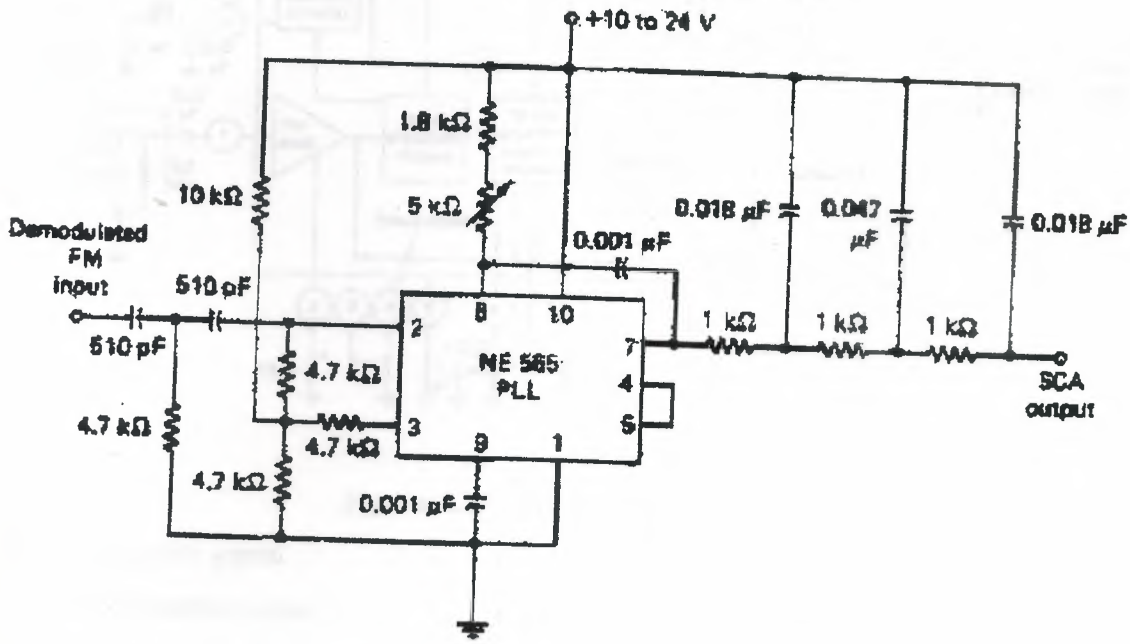

Figure 3.17 SCA PLL decoder

A resistive voltage divider is used to establish a bias voltage for the input (pins 2 and 3), The demodulated FM signal is fed to the input through a two-stage high-pass filter (510 pF, 4.7 kΩ, 510 pF, 4.7 kΩ), both to allow capacitive coupling and to attenuate the stronger level of the stereo signals. The PLL is tuned to about 67 kHz, with the 0,001-μF capacitor from pin 9 to ground and the 5-kΩ potentiometer providing fine adjustment. The demodulated output at pin 7 is fed through a three-stage low-pass filter to provide deemphasis and attenuate the high-frequency noise which often accompanies SCA transmission.

3.6.2 LIC Stereo Decoder

The decoding of the stereo signals is normally accomplished via special function ICs. The RCA CA3090 is such a device with a functional block diagram provided in Fig. 3.18.

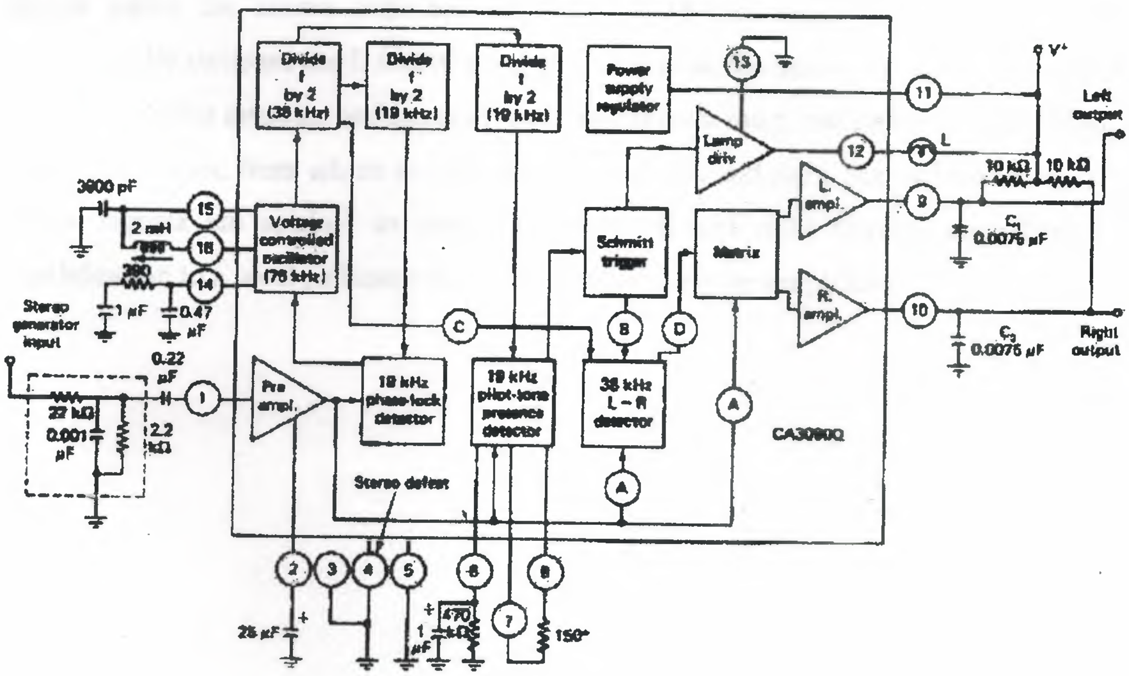

Figure 3.18 CA3090 stereo decoder.

(A) Composite signal
(B) Stereo enable signal
(C) Stereo gating signal
(D) Difference signal

L = 12V, 14mA lamp
C_1 , C_2 provide deemphasis
Determines sensitivity of pilot-tone presence detector

The input signal from the detector is amplified by a low-distortion preamplifier and simultaneously applied to both the 19- and 38-kHz synchronous detectors . A 76-kHz signal, generated by a local voltage-controlled oscillator (VCO), is counted down by two frequency dividers to a 38-kHz signal and to two 19-kHz signals in phase quadrature. The 19-kHz pilot tone supplied by the FM detector is compared to the locally generated 19 kHz signal in the synchronous detector. The resultant signal controls the voltage controlled oscillator so that it produces an output signal to phase-lock the stereo decoder with the pilot tone. A second synchronous detector compares the locally generated 19 kHz signal with the 19-kHz pilot tone. If the pilot tone exceeds an externally adjustable threshold voltage, a Schmitt trigger circuit is energized. The signal from the Schmitt trigger lights the stereo indicator, enables the 38-kHz synchronous detector, and automatically switches the CA3090 from monaural to stereo operation. The output signal from the 38-kHz detector and the composite signal from the preamplifier are applied to a matrixing circuit, from which emerge the resultant left and right channel audio signals. These signals are applied to their respective left and right channel amplifiers for amplification to a level sufficient to drive most audio power amplifiers.

CONCLUSION

In this project, I have tried to analyze Frequency Modulation broadcasting relating to transmission and reception. The factors effecting the transmission and reception and their appropriate solutions. The noise effecting the frequency modulated signal was removed by a limitter circuit and through the use of detector circuits that are in sensitive to the amplitude changes. The inherent ability of FM to minimize the effect of the undesired signals in the transmission also applies to the reception of an undesired station operating at the same frequency at the desired station and was observed to be the capture effect. The preemphasis, which involves increasing the relative strength of the higher frequency components of the audio signal before it is fed to the modulator, was also concluded. Bessel functions were used for the solution of frequency components of a FM wave.

The zero carrier condition was applied for determining the deviation produced in an FM modulator. It has been concluded that broadcast AM receivers normally operate quite satisfactorily without any RF amplifiers, whereas it is very rare case with FM receivers except for frequencies in excess of 1GHz. Phase locked loop was founded to eliminate the need for the intricate coil adjustments of the discriminator.

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