Wide-Band Frequency Modulation

- Spectral analysis of the wide-band FM wave

$$s(t) = A_c \cos[2\pi f_c t + \beta \sin(2\pi f_m t)]$$

or

$$s(t) = \Re\left[A_c \exp[j2\pi f_c t + j\beta \sin(2\pi f_m t)]\right] = \Re[\tilde{s}(t) \exp(j2\pi f_c t)]$$

where $\tilde{s}(t) = A_c \exp\left[j\beta \sin(2\pi f_m t)\right]$ is called "complex envelope".

Note that the complex envelope is a periodic function of time with a fundamental frequency f_m which means

$$\tilde{s}(t) = \tilde{s}(t + kT_m) = \tilde{s}(t + \frac{k}{f_m})$$

where $T_m = 1/f_m$

- Then we can rewrite

$$\begin{aligned}\tilde{s}(t) &= \tilde{s}(t + k/f_m) \\ &= A_c \exp[j\beta \sin(2\pi f_m (t + k/f_m))] \\ &= A_c \exp[j\beta \sin(2\pi f_m t + 2k\pi)] \\ &= A_c \exp[j\beta \sin(2\pi f_m t)]\end{aligned}$$

- Fourier series form

$$\tilde{s}(t) = \sum_{n=-\infty}^{\infty} c_n \exp(j2\pi n f_m t)$$

where

$$\begin{aligned}c_n &= f_m \int_{-1/(2f_m)}^{1/(2f_m)} \tilde{s}(t) \exp(-j2\pi n f_m t)\, dt \\ &= f_m A_c \int_{-1/(2f_m)}^{1/(2f_m)} \exp[j\beta \sin(2\pi f_m t) - j2\pi n f_m t]\, dt\end{aligned}$$

- Define the new variable: $x = 2\pi f_m t$

  Then we can rewrite

$$c_n = \frac{A_c}{2\pi} \int_{-\pi}^{\pi} \exp[j(\beta \sin x - nx)]\, dx$$

- nth order Bessel function of the first kind and argument β

$$J_n(\beta) = \frac{1}{2\pi} \int_{-\pi}^{\pi} \exp[j(\beta \sin x - nx)]\, dx$$

- Accordingly

$$c_n = A_c J_n(\beta)$$

  which gives

$$\tilde{s}(t) = A_c \sum_{n=-\infty}^{\infty} J_n(\beta) \exp(j2\pi n f_m t)$$

- Then the FM wave can be written as

$$\begin{aligned} s(t) &= \Re[\tilde{s}(t)\exp(j2\pi f_c t)] \\ &= \Re\left[A_c \sum_{n=-\infty}^{\infty} J_n(\beta)\exp[j2\pi n(f_c + f_m)t]\right] \\ &= A_c \sum_{n=-\infty}^{\infty} J_n(\beta)\cos[2\pi(f_c + nf_m)t] \end{aligned}$$

- Fourier transform

$$S(f) = \frac{A_c}{2} \sum_{n=-\infty}^{\infty} J_n(\beta)\left[\delta(f - f_c - nf_m) + \delta(f + f_c + nf_m)\right]$$

which shows that the spectrum consists of an infinite number of delta functions spaced at $f = f_c \pm nf_m$ for $n = 0, +1, +2, ...$

Properties of Single-Tone FM for Arbitrary Modulation Index β

1. For different values of n

$$
\begin{aligned}
J_n(\beta) &= J_{-n}(\beta), && \text{for } n \text{ even} \\
J_n(\beta) &= -J_{-n}(\beta), && \text{for } n \text{ odd}
\end{aligned}
$$

2. For small value of β

$$
\begin{aligned}
J_0(\beta) &\approx 1, \\
J_1(\beta) &\approx \frac{\beta}{2} \\
J_n(\beta) &\approx 0, \qquad n > 2
\end{aligned}
$$

6. The equality holds exactly for arbitrary β

$$
\sum_{n=-\infty}^{\infty} J_n^2(\beta) = 1
$$

1. The spectrum of an FM wave contains a carrier component and and an infinite set of side frequencies located symmetrically on either side of the carrier at frequency separations of $f_m,\ 2f_m,\ 3f_m$...

2. The FM wave is effectively composed of a carrier and a single pair of side-frequencies at $f_c \pm f_m$.

3. The amplitude of the carrier component of an FM wave is dependent on the modulation index β .The average power of such as signal developed across a 1-ohm resistor is also constant:

$$P_{\mathrm{av}} = \frac{1}{2} A_c^2$$

The average power of an FM wave may also be determined from

$$P_{\mathrm{av}} = \frac{1}{2} A_c^2 \sum_{n=-\infty}^{\infty} J_n^2(\beta)$$

Case 1

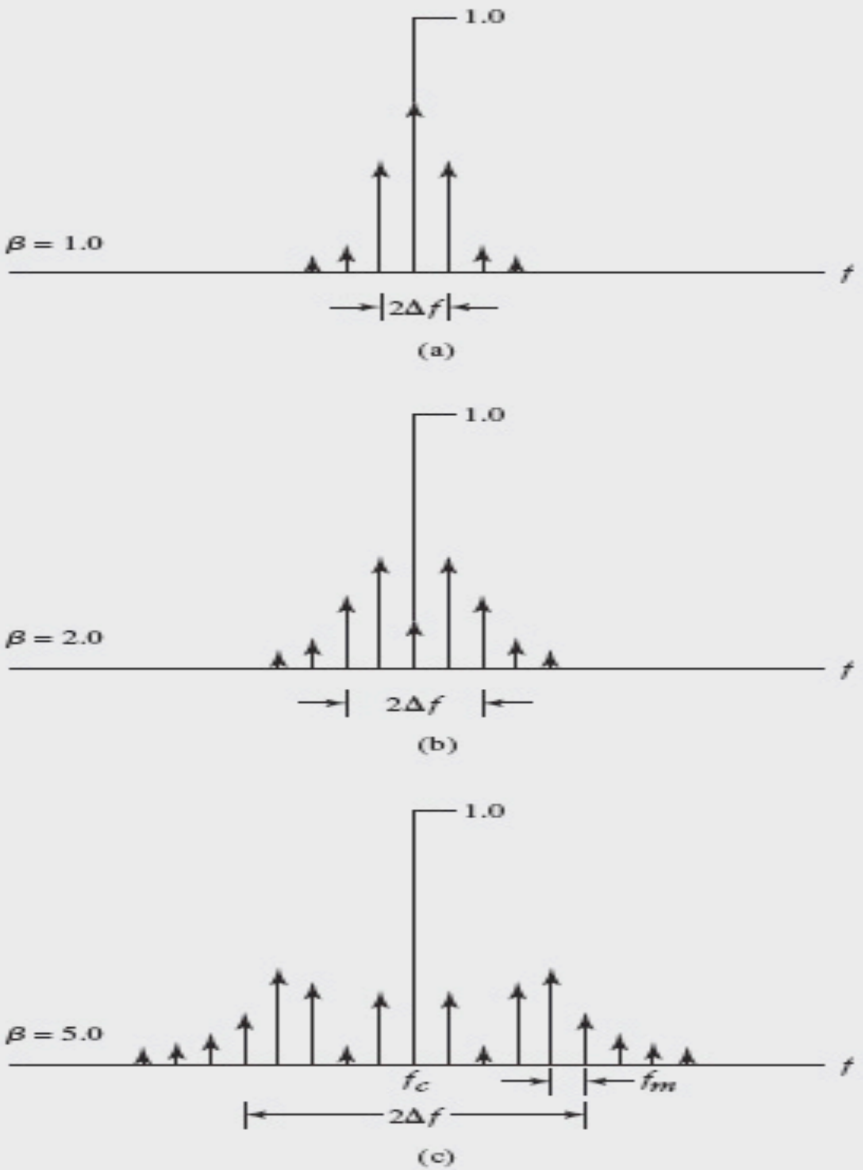


FIGURE 4.7 Discrete amplitude spectra of an FM wave, normalized with respect to the unmodulated carrier amplitude, for the case of sinusoidal modulation of fixed frequency and varying amplitude. Only the spectra for positive frequencies are shown.

[Ref: Haykin & Moher, Textbook]

FM-demodulation using phase Locked Loop:- (PLL)

phase Locked Loop (PLL) is a negative feedback system that consists of three major components

⟨i⟩ A Multiplier used as a phase detector (or) phase comparator.

⟨ii⟩ A- voltage Controlled oscillator ⟨VCO⟩

⟨iii⟩ A- Loop filter, which is a Low pass filter ⟨LPF⟩.

The Block diagram of PLL is shown in Fig.1.

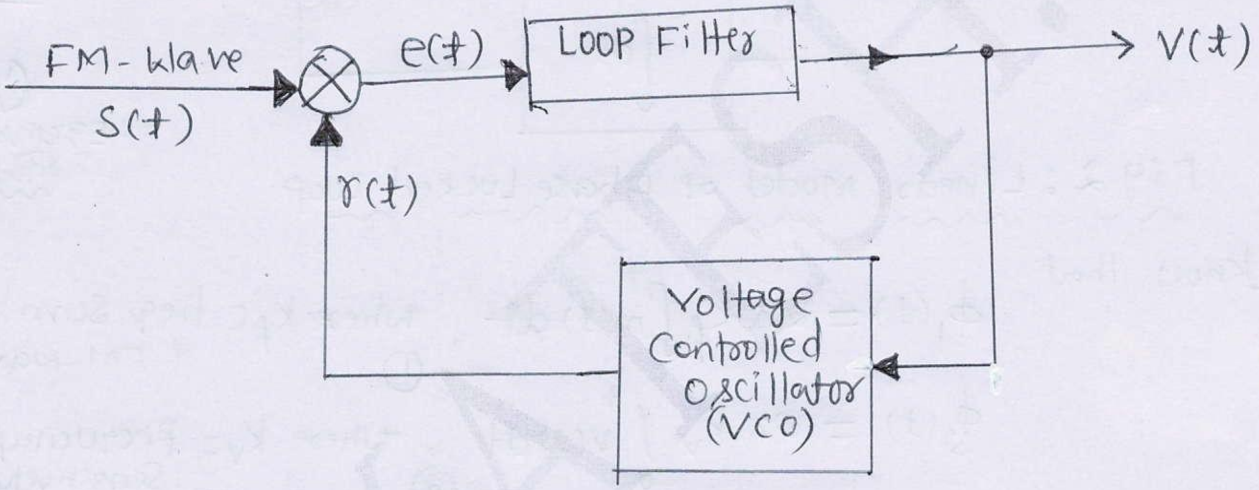


Fig.1: Block diagram of PLL

↳ The VCO output is defined as

$$r(t) = A_v \cos(2\pi f_c t + \phi_2(t)) \quad \text{—} ①$$

Where $\phi_2(t) = 2\pi K_v \int_0^t v(t)\,dt$.

↳ Then, the incoming signal (FM) (S(t)) and the VCO output r(t) are applied to the multiplier, then it gives error signal,

$$e(t) = r(t) \cdot S(t) \quad \text{—} ②$$

Where $S(t) = A_c \sin[2\pi f_c t + \phi_1(t)]$ —③

Where $\phi_1(t) = 2\pi K_f \int_0^t m(t)\,dt$. —④

⟨i⟩ Linear Model of phase-Locked-Loop (Linear-PLL):-

V.T.U Q.P. 8 Mark

The phase locked loop (PLL) is said to be in phase-Lock, when the phase error $\boxed{\phi_e(t)=0}$

The Linear model of PLL for the demodulation of FM-signal is shown in Figure.2.

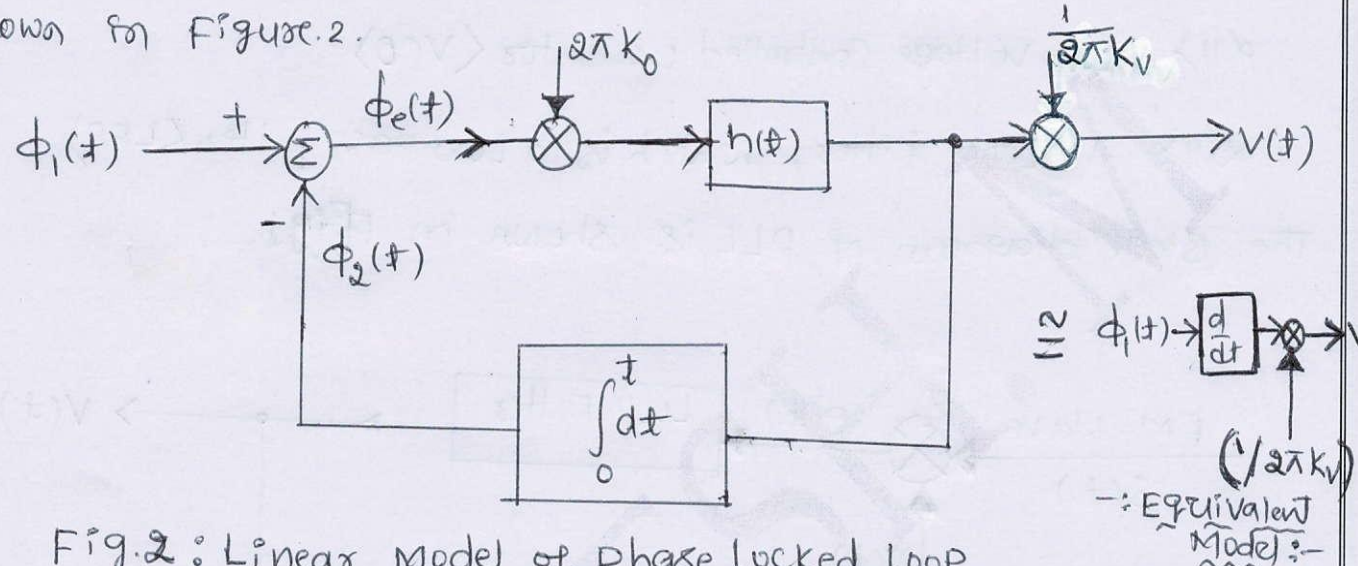


Fig.2: Linear Model of Phase Locked Loop

We know that,

$$\phi_1(t) = 2\pi K_f \int_0^t m(t)\,dt \quad , \text{ where } K_f = \text{Freq. Sensitivity of FM-wave.} \quad \text{—} (1)$$

$$\phi_2(t) = 2\pi K_V \int_0^t V(t)\,dt \quad , \text{ where } K_V = \text{Frequency Sensitivity Constant of VCO} \quad \text{—} (2)$$

From fig.2,

$$\phi_e(t) = \phi_1(t) - \phi_2(t) \quad \text{—} (3)$$

W.K.T. for phase-lock mode : $\phi_e(t) = 0$ (Assuming Small error ≃ 0)

∴ Equation (3) ⟹ $0 = \phi_1(t) - \phi_2(t)$

$$\therefore \phi_1(t) = \phi_2(t)$$

Using equations (1) & (2) we get

$$2\pi K_f \int_0^t m(t)\,dt = 2\pi K_V \int_0^t v(t)\,dt$$

$$K_f \int_0^t m(t)\,dt = K_v \int_0^t v(t)\,dt \quad \text{—} (4)$$

Differentiating both sides of equation (4), we get

$$K_f\, m(t) = K_v\, V(t).$$

$$\therefore \boxed{V(t) = \frac{K_f}{K_v} \cdot m(t) = K\,m(t)} \quad ; \text{ where } K = \frac{K_f}{K_v}$$

$$\text{i.e., } \boxed{V(t) \propto m(t)}$$

Thus, the output V(t) of the low pass-loop filter [h(t)] is proportional to the original modulating signal.

i.e., The message signal present in FM-modulated wave s(t) is recovered and it is produced at the output of loop filter.

—*—*—*—

**

3.9: Non-Linear effects in FM-wave :- <VIU Q.P>

Q) Write a short note on Non-linear effects in FM-system.

↳ Non-linea[redacted]ffects can be of two-[redacted]es

(i) Stron[redacted] (ii) Weak.

* Non-linearit[redacted] said to be [redacted]g, if it is intentionally introduced int[redacted] cir[redacted] in a controlled manner.

  Ex: square law [redacted].

* Non linearity is [redacted]be weak, when it is inherently present in th[redacted]rcuit.

The effect [redacted] non-liniariti[redacted]ll limit m(t) levels in the system[redacted]earity is present.

In FM-generation system, weak non-[redacted]earity is present. the effect of weak non-linearity in FM-systems can be by considering the input and output relation of the memoryless-non linear device used in the frequency multiplier.

8) The techniques such as equalization, especially adaptive versions, are easier to implement with digital transmission techniques.

* COMPARISON BETWEEN ANALOG & DIGITAL COMM SYSTEMS:

S.NO	Parameter	Analog system	Digital system
1)	Bandwidth	Less	More
2)	Error correction and Detection	Not possible	Possible
3)	Immune to Noise	Less	More
4)	System Complexity	Less	More
5)	System Cost	More	Less
6)	Quality of reconstruction	Good	Very Good
7)	Synchronisation	Not required	required
8)	Privacy and security to data	Not possible	possible
9)	Flexibility and Liability	Less	More
10)	Power required	More	Less
11)	Implementation	Difficult	Easy
12)	Programming	Not possible	Possible

rectangular pulse of duration Δt and amplitude $g(nT_s)/\Delta t$; the smaller we make Δt, the better will be the approximation.

The ideal sampled signal $g_\delta(t)$ has a mathematical form similar to that of the Fourier transform of a periodic signal. This is readily established by comparing Eq. (7.1) for $g_\delta(t)$ with the Fourier transform of a periodic signal given in Eq. (2.88). This correspondence suggests that we may determine the Fourier transform of the ideal sampled signal $g_\delta(t)$ by applying the duality property of the Fourier transform to the transform pair of Eq. (2.88). By so doing, and using the fact that a delta function is an even function of time, we get the result:

$$g_\delta(t) \rightleftharpoons f_s \sum_{m=-\infty}^{\infty} G(f - mf_s) \tag{7.2}$$

where $G(f)$ is the Fourier transform of the original signal $g(t)$, and f_s is the sampling rate. Equation (7.2) states that *the process of uniformly sampling a continuous-time signal of finite energy results in a periodic spectrum with a period equal to the sampling rate.*

Another useful expression for the Fourier transform of the ideal sampled signal $g_\delta(t)$ may be obtained by taking the Fourier transform of both sides of Eq. (7.1) and noting that the Fourier transform of the delta function $\delta(t - nT_s)$ is equal to $\exp(-j2\pi f T_s)$. Let $G_\delta(f)$ denote the Fourier transform of $g_\delta(t)$. We may therefore write

$$G_\delta(f) = \sum_{n=-\infty}^{\infty} g(nT_s)\exp(-j2\pi n f T_s) \tag{7.3}$$

This relation is called the *discrete-time Fourier transform* and was briefly discussed in Chapter 2. It may be viewed as a complex Fourier series representation of the periodic frequency function $G_\delta(f)$, with the sequence of samples $\{g(nT_s)\}$ defining the coefficients of the expansion.

The relations, as derived here, apply to any continuous-time signal $g(t)$ of finite energy and infinite duration. Suppose, however, that the signal $g(t)$ is *strictly band-limited*, with no frequency components higher than W Hertz. That is, the Fourier transform $G(f)$ of the signal $g(t)$ has the property that $G(f)$ is zero for $|f| \geq W$, as illustrated in Figure 7.2*a*; the shape of the spectrum shown in this figure is intended for the purpose of illustration only. Suppose also that we choose the sampling period $T_s = 1/2W$. Then the corresponding spectrum $G_\delta(f)$ of the sampled signal $g_\delta(t)$ is as shown in Figure 7.2*b*. Putting $T_s = 1/2W$ in Eq. (7.3) yields

$$G_\delta(f) = \sum_{n=-\infty}^{\infty} g\left(\frac{n}{2W}\right)\exp\left(-\frac{j\pi n f}{W}\right) \tag{7.4}$$

From Eq. (7.2), we readily see that the Fourier transform of $g_\delta(t)$ may also be expressed as

$$G_\delta(f) = f_s G(f) + f_s \sum_{\substack{m=-\infty \\ m\neq 0}}^{\infty} G(f - mf_s) \tag{7.5}$$

Hence, under the following two conditions:

1. $G(f) = 0$ for $|f| \geq W$

2. $f_s = 2W$

we find from Eq. (7.5) that

$$G(f) = \frac{1}{2W} G_\delta(f), \qquad -W < f < W \tag{7.6}$$

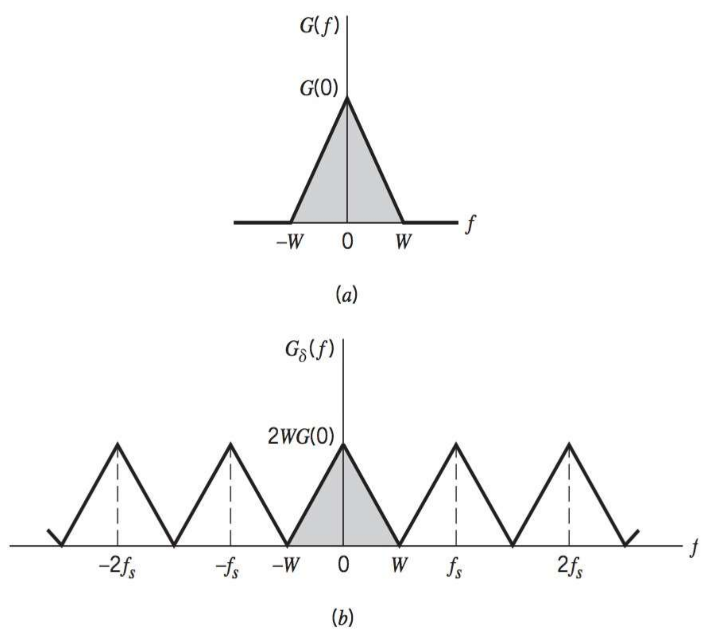


FIGURE 7.2 (a) Spectrum of a strictly band-limited signal $g(t)$. (*b*) Spectrum of sampled version of $g(t)$ for a sampling period $T_s = 1/2W$.

Substituting Eq. (7.4) in Eq. (7.6), we may also write

$$G(f) = \frac{1}{2W}\sum_{n=-\infty}^{\infty} g\left(\frac{n}{2W}\right)\exp\left(-\frac{j\pi nf}{W}\right), \qquad -W<f<W \tag{7.7}$$

Therefore, if the sample values $g(n/2W)$ of a signal $g(t)$ are specified for all time, then the Fourier transform $G(f)$ of the signal is uniquely determined by using the discrete-time Fourier transform of Eq. (7.7). Because $g(t)$ is related to $G(f)$ by the inverse Fourier transform, it follows that the signal $g(t)$ is itself uniquely determined by the sample values $g(n/2W)$ for $-\infty<n<\infty$. In other words, the sequence $\{g(n/2W)\}$ has all the information contained in $g(t)$.

Consider next the problem of reconstructing the signal $g(t)$ from the sequence of sample values $[g(n/2W)]$. Substituting Eq. (7.7) in the formula for the inverse Fourier transform defining $g(t)$ in terms of $G(f)$, we get

$$\begin{aligned} g(t) &= \int_{-\infty}^{\infty} G(f)\exp(j2\pi ft)\,df \\ &= \int_{-W}^{W} \frac{1}{2W}\sum_{n=-\infty}^{\infty} g\left(\frac{n}{2W}\right)\exp\left(-\frac{j\pi nf}{W}\right)\exp(j2\pi ft)\,df \end{aligned}$$

Interchanging the order of summation and integration:

$$g(t) = \sum_{n=-\infty}^{\infty} g\left(\frac{n}{2W}\right)\frac{1}{2W}\int_{-W}^{W}\exp\left[j2\pi f\left(t-\frac{n}{2W}\right)\right]df \tag{7.8}$$

The integral term in Eq. (7.8) is readily evaluated, yielding the final result

$$\begin{aligned} g(t) &= \sum_{n=-\infty}^{\infty} g\left(\frac{n}{2W}\right)\frac{\sin(2\pi Wt-n\pi)}{(2\pi Wt-n\pi)} \\ &= \sum_{n=-\infty}^{\infty} g\left(\frac{n}{2W}\right)\operatorname{sinc}(2Wt-n), \qquad -\infty<t<\infty \end{aligned} \tag{7.9}$$

Equation (7.9) provides an *interpolation formula* for reconstructing the original signal $g(t)$ from the sequence of sample values $\{g(n/2W)\}$, with the sinc function $\operatorname{sinc}(2Wt)$ playing the role of an *interpolation function*. Each sample is multiplied by a delayed version of the interpolation function, and all the resulting waveforms are added to obtain $g(t)$. Looking at Eq. (7.9) in another way, it represents the convolution (or filtering) of the impulse train $g_\delta(t)$ given by Eq. (7.1) with the impulse response $\operatorname{sinc}(2Wt)$. Consequently, any impulse response that plays the same role as $\operatorname{sinc}(2Wt)$ is also referred to as a *reconstruction filter.*

We may now state the *sampling theorem* for strictly band-limited signals of finite energy in two equivalent parts:

1. *A band-limited signal of finite energy, which only has frequency components less than W Hertz, is completely described by specifying the values of the signal at instants of time separated by 1/2W seconds.*
2. *A band-limited signal of finite energy, which only has frequency components less than W Hertz, may be completely recovered from a knowledge of its samples taken at the rate of 2W samples per second.*

The sampling rate of $2W$ samples per second, for a signal bandwidth of W Hertz, is called the *Nyquist rate*; its reciprocal $1/2W$ (measured in seconds) is called the *Nyquist interval.*

The Whittakers, Father and Son

The exact origin of the sampling theorem has an intriguing history of its own. The earliest and most highly cited paper is that of E. T. Whittaker, published in 1915. In that paper, Whittaker described an idea that he termed the *cardinal function*, which was subsequently, in 1929 renamed the *cardinal series* by his son, J. M. Whittaker. In his 1915 paper, the senior Whittaker showed (among other findings) that if a function of time is band-limited, then the cardinal series is applicable to that function.

The *sampling theorem*, under that very name, is mentioned (perhaps for the first time) in Shannon's 1949 paper on information theory. For the derivation of the theorem, the reader is referred to another Shannon paper written in 1949 on "Communication in the presence of noise." In this latter paper, Shannon does make reference to a book by J. M. Whittaker on *Interpolation Function Theory*, published in 1935.

For a more detailed account of the history of the sampling theorem, see Chapter 1 of the book by Marks (1991), which, interestingly enough, is entitled *Introduction to Shannon Sampling and Interpolation Theory*.

The derivation of the sampling theorem, as described herein, is based on the assumption that the signal $g(t)$ is strictly band limited. In practice, however, an information-bearing signal is *not* strictly band limited, with the result that some degree of undersampling is encountered. Consequently, some *aliasing* is produced by the sampling process. Aliasing refers to the phenomenon of a high frequency component in the spectrum of the signal seemingly taking on the identity of a lower frequency in the spectrum of its sampled version, as illustrated in Figure 7.3. The aliased spectrum shown by the solid curve in Figure 7.3*b* pertains to an "undersampled" version of the message signal represented by the spectrum of Figure 7.3*a*. To combat the effects of aliasing in practice, we may use two corrective measures, as described here:

1. Prior to sampling, a low-pass *pre-alias filter* is used to attenuate those high-frequency components of the signal that are not essential to the information being conveyed by the signal.
2. The filtered signal is sampled at a rate slightly higher than the Nyquist rate.

The use of a sampling rate higher than the Nyquist rate also has the beneficial effect of easing the design of the *reconstruction filter* used to recover the original signal from its sampled version. Consider the example of a message signal that has been pre-alias (low-pass) filtered, resulting in the spectrum shown in Figure 7.4*a*. The corresponding spectrum of the instantaneously sampled version of the signal is shown in Figure 7.4*b*, assuming a sampling rate higher than the Nyquist rate. According to Figure 7.4*b*, we readily see that the design of the reconstruction filter may be specified as follows (see Figure 7.4*c*):

- The reconstruction filter is low-pass with a passband extending from $-W$ to W, which is itself determined by the pre-alias filter.
- The filter has a transition band extending (for positive frequencies) from W to $f_s - W$, where f_s is the sampling rate.

The fact that the reconstruction filter has a well-defined transition band means that it is physically realizable. This is to be compared to the implementation of the ideal reconstruction filter corresponding to sinc($2Wt$) that would be necessary if the signal was not oversampled.

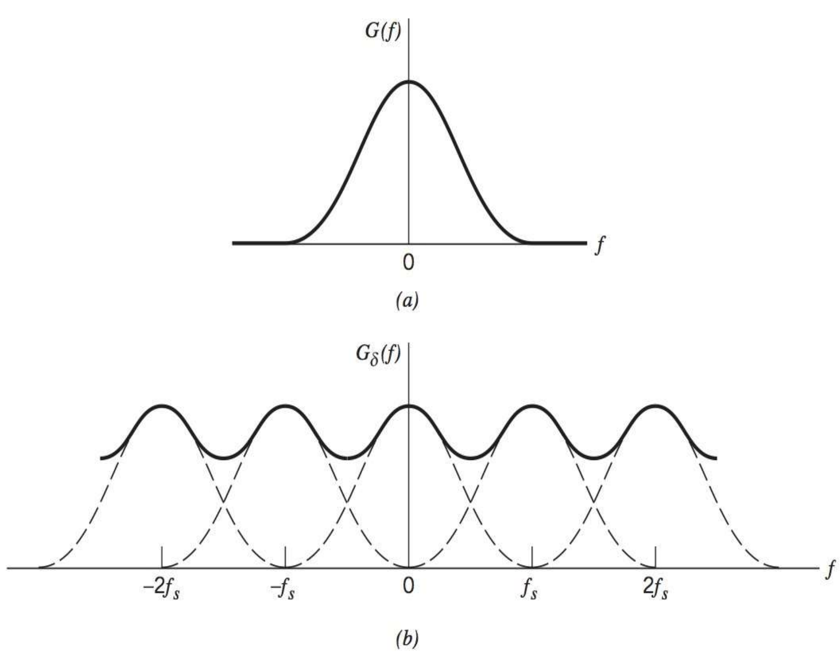


FIGURE 7.3 (*a*) Spectrum of a signal, (*b*) spectrum of an undersampled version of the signal exhibiting the aliasing phenomenon.

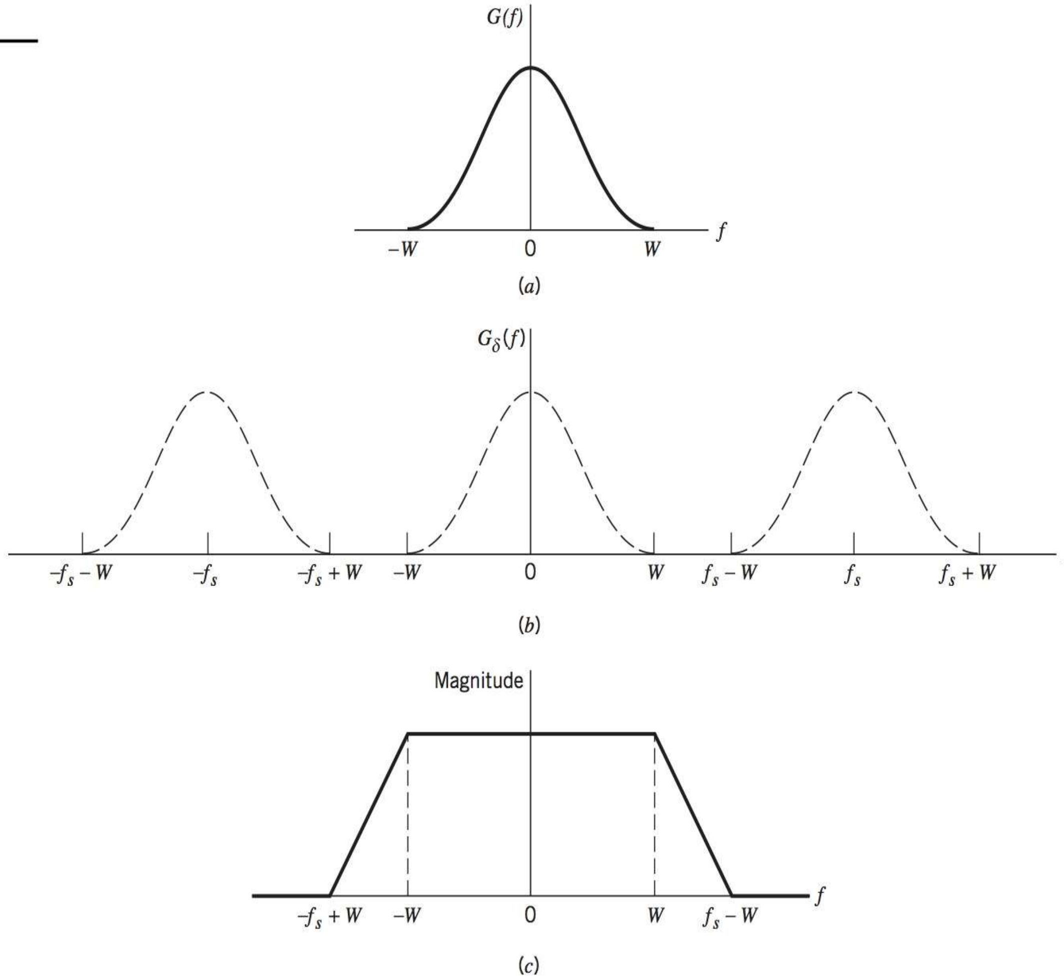


FIGURE 7.4 (*a*) Pre-alias filtered spectrum of an information-bearing signal. (*b*) Spectrum of instantaneously sampled version of the signal, assuming the use of a sampling rate greater than the Nyquist rate. (*c*) Amplitude response of reconstruction filter.

7.4 PULSE-AMPLITUDE MODULATION

Now that we understand the essence of the sampling process, we are ready to formally define pulse-amplitude modulation, which is the simplest and most basic form of analog pulse modulation. In *pulse-amplitude modulation* (PAM), *the amplitudes of regularly spaced pulses are varied in proportion to the corresponding sample values of a continuous message signal*; the pulses can be of a rectangular form or some other appropriate shape. Pulse-amplitude modulation as defined here is somewhat similar to natural sampling, where the message signal is multiplied by a periodic train of rectangular pulses. However, in natural sampling the top of each modulated rectangular pulse varies with the message signal, whereas in PAM it is maintained flat; natural sampling is explored further in Problem 7.1.

The waveform of a PAM signal is illustrated in Figure 7.5. The dashed curve in this figure depicts the waveform of a message signal $m(t)$, and the sequence of amplitude-modulated rectangular pulses shown as solid lines represents the corresponding PAM signal $s(t)$. There are two operations involved in the generation of the PAM signal:

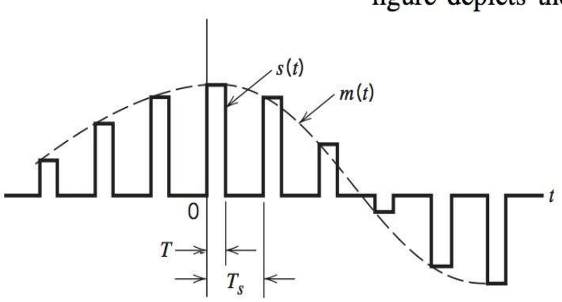


FIGURE 7.5 Flat-top samples.

1. *Instantaneous sampling* of the message signal $m(t)$ every T_s seconds, where the sampling rate $f_s = 1/T_s$ is chosen in accordance with the sampling theorem.
2. *Lengthening* the duration of each sample so obtained to some constant value T.

35. Generation of FM-Waves :-

There are two basic methods of generating FM-waves.

⟨i⟩ Direct Method.

⟨ii⟩ Indirect Method ⟨Armstrong Modulator⟩

*** IMP**

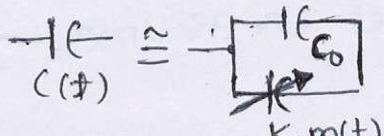
⟨i⟩ Generation of frequency modulated signal using DIRECT-METHOD :-

⟨Q⟩ Explain generation of frequency modulated signal using direct method. V.T.U June/July-2017 (5M)

↳ The Direct method uses a sinusoidal oscillator, with one of the reactive elements (example: Capacitive element) in the tank circuit of the oscillator being directly controlled by the message signal, m(t).

↳ In direct method of FM-signal generation, the instantaneous frequency of the carrier wave is varied directly in accordance with the message signal.

↳ Fig. 1, shows a Hartly oscillator in which the Capacitive Component of the tank circuit is,



$$C(t) = C_0 + K_c m(t) \quad \text{——} \quad (*)$$

where, C_0 = Total Capacitance in the absence of modulation.

K_c = Variable Capacitor Sensitivity to voltage change.

$m(t)$ = message signal = $A_m \cos(2\pi f_m t)$.

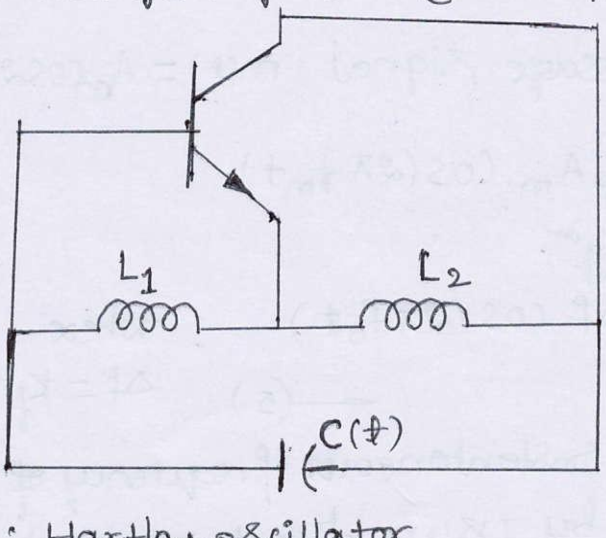


Fig.1: Hartley oscillator

The frequency of the Hartley oscillator is given by 20

$$f_i(t) = \frac{1}{2\pi\sqrt{(L_1+L_2)\,C(t)}} \quad ; \text{Where } C(t) = C_0 + K_c m(t)$$

$$\therefore \quad f_i(t) = \frac{1}{2\pi\sqrt{(L_1+L_2)\left[C_0 + K_c m(t)\right]}}$$

$$f_i(t) = \frac{1}{2\pi\sqrt{(L_1+L_2)\,C_0\left[1+\frac{K_c}{C_0}m(t)\right]}}$$

$$f_i(t) = \frac{f_0}{\sqrt{1+\frac{K_c}{C_0}m(t)}} \quad ; \text{where } f_0 = \frac{1}{2\pi\sqrt{(L_1+L_2)\,C_0}}$$

$$f_i(t) = f_0\left(1+\frac{K_c}{C_0}\cdot m(t)\right)^{-1/2} \quad \text{——(1)}$$

Using Binomial theorem, $(1+x)^{-1/2} = \left(1-\frac{x}{2}\right)$

$$\therefore \quad \left(1+\frac{K_c}{C_0}m(t)\right)^{-1/2} = \left(1-\frac{K_c\cdot m(t)}{2C_0}\right) \quad \text{——(2)}$$

Using equation (2) in (1) we get

$$f_i(t) = f_0\left(1-\frac{K_c\, m(t)}{2C_0}\right) \quad \text{——(3)}$$

Let us assume, $\frac{-K_c}{2C_0} = \frac{K_f}{f_0}$, where K_f = frequency sensitivity parameter.

$$\therefore \quad f_i(t) = f_0\left(1+\frac{K_f}{f_0}\cdot m(t)\right)$$

$$f_i(t) = f_0 + K_f m(t) \quad \text{——(4)}$$

for sinusoidal message signal, $m(t) = A_m \cos 2\pi f_m t$

$$\therefore \quad f_i(t) = f_0 + \underbrace{K_f A_m}_{\Downarrow}\cdot\cos(2\pi f_m t)$$

$$\therefore \quad f_i(t) = f_0 + \Delta f\cos(2\pi f_m t) \quad \text{——(5)}$$

; where $\Delta f = K_f A_m$ = Maximum frequency deviation.

Equation (5) gives, the instantaneous frequency of FM-wave generated by using direct method.

↳ Therefore, the direct method is stright forward to implement and is capable of providing large frequency deviation (Δf).

↳ One of the major limitation of the direct method is, "the carrier frequency is not obtained from a highly stable oscillator.

↳ To overcome, this limitation. a closed loop feedback system for the carrier frequency stabilization is used to provide frequency stabilized FM wave. This arrangement is shown in Fig. 2.

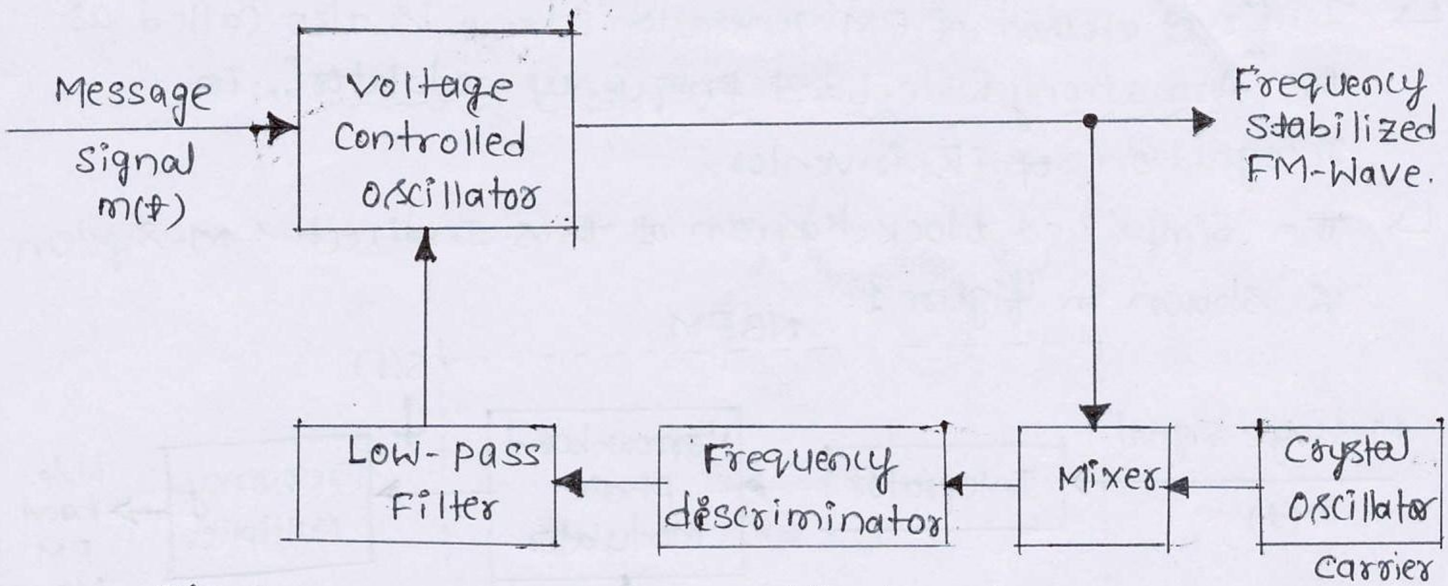


Fig. 2: Carrier frequency stabilization of Direct method FM-generation

↳ Fig. 2 consists of crystal oscillator, Mixer, Frequency descriminator, Low-pass filter and Voltage Controlled oscillator (VCO).

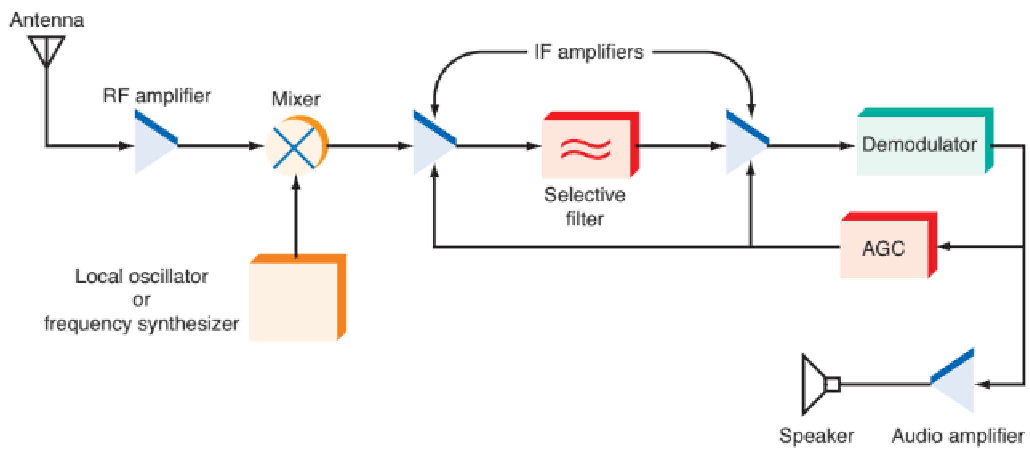
↳ This configuration provides

- Good frequency stability
- Required frequency deviation to generate WBFM.
- Constant proportionality between output frequency change to input voltage change.

∴ the required WBFM-wave is obtained.

Superheterodyne Receiver

One type of receiver that can provide the capability and performance required for the modern communication systems is the superheterodyne receiver. Superheterodyne receivers convert all incoming signals to a lower frequency, known as the intermediate frequency (IF), at which a single set of amplifiers and filters provides a fixed level of sensitivity and selectivity. Most of the gain and selectivity in a superheterodyne receiver are obtained in the IF amplifiers. The key circuit is the mixer, which acts as a simple amplitude modulator to produce sum and difference frequencies. The incoming signal is mixed with a local oscillator signal to obtain this conversion. The below figure shows a general block diagram of a superheterodyne receiver.



Superheterodyne Receiver

RF Amplifiers:

The antenna picks up the weak radio signal and feeds it to the RF amplifier, also called a low-noise amplifier (LNA). Because RF amplifiers provide some initial gain and selectivity, they are sometimes referred to as preselectors. Tuned circuits help select the desired signal or the frequency range in which the signal resides. The tuned circuits in fixed-tuned receivers can be given a very high Q, excellent selectivity can be obtained. For receivers that must be tuned over a broad range of frequencies, selectivity is mostly compromised. The tuned circuits must resonate over a wide frequency range. Therefore, the Q, bandwidth, and selectivity of the amplifier change with frequency.

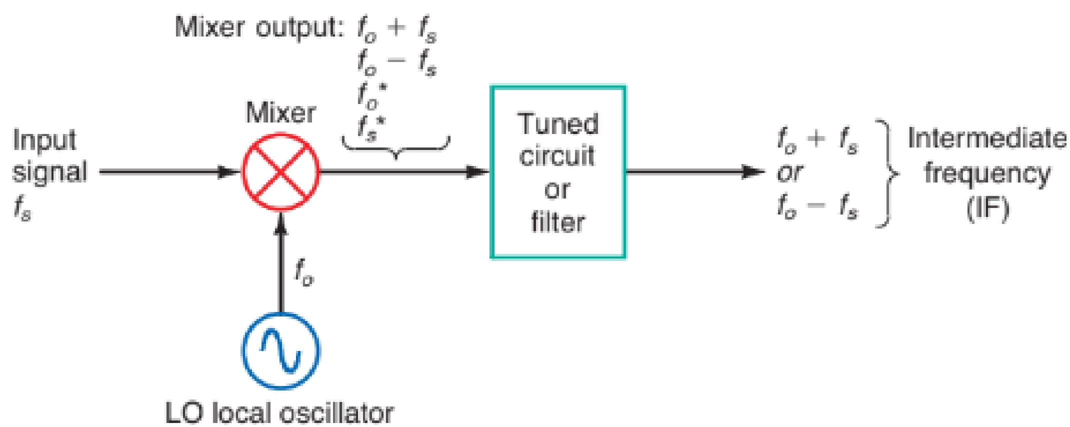
Mixers and Frequency Conversion:

The output of the RF amplifier is applied to the input of the mixer. The mixer also receives an input from a local oscillator or frequency synthesizer. The mixer output is the input signal, the local oscillator signal, and the sum and difference frequencies of these signals. Usually, a tuned circuit at the output of the mixer selects the difference frequency, or intermediate frequency (IF). The sum frequency may also be selected as the IF in some applications. The mixer may be a diode, a balanced modulator, or a transistor. MOSFETs and hot carrier diodes are preferred as mixers because of their low-noise characteristics.

The function performed by the mixer is called heterodyning. Mixers accept two inputs. The signal fs, which is to be translated to another frequency, is applied to one input, and the sine wave from a local oscillator f_o is applied to the other input. The signal to be translated can be a simple sine wave or any complex modulated signal containing sidebands. Like an amplitude modulator, a mixer essentially performs a mathematical multiplication of its two input signals. The output of the mixer, therefore, consists of signals f_o, f_s , f_s+f_o and f_s-f_o . The filter takes the required combination of the signal frequency.

The process is also termed as frequency translation or conversion, f_s+f_o is up-conversion and f_s-f_o is termed as down-conversion

Concept of a mixer.

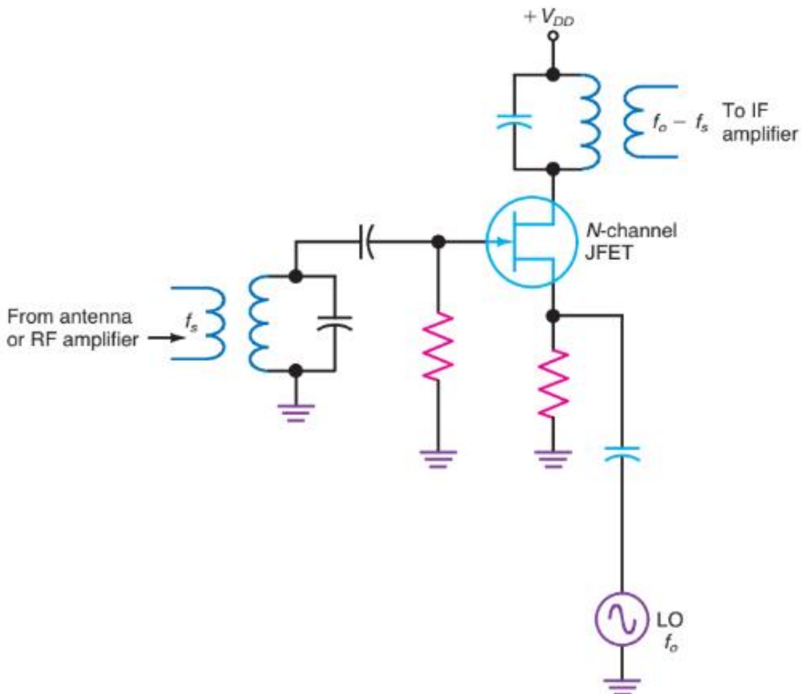


The AM and FM specification for IF is given in the table below

	AM Radio	FM Radio
RF carrier range	0.535–1.605 MHz	88–108 MHz
Mid-band frequency of IF section	0.455 MHz	10.7 MHz
IF bandwidth	10 kHz	200 kHz

a) JFET Mixer:

FETs make good mixers because they provide gain, have low noise, and offer a nearly perfect square-law response. The FET mixer is biased such that, it operates in the nonlinear portion of its range. The input signal is applied to the gate, and the local oscillator signal is coupled to the source. Again, the tuned circuit in the drain selects the difference frequency. The circuit diagram of a JFET mixer is given below.



JFET Mixer

Local Oscillators:

The local oscillator is made tunable so that its frequency can be adjusted over a relatively wide range. As the local oscillator frequency is changed, the mixer translates a wide range of input frequencies to the fixed IF. There are no set rules for deciding which of these to choose. However, at lower frequencies, say, those less than about 100 MHz, the local oscillator frequency is traditionally higher than the incoming signal's frequency, and at higher frequencies, those above 100 MHz, the local oscillator frequency is lower than the input signal frequency.

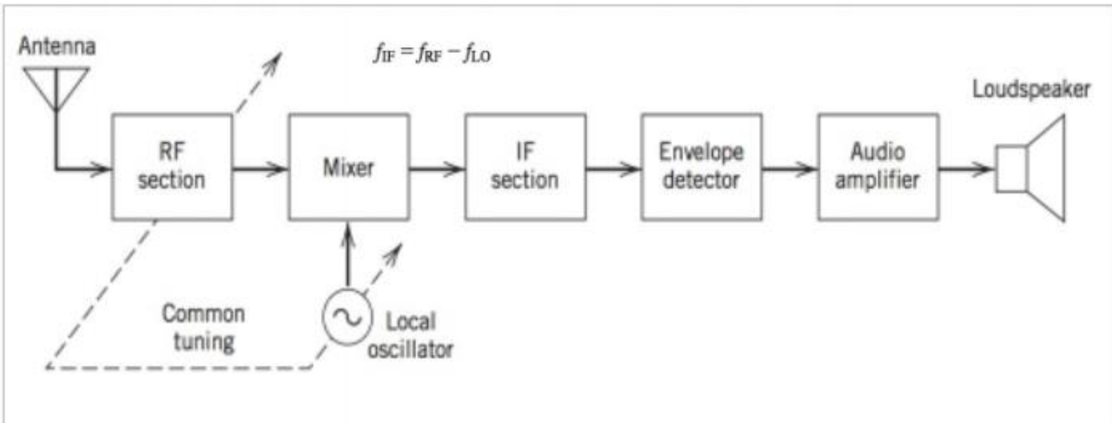
IF Amplifiers

The output of the mixer is an IF signal containing the same modulation that appeared on the input RF signal. This signal is amplified by one or more IF amplifier stages, and most of the receiver gain is obtained in these stages. Selectively tuned circuits provide fixed selectivity. Since the intermediate frequency is usually much lower than the input signal frequency, IF amplifiers are easier to design, and good selectivity is easier to obtain.

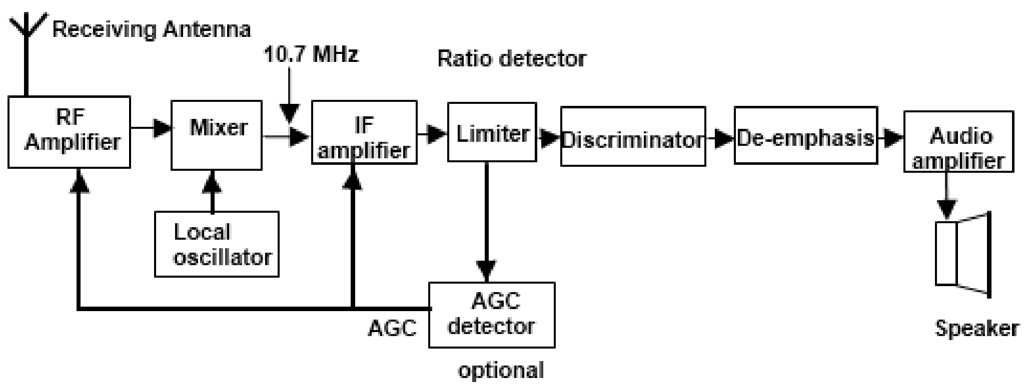
Demodulators:

The highly amplified IF signal is finally applied to the demodulator, or detector, which recovers the original modulating information. The demodulator may be a diode detector (for AM), a quadrature detector (for FM), or a product detector (for SSB). In modern digital superheterodyne radios, the IF signal is first digitized by an analog-to-digital converter (ADC) and then sent to a digital signal processor (DSP) where the demodulation is carried out by a programmed algorithm. The recovered signal in digital form is then converted back to analog by a digital-to-analog converter (DAC). The output of the demodulator or DAC is then usually fed to an audio amplifier with sufficient voltage and power gain to operate a speaker.

AM and FM superheterodyne receiver block diagram is given below



Block diagram of AM Receiver



Block diagram of FM Receiver

Automatic Gain Control:

The output of a demodulator is usually the original modulating signal, the amplitude of which is directly proportional to the amplitude of the received signal. The recovered signal, which is usually ac, is rectified and filtered into a dc voltage by a circuit known as the automatic gain control (AGC) circuit. This dc voltage is fed back to the IF amplifiers, and sometimes the RF amplifier, to control receiver gain. AGC circuits help maintain a constant output voltage level over a wide range of RF input signal levels; they also help the receiver to function over a wide range so that strong signals do not produce performance-degrading distortion.

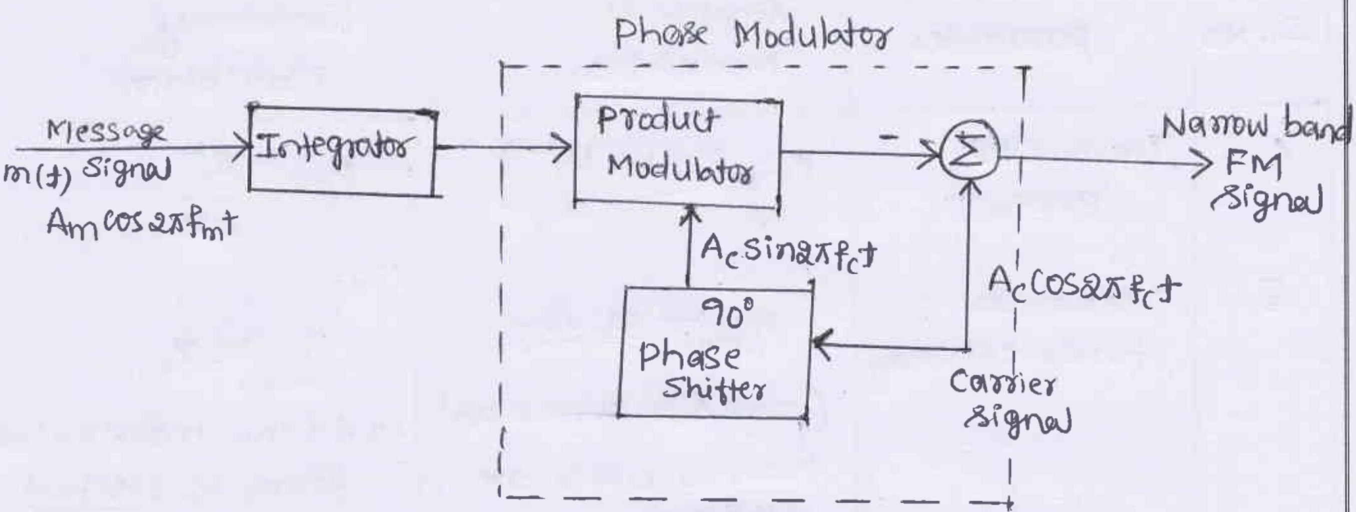


Figure 1: Indirect Method of Generating FM signal (Narrow band) Using phase modulator :-

Figure 1, shows the indirect method of generating Narrow band FM signal shown in figure (5), using phase modulator.

Note: The Band width required to transmit Narrow band FM signal is same as that of AM-signal transmission channel bandwidth "$2f_m$".

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* Comparison between AM and FM :-

Sl. No.	parameter	Amplitude Modulation	Frequency Modulation
1.	Altering parameter of Carrier	Amplitude	Frequency.
2.	Constant parameter of Carrier	Frequency and Phase	Amplitude, phase
3.	Modulation Index	$\mu = K_a A_m < 1$	$\beta = \frac{\Delta f_{max}}{f_m}$ $\beta < 1$ (Narrow band) $\beta > 1$ (Wide band)

Sl. No	parameter	Amplitude Modulation	Frequency Modulation
4.	Transmitted power	$P_t = P_c\left[1+\frac{\mu_t^2}{2}\right]$	$P_t = \frac{A_c^2}{2R}$
5.	Maximum power efficiency	$\eta_{max} = 33.33\%$ (one side power and Carrier power are wasted)	$\eta_{max} = 100\%$ i.e, All the transmitted power is <u>useful</u> <u>power</u>
6.	Bandwidth	• $BW = 2f_m$ • Band width is Independent of Modu--lation Index	$BW = 2f_m$ Narrowband (or) $BW = 2f_m + 2\Delta f_{max}$ (wide band) • Band width depends on Modulation Index (wide band FM)
7.	Range of Communication	Covers Large Area Ex: Radio	Covers limited Area EX: FM-channels
8.	complexity	Less Complex	More Complex
9.	Cost	Inexpensive	Expensive
10.	Noise Immunity	Affected by Noise	Immune to Noise.
11.	Types	- DSBSC • SSBSC	• Narrow band • Wide band
12.	Applications	Long distance Communication Ex: Radi	Short distance Communi-cation <u>Ex</u>: FM- stations.

7b Carsons rule Transmission bandwidth = 2(frequency deviation + modulating frequency)

= 2(300+15) = 630kHz

Universal Curve method

Transmission bandwidth = 2nmaxfm = 50 x 15k = 750kHz.