MALLA REDDY ENGINEERING COLLEGE

(Autonomous)

LECTURE NOTES

ON

ANALOG AND DIGITAL COMMUNICATIONS

(80409)

B.Tech-ECE-IV semester

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ELECTRONICS AND COMMUNICATION ENGINEERING

MALLA REDDY ENGINEERING COLLEGE (Autonomous)

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<tr><td>2018-19 Onwards (MR-18)</td><td>MALLA REDDY ENGINEERING COLLEGE (Autonomous)</td><td colspan="3">B.Tech. IV Semester</td></tr>
<tr><td>Code: 80409</td><td rowspan="2">ANALOG & DIGITAL COMMUNICATIONS</td><td>L</td><td>T</td><td>P</td></tr>
<tr><td>Credits:4</td><td>3</td><td>1</td><td>-</td></tr>
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Pre-Requisites: Signals and Systems, Probability Theory and Stochastic Processes.

Course Objectives: This course introduces the concept of modulation and various techniques for amplitude modulation of analog signals. This course also introduces the concept of angle modulation techniques for Frequency modulation of analog signals. This course also introduces the radio transmitters and receivers, the effect of noise on communication systems and various pulse analog & digital binary modulation techniques.

MODULE I: Amplitude Modulation Techniques [13 Periods]

Introduction to communication system, Need for modulation, Amplitude Modulation, Definition, Time domain and frequency domain description of AM system, single tone modulation, power relations in AM waves. Time domain and frequency domain description of DSB-SC**,** SSB-SCand VSB-SC systems**.** Comparison of AM Techniques, Applications of different AM Systems.

MODULE II: Frequency Modulation Techniques [12 Periods]

Basic concepts, Frequency Modulation: Single tone frequency modulation, Spectrum Analysis of Sinusoidal FM Wave, Narrow band FM, Wide band FM, Constant Average Power, Transmission bandwidth of FM Wave - Generation of FM Waves with Direct and Indirect methods, Detection of FM Waves: Balanced Frequency discriminator, Phase locked loop, Comparison of FM and AM.

MODULE III: Radio Transmitters &Receivers [14 Periods]

Transmitters: Block diagram of AM Transmitter and FM Transmitter. Types of Noise: Resistive (Thermal) Noise Source, Shot noise, Extraterrestrial Noise, Arbitrary Noise Sources, White Noise, Narrowband Noise- In phase and Quadrature phase components and its Properties, Average Noise Figures, Average Noise Figure of cascaded networks. Noise Analysis in AM and FM Systems.

Radio Receivers: Introduction, Receiver Types - Tuned radio frequency receiver, Super hetrodyne receiver, RF section and Characteristics - Frequency changing and tracking, Intermediate frequency, AGC, AM & FM Receivers, Comparison with AM Receiver, Amplitude limiting. Frequency Division Multiplexing.

MODULE IV: Elements of Digital Communication Systems [13 Periods]

Model of Digital Communication System, Advantages of Digital Communication Systems.

Pulse Analog Modulation: Introduction, PAM, PWM, PPM Modulation and Demodulation Techniques.

Pulse Digital Modulation: PCM Generation and Reconstruction, Quantization Noise, Non Uniform Quantization and Companding, DPCM, Adaptive DPCM, DM and Adaptive DM, Noise in PCM and DM.

MODULE V: Digital Binary Carrier Modulation Schemes **[12 Periods]**

Introduction, ASK, ASK Modulator, Coherent ASK Detector, Non-Coherent ASK Detector, FSK, Bandwidth and Frequency Spectrum FSK, Non Coherent FSK Detector, Coherent FSK Detector, FSK Detection using PLL, BPSK, Coherent PSK Detection, Differential PSK.

Text Books:

1.H Taub& D. Schilling, GautamSahe, "Principles of Communication Systems", TMH, 3rd Edition, 2007.

2.Sam Shanmugam, "Digital and Analog Communication Systems", John Wiley, 2005

Reference Books:

1. Simon Haykin, John Wiley, "Digital Communication", 1st Edition, 2005.
2. B.P. Lathi, "Communication Systems", BS Publication, 2006.

MODULE I

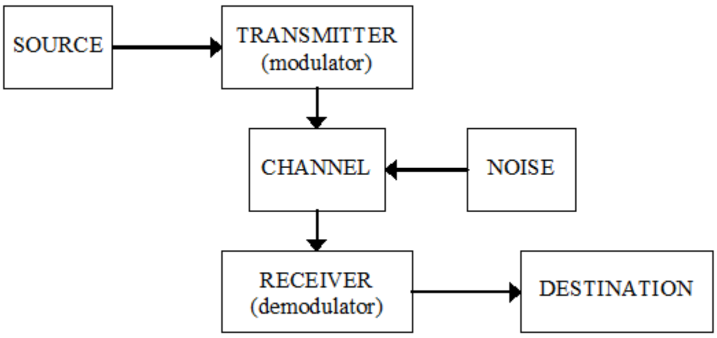
Amplitude Modulation Techniques

Introduction to Communication System

Communication is the process by which information is exchanged between individuals through a medium.

Communication can also be defined as the transfer of information from one point in space and time to another point.

The basic block diagram of a communication system is as follows.



- **Transmitter:** Couples the message into the channel using high frequency signals.
- **Channel:** The medium used for transmission of signals
- **Modulation:** It is the process of shifting the frequency spectrum of a signal to a frequency range in which more efficient transmission can be achieved.
- **Receiver:** Restores the signal to its original form.
- **Demodulation:** It is the process of shifting the frequency spectrum back to the original baseband frequency range and reconstructing the original form.

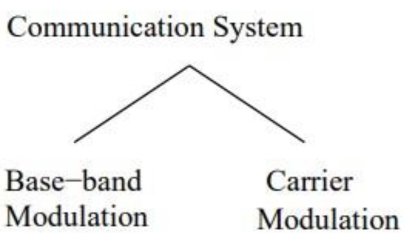
Modulation:

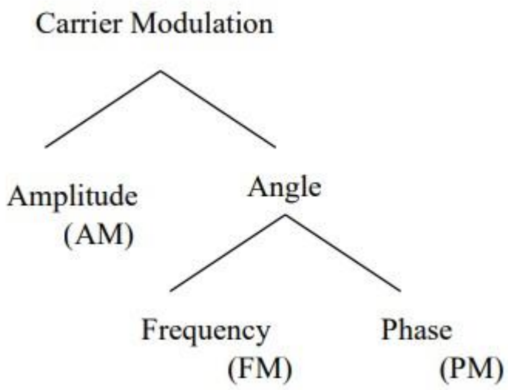
Modulation is a process that causes a shift in the range of frequencies in a signal.

• Signals that occupy the same range of frequencies can be separated.

• Modulation helps in noise immunity, attenuation - depends on the physical medium.

The below figure shows the different kinds of analog modulation schemes that are available





Modulation is operation performed at the transmitter to achieve efficient and reliable information transmission.

For analog modulation, it is frequency translation method caused by changing the appropriate quantity in a carrier signal.

It involves two waveforms:

- A modulating signal/baseband signal – represents the message.
- A carrier signal – depends on type of modulation.

•Once this information is received, the low frequency information must be removed from the high frequency carrier. •This process is known as “Demodulation”.

Need for Modulation:

- Baseband signals are incompatible for direct transmission over the medium so, modulation is used to convey (baseband) signals from one place to another.
- Allows frequency translation:
  - Frequency Multiplexing
  - Reduce the antenna height
  - Avoids mixing of signals
  - Narrowbanding
- Efficient transmission
- Reduced noise and interference

Amplitude Modulation (AM)

Amplitude Modulation is the process of changing the amplitude of a relatively high frequency carrier signal in accordance with the amplitude of the modulating signal (Information).

The carrier amplitude varied linearly by the modulating signal which usually consists of a range of audio frequencies. The frequency of the carrier is not affected.

- Application of AM - Radio broadcasting, TV pictures (video), facsimile transmission
- Frequency range for AM - 535 kHz – 1600 kHz
- Bandwidth - 10 kHz

Various forms of Amplitude Modulation

- Conventional Amplitude Modulation (Alternatively known as Full AM or Double

Sideband Large carrier modulation (DSBLC) /Double Sideband Full Carrier (DSBFC)

- Double Sideband Suppressed carrier (DSBSC) modulation
- Single Sideband (SSB) modulation
- Vestigial Sideband (VSB) modulation

Time Domain and Frequency Domain Description

It is the process where, the amplitude of the carrier is varied proportional to that of the message signal.

Let m (t) be the base-band signal, m (t) ←→ M (ω) and c (t) be the carrier, $c(t) = A_c \cos(\omega_c t)$. fc is chosen such that fc >> W, where W is the maximum frequency component of m(t). The amplitude modulated signal is given by

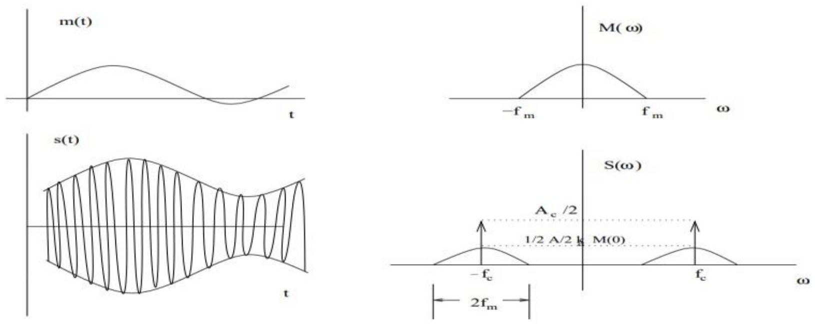
$$s(t) = A_c [1 + k_a m(t)] \cos(\omega_c t)$$

Fourier Transform on both sides of the above equation

$$S(\omega) = \pi A_c/2 (\delta(\omega - \omega_c) + \delta(\omega + \omega_c)) + k_a A_c/2 (M(\omega - \omega_c) + M(\omega + \omega_c))$$

k_a is a constant called amplitude sensitivity.

$k_a m(t) < 1$ and it indicates percentage modulation.



Single Tone Modulation:

Consider a modulating wave m(t) that consists of a single tone or single frequency

$$m(t) = A_m \cos(2\pi f_m t) \quad \text{......................(1)}$$

where A_m is peak amplitude of the sinusoidal modulating wave

f_m is the frequency of the sinusoidal modulating wave

Let A_c be the peak amplitude and f_c be the frequency of the high frequency carrier signal. Then the corresponding single-tone AM wave is given by

$$s(t) = A_c[1 + m\cos(2\pi f_m t)]Cos(2\pi f_c t) \quad \text{......................(2)}$$

Let A_{max} and A_{min} denote the maximum and minimum values of the envelope of the modulated wave. Then from the above equation (2.12), we get

$$\frac{A_{max}}{A_{min}} = \frac{A_c(1+m)}{A_c(1-m)}$$

$$m = \frac{A_{max} - A_{min}}{A_{max} + A_{min}}$$

component given by

Expanding the equation (2), we get

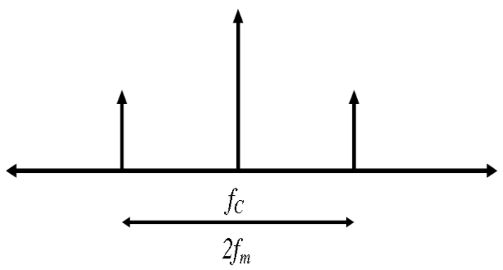
$$s(t) = A_c \cos(2\pi f_c t) + \frac{1}{2} m A_c \cos[2\pi(f_c + f_m)t] + \frac{1}{2} m A_c \cos[2\pi(f_c - f_m)t]$$

The Fourier transform of $s(t)$ is obtained as follows.

$$s(f) = \frac{1}{2} A_c[\delta(f - f_c) + \delta(f + f_c)] + \frac{1}{4} m A_c[\delta(f - f_c - f_m) + \delta(f + f_c + f_m)]$$

$$+ \frac{1}{4} m A_c[\delta(f - f_c + f_m) + \delta(f + f_c - f_m)]$$

Thus the spectrum of an AM wave, for the special case of sinusoidal modulation consists of delta functions at $\pm f_c$, $f_c \pm f_m$, and $-f_c \pm f_m$. The spectrum for positive frequencies is as shown in figure



Power relations in AM waves:

Consider the expression for single tone/sinusoidal AM wave

$$s(t) = A_c Cos(2\pi f_c t) + \frac{1}{2} m A_c Cos[2\pi(f_c + f_m)t] + \frac{1}{2} m A_c Cos[2\pi(f_c - f_m)t] \quad \text{.................(1)}$$

This expression contains three components. They are carrier component, upper side band and lower side band. Therefore Average power of the AM wave is sum of these three components.

Therefore the total power in the amplitude modulated wave is given by

$$Pt = \frac{V^2_{car}}{R} + \frac{V^2_{LSB}}{R} + \frac{V^2_{USB}}{R} \quad \text{........................(2)}$$

Where all the voltages are rms values and R is the resistance, in which the power is dissipated.

Therefore total average power is given by

$$P_t = P_c + P_{LSB} + P_{USB}$$

$$P_t = P_c + \frac{m^2}{4}P_c + \frac{m^2}{4}P_c$$

$$P_t = P_c\left(1 + \frac{m^2}{4} + \frac{m^2}{4}\right)$$

$$P_t = P_c\left(1 + \frac{m^2}{2}\right) \qquad \text{.............................(3)}$$

$$P_C = \frac{V^2{}_{car}}{R} = \frac{\left(A_c/\sqrt{2}\right)^2}{R} = \frac{A_c{}^2}{2R}$$

$$P_{LSB} = \frac{V_{LSB}{}^2}{R} = \left(\frac{mA_c}{2\sqrt{2}}\right)^2 \frac{1}{R} = \frac{m^2 A_c{}^2}{8R} = \frac{m^2}{4}P_c$$

$$P_{USB} = \frac{V_{USB}{}^2}{R} = \left(\frac{mA_c}{2\sqrt{2}}\right)^2 \frac{1}{R} = \frac{m^2 A_c{}^2}{8R} = \frac{m^2}{4}P_c$$

The ratio of total side band power to the total power in the modulated wave is given by

$$\frac{P_{SB}}{P_t} = \frac{P_c(m^2/2)}{P_c(1 + m^2/2)}$$

$$\frac{P_{SB}}{P_t} = \frac{m^2}{2 + m^2} \qquad \text{.......................(4)}$$

This ratio is called the efficiency of AM system

Advantages and Disadvantages of

AM: Advantages of AM:

- Generation and demodulation of AM wave are easy.
- AM systems are cost effective and easy to build

Disadvantages:

- AM contains unwanted carrier component, hence it requires more transmission power.
- The transmission bandwidth is equal to twice the message bandwidth.

To overcome these limitations, the conventional AM system is modified at the cost of increased system complexity. Therefore, three types of modified AM systems are discussed.

DSB-SC Time domain and Frequency domain Description:

DSBSC modulators make use of the multiplying action in which the modulating signal multiplies the carrier wave. In this system, the carrier component is eliminated and both upper and lower side bands are transmitted. As the carrier component is suppressed, the power required for transmission is less than that of AM.

If $m(t)$ is the message signal and $c(t) = A_c \cos(2\pi f_c t)$ is the carrier signal, then DSBSC modulated wave $s(t)$ is given by

$$s(t) = c(t)\, m(t)$$

$$s(t) = A_c \cos(2\pi f_c t) m(t)$$

Consequently, the modulated signal s(t) under goes a phase reversal , whenever the message signal m(t) crosses zero as shown below.

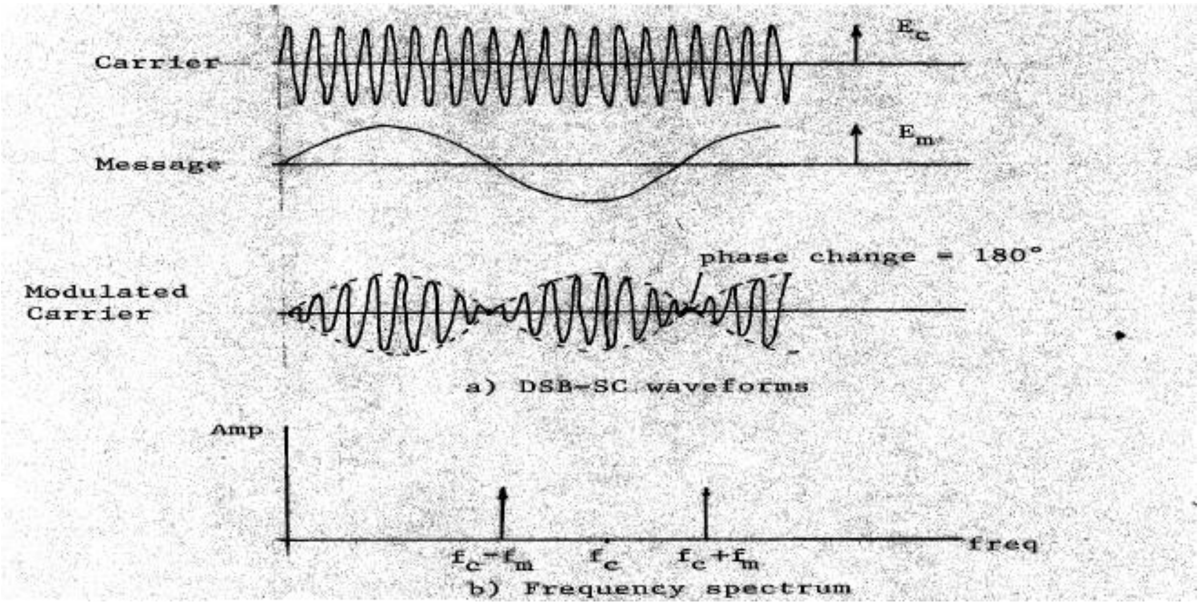


Fig.1. (a) DSB-SC waveform (b) DSB-SC Frequency Spectrum

The envelope of a DSBSC modulated signal is therefore different from the message signal and the Fourier transform of s(t) is given by

$$S(f)=\frac{A_c}{2}\left[M(f-f_c)+M(f+f_c)\right]$$

For the case when base band signal *m(t)* is limited to the interval $-W<f<W$ as shown in figure below, we find that the spectrum *S(f)* of the DSBSC wave *s(t)* is as illustrated below. Except for a change in scaling factor, the modulation process simply translates the spectrum of the base band signal by f_c. The transmission bandwidth required by DSBSC modulation is the same as that for AM.

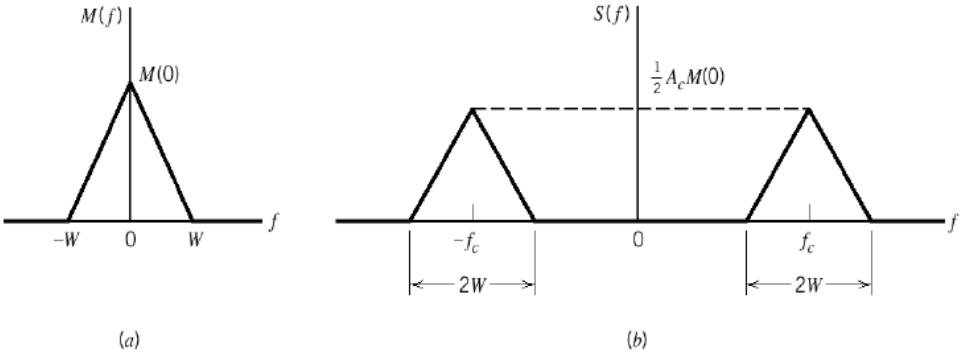


Figure: Message and the corresponding DSBSC spectrum

SSB-SC Time domain and Frequency domain Description:

Standard AM and DSBSC require transmission bandwidth equal to twice the message bandwidth. In both the cases spectrum contains two side bands of width W Hz, each. But the upper and lower sides are uniquely related to each other by the virtue of their symmetry about the carrier frequency. That is, given the amplitude and phase spectra of either side band, the other can be uniquely determined. Thus if only one side band is transmitted, and if both the carrier and the other side band are suppressed at the transmitter, no information is lost. This kind of modulation is called SSBSC and spectral comparison between DSBSC and SSBSC is shown in the figures 1 and 2.

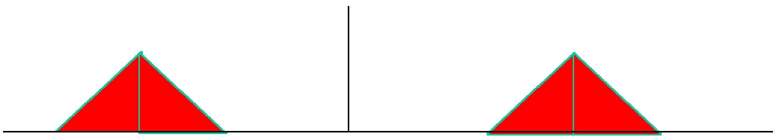


Figure.1 : Spectrum of the DSBSC wave

Figure .2 : Spectrum of the SSBSC wave

Frequency Domain Description

Consider a message signal $m(t)$ with a spectrum $M(f)$ band limited to the interval $-w < f < w$ as shown in figure 3 , the DSBSC wave obtained by multiplexing $m(t)$ by the carrier wave $c(t) = A_c \cos(2\pi f_c t)$ and is also shown, in figure 4 . The upper side band is represented in duplicate by the frequencies above f_c and those below $-f_c$, and when only upper

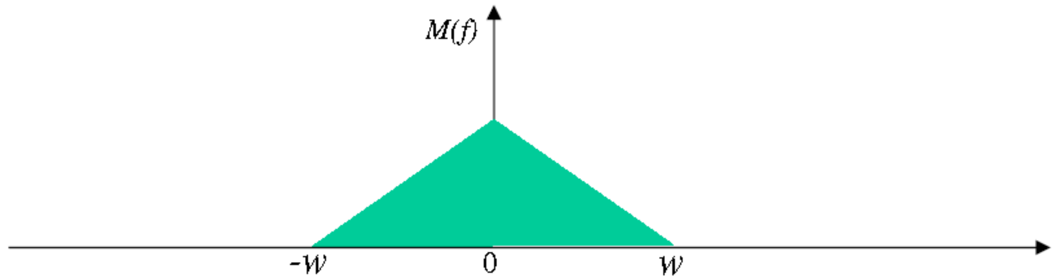


Figure 3. : Spectrum of message wave

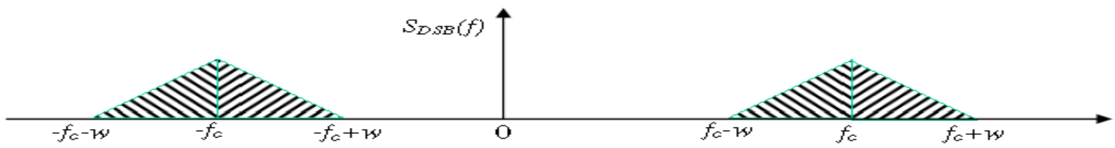


Figure .4 : Spectrum of DSBSC wave

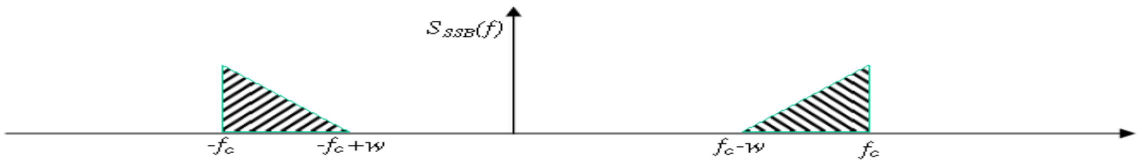


Figure.5 : Spectrum of SSBSC-LSB wave

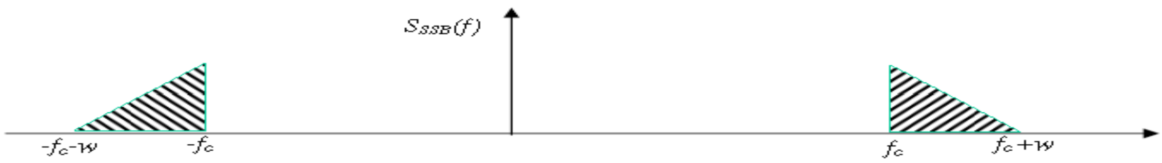


Figure .6 : Spectrum of SSBSC-USB wave

side band is transmitted; the resulting SSB modulated wave has the spectrum shown in figure 1. Similarly, the lower side band is represented in duplicate by the frequencies below fc and those above -fc and when only the lower side band is transmitted, the spectrum of the corresponding SSB modulated wave shown in figure 5.Thus the essential function of the SSB modulation is to translate the spectrum of the modulating wave, either with or without inversion, to a new location in the frequency domain. The advantage of SSB modulation is reduced bandwidth and the elimination of high power carrier wave. The main disadvantage is the cost and complexity of its implementation.

Time Domain Description:

The time domain description of an SSB wave s(t) in the canonical form is given by the equation 1.

$$s(t) = s_I(t)\cos(2\pi f_c t) - s_Q(t)\sin(2\pi f_c t) \quad \text{------------------- (1)}$$

where $S_I(t)$ is the in-phase component of the SSB wave and $S_Q(t)$ is its quadrature component. The in-phase component $S_I(t)$ except for a scaling factor, may be derived from $S(t)$ by first multiplying $S(t)$ by $\cos(2\pi f_c t)$ and then passing the product through a low-pass filter. Similarly, the quadrature component $S_Q(t)$, except for a scaling factor, may be derived from $s(t)$ by first multiplying $s(t)$ by $\sin(2\pi f_c t)$ and then passing the product through an identical filter.

The Fourier transformation of $S_I(t)$ and $S_Q(t)$ are related to that of SSB wave as follows, respectively.

$$S_I(f) = \begin{cases} S(f - f_c) + S(f + f_c), -w \le f \le w \\ 0, elsewhere \end{cases} \quad \text{------------------- (2)}$$

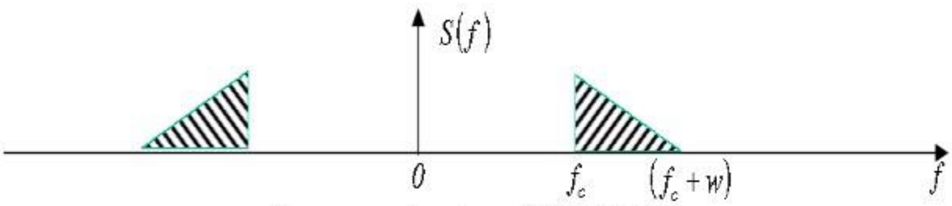


Figure 10 : Spectrum of SSBSC-USB

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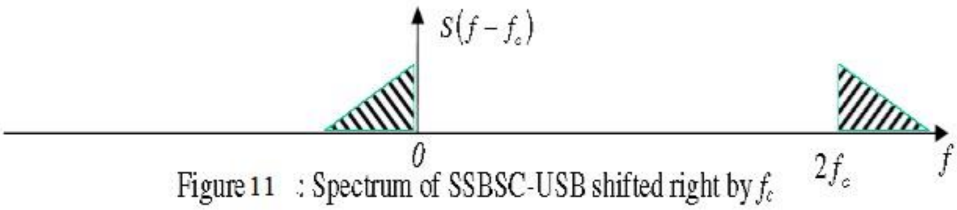


Figure 11 : Spectrum of SSBSC-USB shifted right by f_c

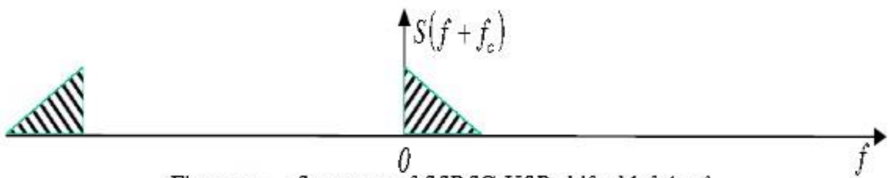


Figure 12 : Spectrum of SSBSC-USB shifted left by f_c

$$S_Q(f) = \begin{cases} j[S(f - f_c) - S(f + f_c)], -w \le f \le w \\ 0, elsewhere \end{cases} \quad \text{------------------ (3)}$$

where $-w < f < w$ defines the frequency band occupied by the message signal $m(t)$.

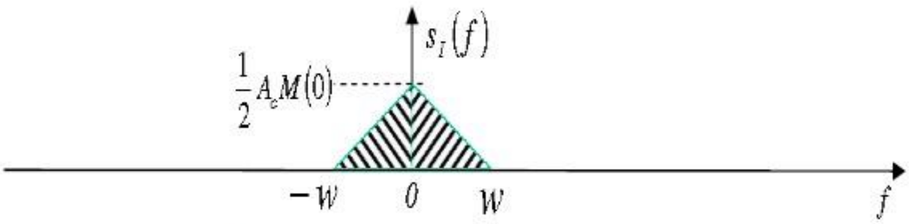


Figure 13 : Spectrum of in-phase component of SSBSC-USB

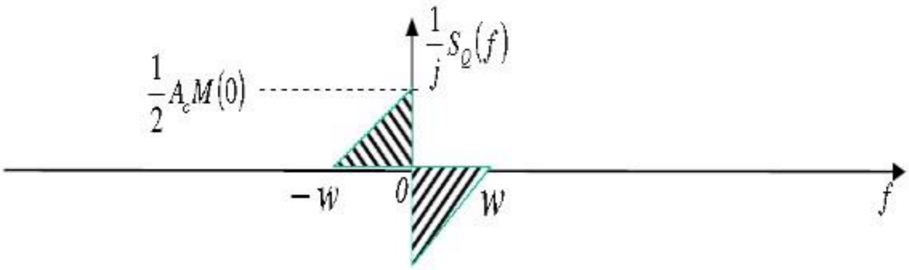


Figure 14 : Spectrum of quadrature component of SSBSC-USB

From the figure 13 , it is found that

$$S_I(f)=\frac{1}{2}A_cM(f)$$

where *M(f)* is the Fourier transform of the message signal *m(t)*. Accordingly in-phase component $S_I(t)$ is defined by equation 4

$$\boxed{s_I(t)=\frac{1}{2}A_cm(t)} \qquad \text{------------------ (4)}$$

Now on the basis of figure14 , it is found that

$$S_Q(f)=\begin{cases}\frac{-j}{2}A_cM(f), f>0\\ 0, f=0\\ \frac{j}{2}A_cM(f), f<0\end{cases}$$

$$S_Q(f)=\frac{-j}{2}A_c\,\text{sgn}(f)M(f) \qquad \text{------------------ (5)}$$

where $\text{sgn}(f)$ is the Signum function.

But from the discussions on Hilbert transforms, it is shown that

$$-j\,\mathrm{sgn}(f)M(f)=\hat{M}(f) \qquad \text{-------------------- (6)}$$

where $\hat{M}(f)$ is the Fourier transform of the Hilbert transform of $m(t)$. Hence the substituting equation (6) in (5), we get

$$S_Q(f)=\frac{1}{2}A_c\hat{M}(f) \qquad \text{------------------------ (7)}$$

Therefore quadrature component $s_Q(t)$ is defined by equation 8

$$s_Q(t)=\frac{1}{2}A_c\hat{m}(t) \qquad \text{------------------------- (8)}$$

Therefore substituting equations (4) and (8) in equation in (1), we find that canonical representation of an SSB wave $s(t)$ obtained by transmitting only the upper side band is given by the equation 9

$$s_U(t)=\frac{1}{2}A_c m(t)\cos(2\pi f_c t)-\frac{1}{2}A_c\hat{m}(t)\sin(2\pi f_c t) \qquad \text{-------------- (9)}$$

Following the same procedure, we can find the canonical representation for an SSB wave
s(t) obtained by transmitting only the lower side band is given by

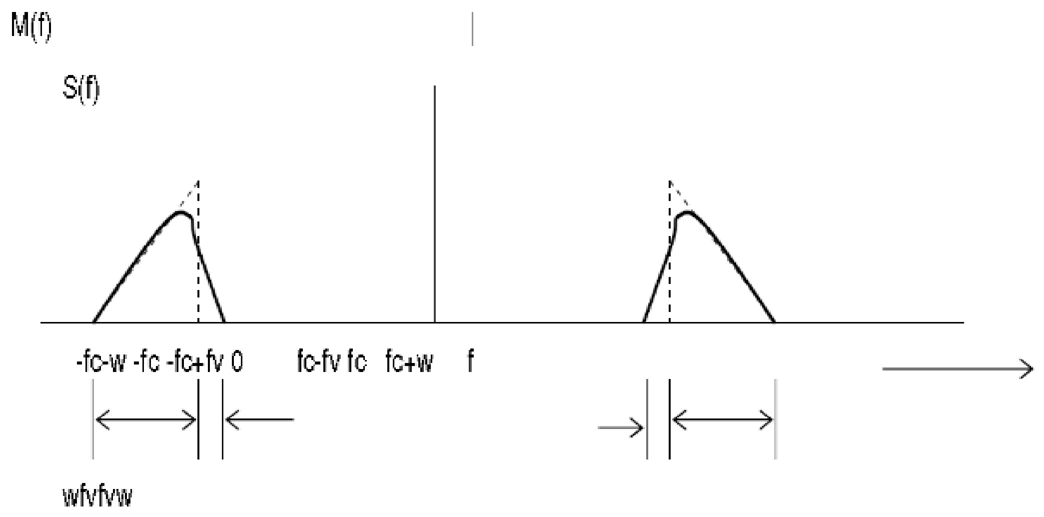
$$s_L(t)=\frac{1}{2}A_c m(t)\cos(2\pi f_c t)+\frac{1}{2}A_c\hat{m}(t)\sin(2\pi f_c t) \qquad \text{--------------- (10)}$$

VSB-SC Time domain and Frequency domain Description:

Vestigial sideband is a type of Amplitude modulation in which one side band is completely passed along with trace or tail or vestige of the other side band. VSB is a compromise between SSB and DSBSC modulation. In SSB, we send only one side band, the Bandwidth required to send SSB wave is w. SSB is not appropriate way of modulation when the message signal contains significant components at extremely low frequencies. To overcome this VSB is used.

Frequency Domain Description

The following Fig illustrates the spectrum of VSB modulated wave s (t) with respect to the message m (t) (band limited)



Fig(b) Spectrum of VSB wave containing vestige of the Lower side band

Assume that the Lower side band is modified into the vestigial side band. The vestige of the lower sideband compensates for the amount removed from the upper sideband. The bandwidth required to send VSB wave is

$$B = w + f_v$$

Where f_v is the width of the vestigial side band.

Similarly, if Upper side band is modified into the vestigial side band then,

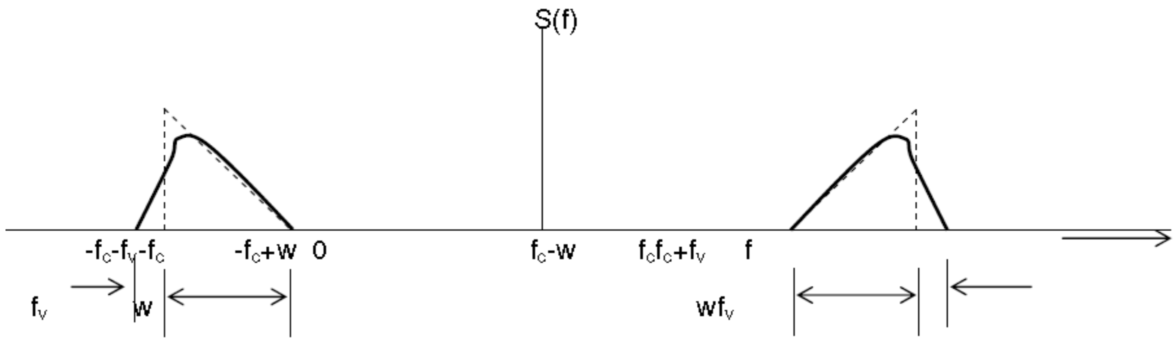


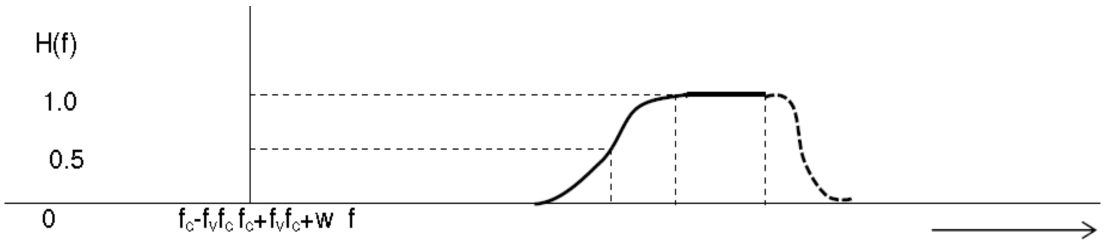
Fig (d) Spectrum of VSB wave containing vestige of the Upper side band

The vestige of the Upper sideband compensates for the amount removed from the Lower sideband. The bandwidth required to send VSB wave is B = w+fv, where fv is the width of the vestigial side band.

Therefore, VSB has the virtue of conserving bandwidth almost as efficiently as SSB modulation, while retaining the excellent low-frequency base band characteristics of DSBSC and it is standard for the transmission of TV signals

Time Domain Description:

Time domain representation of VSB modulated wave, procedure is similar to SSB Modulated waves. Let s(t) denote a VSB modulated wave and assuming that s(t) containing Upper sideband along with the Vestige of the Lower sideband. VSB modulated wave s(t) is the output from Sideband shaping filter, whose input is DSBSC wave. The filter transfer function H(f) is of the form as in fig below,



To determine $\tilde{s}(t)$ we proceed as follows

1. The side band shaping filter transfer function H(f) is replaced by its equivalent complex low pass transfer function denoted by $\tilde{H}(f)$ as shown in fig below

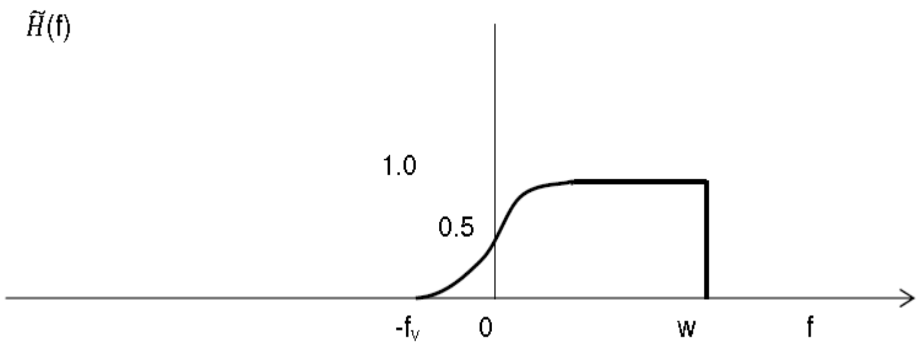


Fig (2) Low pass equivalent to H(f)

We may express $\widetilde{H}$(f) as the difference between two components $\widetilde{Hu}$(f) and $\widetilde{Hv}$(f) as

$\widetilde{H}$(f) = $\widetilde{Hu}$(f) - $\widetilde{Hv}$(f) -----------------------------(4)

These two components are considered individually as follows

i). The transfer function $\widetilde{Hu}$(f) pertains to a complex low pass filter equivalent to a band pass filter design to reject the lower side band completely as

$\widetilde{Hu}$(f)

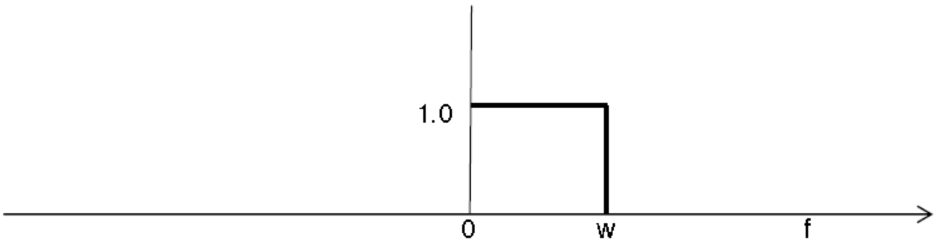


Fig (3) First component of $\widetilde{H}$(f)

$$\widetilde{Hu}(f) = \begin{cases} \frac{1}{2}\,[1+\mathrm{sgn}(f)], & 0<f<w \\ 0, & \text{otherwise} \end{cases} \quad \text{-----------------(5)}$$

ii). The transfer function $\widetilde{Hv}$(f) accounts for the generation of vestige and removal of a corresponding portion from the upper side band.

$\widetilde{Hv}$(f)

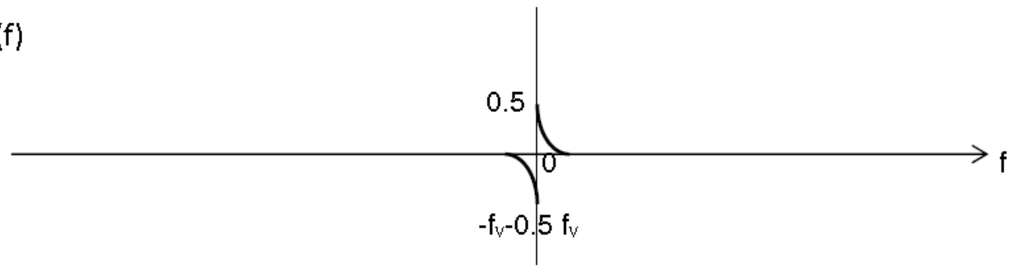


Fig (4) Second component of $\widetilde{H}$(f)

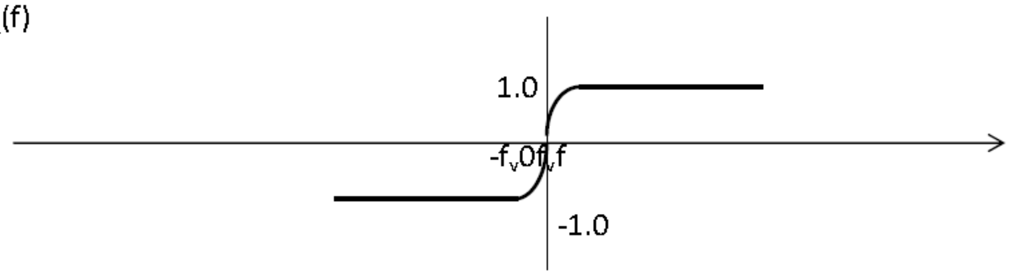
Substitute eqn(5) in eqn (4) we get,

$$\widetilde{H}(f) = \begin{cases} \frac{1}{2}\,[1+\mathrm{sgn}(f) - 2\widetilde{Hv}(f)], & f_v<f<w \\ 0, & \text{otherwise} \end{cases} \quad \text{-----------------(6)}$$

The sgn(f) and $\widetilde{Hv}$(f) are both odd functions of frequency, Hence, both they have purely imaginary Inverse Fourier Transform (FT). Accordingly, we may introduce a new transfer function

$H_Q(f) = 1/j[\text{sgn}(f) - 2\widetilde{Hv}(f)]$ ------------(7)

It has purely Inverse FT and $h_Q(t)$ denote IFT of $H_Q(f)$



Fig(5) Transfer function of the filter $jH_Q(f)$

Rewrite eqn(6) interms of $H_Q(f)$ as

$$\tilde{H}(f) = \begin{cases} \tfrac{1}{2}[1+jH_Q(f)], & f_v<f<w \\ 0, & \text{otherwise} \end{cases} \quad \text{-----------------(8)}$$

2. The DSBSC modulated wave is replaced by its complex envelope as
   $\tilde{S}_{DSBSC}(f) = A_c\, M(f)$ -----------------(9)
3. The desired complex envelope $\tilde{s}(t)$ is determined by evaluating IFT of the product $\tilde{H}(f)\tilde{S}_{DSBSC}(f)$.
   i.e., $\tilde{S}(f) = \tilde{H}(f)\tilde{S}_{DSBSC}(f)$ -----------------------(10)
   $\tilde{S}(f) = A_c/2[1+jH_Q(f)]\, M[f]$ --------------------------(11)
   Take IFT of eqn(11) we get,
   $\tilde{s}(t) = A_c/2[m(t) + jm_Q(t)]$ ----------------------(12)
   Where $m_Q(t)$ is the response produced by passing the message through a low pass filter of impulse response $h_Q(t)$.
   Substitute eqn(12) in eqn(3) and simplify, we get
   $S(t) = A_c/2\, m(t) \cos 2\pi f_c t - A_c/2\, m_Q(t) \sin 2\pi f_c t$ -------------------(13)
   Where $A_c/2\, m(t)$ ----- In-phase component
   $A_c/2\, m_Q(t)$----- Quadrature component

Note:

1. If vestigial side band is increased to full side band, VSB becomes DSCSB ,i.e., mQ(t) = 0.

2. If vestigial side band is reduced to Zero, VSB becomes SSB.
   i.e., $m_Q(t) = \hat{m}(t)$
   Where $\hat{m}(t)$ is the Hilbert transform of m(t).

Comparison of AM Techniques:

Sr. No.	Parameter	Standard AM	SSB	DSBSC	VSB
1	Power	High	Less	Medium	Less than DSBSC but greater than SSB
2	Bandwidth	$2 f_m$	f_m	$2 f_m$	$f_m < B_w < 2 f_m$
3	Carrier supression	No	Yes	Yes	No
4	Receiver complexity	Simple	Complex	Complex	Simple
5	Application	Radio communication	Point to point communication preferred for long distance transmission.	Point to point communication	Television broadcasting
6	Modulation type	Non linear	Linear	Linear	Linear
7	Sideband suppression	No	One sided completely	No	One sideband suppressed partly
8	Transmission efficiency	Minimum	Maximum	Moderate	Moderate

Applications of different AM systems:

- Amplitude Modulation: AM radio, Short wave radio broadcast
- DSB-SC: Data Modems, Color TV's color signals.
- SSB: Telephone
- VSB: TV picture signals

MODULE II

Frequency Modulation Techniques

Basic concepts:

Instantaneous Frequency

The frequency of a cosine function $x(t)$ that is given by

$$x(t) = \cos(\omega_c t + \theta_0)$$

is equal to ω_c since it is a constant with respect to t, and the phase of the cosine is the constant θ_0. The angle of the cosine $\theta(t) = \omega_c t + \theta_0$ is a linear relationship with respect to t (a straight line with slope of ω_c and y–intercept of θ_0). However, for other sinusoidal functions, the frequency may itself be a function of time, and therefore, we should not think in terms of the constant frequency of the sinusoid but in terms of the INSTANTANEOUS frequency of the sinusoid since it is not constant for all t. Consider for example the following sinusoid

$$y(t) = \cos\left[\theta(t)\right],$$

where $\theta(t)$ is a function of time. The frequency of $y(t)$ in this case depends on the function of $\theta(t)$ and may itself be a function of time. The instantaneous frequency of $y(t)$ given above is defined as

$$\omega_i(t) = \frac{d\theta(t)}{dt}.$$

As a checkup for this definition, we know that the instantaneous frequency of $x(t)$ is equal to its frequency at all times (since the instantaneous frequency for that function is constant) and is equal to ω_c. Clearly this satisfies the definition of the instantaneous frequency since $\theta(t) = \omega_c t + \theta_0$ and therefore $\omega_i(t) = \omega_c$.

If we know the instantaneous frequency of some sinusoid from $-\infty$ to sometime t, we can find the angle of that sinusoid at time t using

$$\theta(t) = \int_{-\infty}^{t} \omega_i(\alpha)d\alpha.$$

Changing the angle $\theta(t)$ of some sinusoid is the bases for the two types of angle modulation: Phase and Frequency modulation techniques.

Phase Modulation (PM)

In this type of modulation, the phase of the carrier signal is directly changed by the message signal. The phase modulated signal will have the form

$$g_{PM}(t) = A \cdot \cos\left[\omega_c t + k_p m(t)\right],$$

where A is a constant, ω_c is the carrier frequency, $m(t)$ is the message signal, and k_p is a parameter that specifies how much change in the angle occurs for every unit of change of $m(t)$. The phase and instantaneous frequency of this signal are

$$\theta_{PM}(t) = \omega_c t + k_p m(t),$$

So, the frequency of a PM signal is proportional to the derivative of the message signal.

Frequency Modulation (FM)

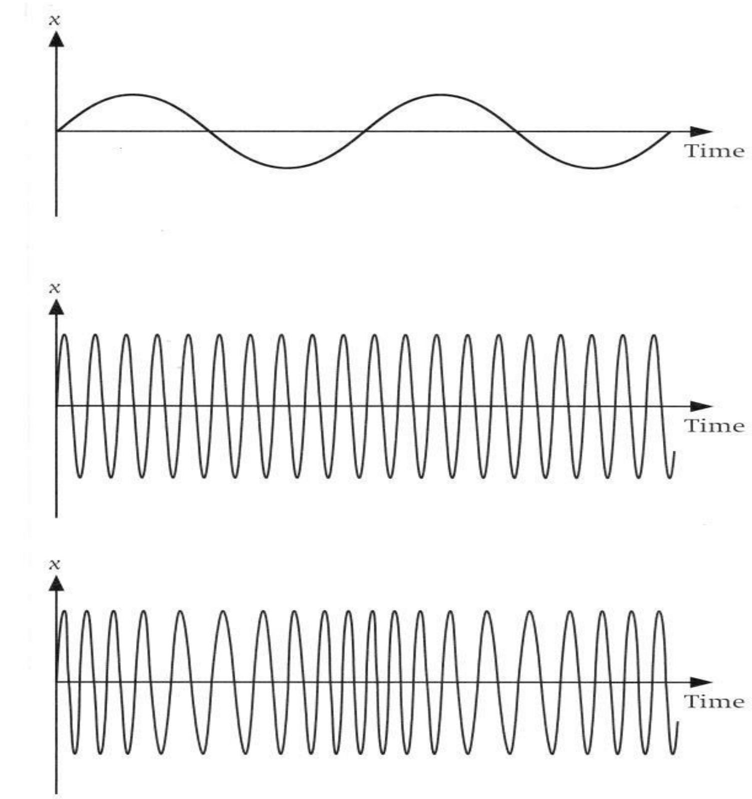
This type of modulation changes the frequency of the carrier (not the phase as in PM) directly with the message signal. The FM modulated signal is

$$g_{FM}(t) = A \cdot \cos\left[\omega_c t + k_f \int_{-\infty}^{t} m(\alpha)d\alpha\right],$$

where k_f is a parameter that specifies how much change in the frequency occurs for every unit change of $m(t)$. The phase and instantaneous frequency of this FM are

Frequency Modulation

In **F**requency **M**odulation (**FM**) the instantaneous value of the information signal controls the frequency of the carrier wave. This is illustrated in the following diagrams.



Notice that as the information signal increases, the frequency of the carrier increases, and as the information signal decreases, the frequency of the carrier decreases

The frequency f_i of the information signal controls the rate at which the carrier frequency increases and decreases. As with AM, f_i must be less than f_c. The amplitude of the carrier remains constant throughout this process.

When the information voltage reaches its maximum value then the change in frequency of the carrier will have also reached its maximum deviation above the nominal value. Similarly when the information reaches a minimum the carrier will be at its lowest frequency below the nominal carrier frequency value. When the information signal is zero, then no deviation of the carrier will occur.

The maximum change that can occur to the carrier from its base value f_c is called the frequency deviation, and is given the symbol Δf_c. This sets the dynamic range (i.e. voltage range) of the transmission. The dynamic range is the ratio of the largest and smallest analogue information signals that can be transmitted.

SINGLE-TONE FREQUENCY MODULATION

Time-Domain Expression

Since the FM wave is a nonlinear function of the modulating wave, the frequency modulation is a nonlinear process. The analysis of nonlinear process is the difficult task. In this section, we will study single-tone frequency modulation in detail to simplify the analysis and to get thorough understanding about FM.

Let us consider a single-tone sinusoidal message signal defined by

$$n(t) = A_n \cos(2\eta f_n t) \qquad (5.13)$$

The instantaneous frequency from Eq. (5.8) is then

$$f(t) = f_c + k_f A_n \cos(2\eta f_n t) = f_c + \Delta f \cos(2\eta f_n t) \qquad (5.14)$$

where

$$\Delta f = k_f A_n$$

$$\theta(t) = 2\pi f_c t + 2\pi k_f \int_0^t A_m \cos(2\pi f_m t)dt$$

$$= 2\pi f_c t + 2\pi k_f \frac{A_m}{2\pi f_m} \sin(2\pi f_m t)$$

$$= 2\pi f_c t + \frac{k_f A_m}{f_m} \sin(2\pi f_m t)$$

$$= 2\pi f_c t + \frac{\Delta f}{f_m} \sin(2\pi f_m t)$$

$$\therefore \theta(t) = 2\pi f_c t + \beta_f \sin(2\pi f_m t)$$

Where

$$\beta_f = \frac{\Delta f}{f_m} = \frac{k_f A_m}{f_m}$$

is the modulation index of the FM wave. Therefore, the single-tone FM wave is expressed by

$$s_{FM}(t) = A_c \cos[2\text{n}f_c t + \text{þ}_f \sin(2\text{n}f_n t)] \qquad (5.18)$$

This is the desired time-domain expression of the single-tone FM wave

Similarly, **single-tone phase modulated wave** may be determined from Eq.as

$$s_{PM}(t) = A_c \cos[2\text{n}f_c t + k_p A_n \cos(2\text{n}f_n t)]$$

$$\text{or, } s_{PM}(t) = A_c \cos[2\text{n}f_c t + \text{þ}_p \cos(2\text{n}f_n t)]$$

where $\text{þ}_p = k_p A_n$

is the modulation index of the single-tone phase modulated wave. The frequency deviation of the single-tone PM wave is

$$s_{FM}(t) = A_c \cos[2\pi f_c t + \beta_f \sin(2\pi f_m t)]$$

Spectral Analysis of Single-Tone FM Wave

The above Eq. can be rewritten as

$$s_{FM}(t) = \text{Re}\{A_c e^{j2\text{n}f_c t} e^{j\text{þ} \sin(2\text{n}f_n t)}\}$$

For simplicity, the modulation index of FM has been considered as þ instead of þ_f afterward. Since $\sin(2\text{n}f_n t)$ is periodic with fundamental period $T = 1/f_n$, the complex expontial $e^{j\text{þ} \sin(2\text{n}f_n t)}$ is also periodic with the same fundamental period. Therefore, this complex exponential can be expanded in Fourier series representation as

$$e^{j\beta \sin(2\pi f_m t)} = \sum_{n=-\infty}^{\infty} c_n e^{j2\pi n f_m t}$$

where the Fourier series coefficients c_n are obtained as

$$c_n = \frac{1}{T}\int_{-T/2}^{T/2} e^{j\beta\sin(2\pi f_m t)} e^{-j2\pi n f_m t} dt \tag{5.24}$$

Let $2\pi f_m t = x$, then Eq. (5.24) reduces to

$$c_n = \frac{1}{2\pi}\int_{-\pi}^{\pi} e^{j\beta\sin(x)} e^{-jnx} dx = \frac{1}{2\pi}\int_{-\pi}^{\pi} e^{j(\beta\sin(x) - nx)} dx \tag{5.25}$$

The integral on the right-hand side is known as the n[th] order Bessel function of the first kind and is denoted by $J_n(\beta)$. Therefore, $c_n = J_n(\beta)$ and Eq. (4.23) can be written as

$$e^{j\beta\sin(2\pi f_m t)} = \sum_{n=-\infty}^{\infty} J_n(\beta) e^{j2\pi n f_m t} \tag{5.26}$$

By substituting Eq. (5.26) in Eq. (5.22), we get

$$s_{FM}(t) = \mathrm{Re}\left\{A_c \sum_{n=-\infty}^{\infty} J_n(\beta) e^{j2\pi n f_m t}\, e^{j2\pi f_c t}\right\}$$
$$= A_c \sum_{n=-\infty}^{\infty} J_n(\beta)\cos[2\pi(f_c + n f_m)t] \tag{5.27}$$

Taking Fourier transform of Eq. (5.27), we get

$$S(f) = \frac{1}{2} A_c \sum_{n=-\infty}^{\infty} J_n(\beta)[\delta(f - f_c - n f_m) + \delta(f + f_c + n f_m)] \tag{5.28}$$

From the spectral analysis we see that there is a carrier component and a number of side-frequencies around the carrier frequency at $\pm n f_m$.

The Bessel function may be expanded in a power series given by

$$J_n(\beta) = \sum_{k=0}^{\infty} \frac{(-1)^k \left(\frac{1}{2}\beta\right)^{n+2k}}{k!\,(k+n)!} \tag{5.29}$$

Plots of Bessel function $J_n(\beta)$ versus modulation index β for $n = 0, 1, 2, 3, 4$ are shown in Figure 5.3.

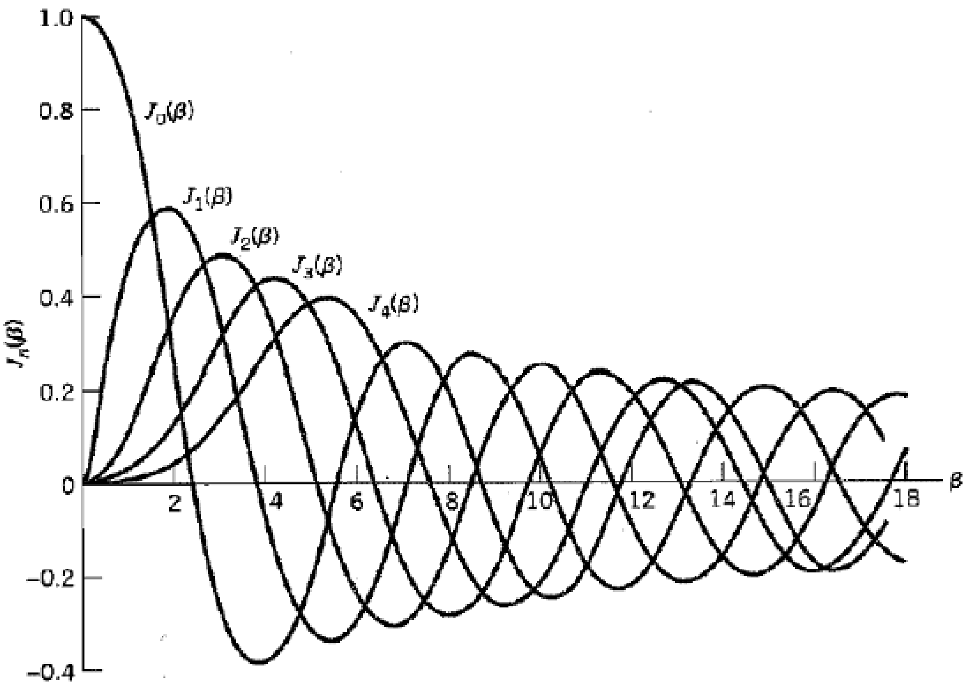


Figure 5.3 Plot of Bessel function as a function of modulation index.

Figure 5.3 shows that for any fixed value of n, the magnitude of $J_n(\beta)$ decreases as β increases. One property of Bessel function is that

$$J_{-n}(\beta) = \begin{cases} J_n(\beta), & n \text{ even} \\ -J_n(\beta), & n \text{ odd} \end{cases} \tag{5.30}$$

One more property of Bessel function is that

$$\sum_{n=-\infty}^{\infty} J_n^2(\beta) = 1 \tag{5.31}$$

(iii) The average power of the FM wave remains constant. To prove this, let us determine the average power of Eq. (5.27) which is equal to

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

Using Eq. (5.31), the average power P is now

$$P = \frac{1}{2} A_c^2$$

TRANSMISSION BANDWIDTH OF FM WAVE

The transmission bandwidth of an FM wave depends on the modulation index þ. The modulation index, on the other hand, depends on the modulating amplitude and modulating frequency. It is almost impossible to determine the exact bandwidth of the FM wave. Rather, we use a rule-of-thumb expression for determining the FM bandwidth.

For single-tone frequency modulation, the approximated bandwidth is determined by the expression

$$B = 2(\Delta f + f_m) = 2(\beta + 1)f_m = 2\Delta f\left(1 + \frac{1}{\beta}\right)$$

This expression is regarded as the Carson's rule. The FM bandwidth determined by this rule accommodates at least 98 % of the total power.

For an arbitrary message signal n(t) with bandwidth or maximum frequency W, the bandwidth of the corresponding FM wave may be determined by Carson's rule as

$$B = 2(\Delta f + W) = 2(D + 1)W = 2\Delta f\left(1 + \frac{1}{D}\right)$$

GENERATION OF FM WAVES

FM waves are normally generated by two methods: indirect method and direct method.

Indirect Method (Armstrong Method) of FM Generation

In this method, narrow-band FM wave is generated first by using phase modulator and then the wideband FM with desired frequency deviation is obtained by using frequency multipliers.

The above eq is the expression for narrow band FM wave

$$s(t) = A_c \cos(2\pi f_c t) - A_c \sin(2\pi f_c t)\emptyset(t)$$

$$\text{or, } s(t) = A_c \cos(2\pi f_c t) - 2\pi A_c k_f \sin(2\pi f_c t)\int_0^t m(t)dt$$

In this case

$$\cos[\emptyset(t)] \approx 1 \text{ and } \sin[\emptyset(t)] \approx \emptyset(t)$$

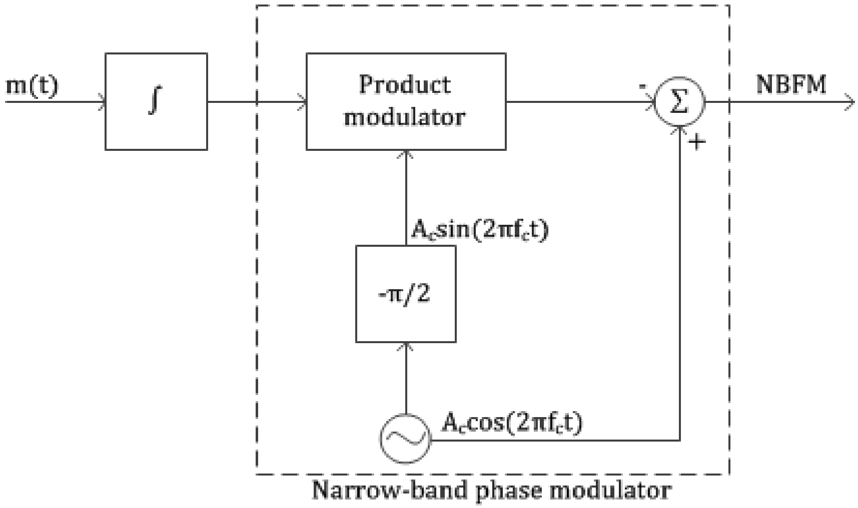


Fig: Narrowband FM Generator

The frequency deviation Δf is very small in narrow-band FM wave. To produce wideband FM, we have to increase the value of Δf to a desired level. This is achieved by means of one or multiple frequency multipliers. A frequency multiplier consists of a nonlinear device and a bandpass filter. The n^{th} order nonlinear device produces a dc component and n number of frequency modulated waves with carrier frequencies f_c, $2f_c$, … nf_c and frequency deviations Δf, $2\Delta f$, … $n\Delta f$, respectively. If we want an FM wave with frequency deviation of $6\Delta f$, then we may use a 6^{th} order nonlinear device or one 2^{nd} order and one 3^{rd} order nonlinear devices in cascade followed by a bandpass filter centered at $6f_c$. Normally, we may require very high value of frequency deviation. This automatically increases the carrier frequency by the same factor which may be higher than the required carrier frequency. We may shift the carrier frequency to the desired level by using mixer which does not change the frequency deviation.

The narrowband FM has some distortion due to the approximation made in deriving the expression of narrowband FM from the general expression. This produces some amplitude modulation in the narrowband FM which is removed by using a limiter in frequency multiplier

Direct Method of FM Generation

In this method, the instantaneous frequency f(t) of the carrier signal c(t) is varied directly with the instantaneous value of the modulating signal n(t). For this, an oscillator is used in which any one of the reactive components (either C or L) of the resonant network of the oscillator is varied linearly with n(t). We can use a varactor diode or a varicap as a voltage-variable capacitor whose capacitance solely depends on the reverse-bias voltage applied across it. To vary such capacitance linearly with n(t), we have to reverse-bias the diode by the fixed DC voltage and operate within a small linear portion of the capacitance-voltage characteristic curve. The unmodulated fixed capacitance C_0 is linearly varied by n(t) such that the resultant capacitance becomes

$$C(t) = C_0 - kn(t)$$

where the constant k is the sensitivity of the varactor diode (measured in capacitance per volt).

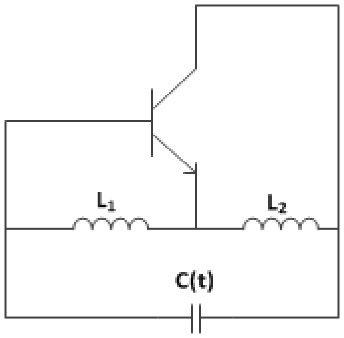


fig: Hartley oscillator for FM generation

The above figure shows the simplified diagram of the Hartley oscillator in which is implemented the above discussed scheme. The frequency of oscillation for such an oscillator is given

$$f(t) = \frac{1}{2\pi\sqrt{(L_1 + L_2)C(t)}}$$

$$f(t) = \frac{1}{2\pi\sqrt{(L_1 + L_2)(C_0 - km(t))}}$$

$$= \frac{1}{2\pi\sqrt{(L_1 + L_2)C_0}} \frac{1}{\sqrt{1 - \frac{km(t)}{C_0}}}$$

$$\text{or,} \qquad f(t) = f_c\left(1 - \frac{km(t)}{C_0}\right)^{-1/2}$$

where f_c is the unmodulated frequency of oscillation. Assuming,

$$\frac{km(t)}{C_0} \ll 1$$

we have from binomial expansion,

$$\left(1 - \frac{km(t)}{C_0}\right)^{-1/2} \approx 1 + \frac{km(t)}{2C_0}$$

$$f(t) \approx f_c\left(1 + \frac{km(t)}{2C_0}\right)$$

$$= f_c + \frac{kf_c m(t)}{2C_0}$$

$$\text{or, } f(t) = f_c + k_f m(t)$$

$$k_f = \frac{kf_c}{2C_0}$$

is the frequency sensitivity of the modulator. The Eq. (5.42) is the required expression for the instantaneous frequency of an FM wave. In this way, we can generate an FM wave by direct method.

Direct FM may be generated also by a device in which the inductance of the resonant circuit is linearly varied by a modulating signal n(t); in this case the modulating signal being the current.

The main advantage of the direct method is that it produces sufficiently high frequency deviation, thus requiring little frequency multiplication. But, it has poor frequency stability. A feedback scheme is used to stabilize the frequency in which the output frequency is compared with the constant frequency generated by highly stable crystal oscillator and the error signal is feedback to stabilize the frequency.

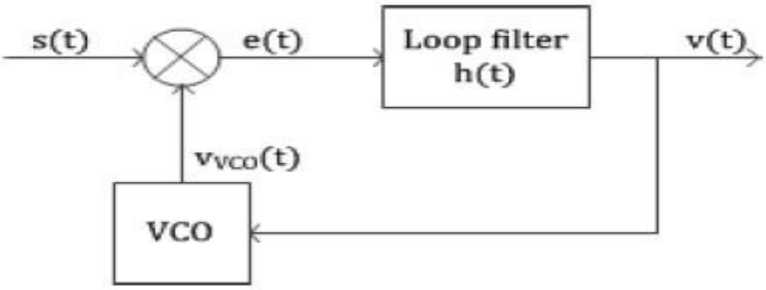
DETECTION OF FM WAVES

Phase-Locked Loop (PLL) as FM Demodulator

A PLL consists of a multiplier, a loop filter, and a VCO connected together to form a feedback loop as shown in Fig. 5.15. Let the input signal be an FM wave as defined by

$$s(t) = A_c \cos[2nf_c t + \emptyset_1(t)]$$

$$\emptyset_1(t) = 2\pi k_f \int_0^t m(t)dt$$



Let the VCO output be defined by

$$v_{VCO}(t) = A_v \sin[2nf_c t + \emptyset_2(t)]$$

$$\emptyset_2(t) = 2nk_v \int v(t)dt$$

Here k_v is the frequency sensitivity of the VCO measured in hertz per volt. The multiplication of $s(t)$ and $v_{VCO}(t)$ results

$$s(t)v_{VCO}(t) = A_c \cos[2\pi f_c t + \emptyset_1(t)]\, A_v \sin[2\pi f_c t + \emptyset_2(t)]$$

$$= \frac{A_c A_v}{2}\sin[4\pi f_c t + \emptyset_1(t) + \emptyset_2(t)] + \frac{A_c A_v}{2}\sin[\emptyset_2(t) - \emptyset_1(t)]$$

The high-frequency component is removed by the low-pass filtering of the loop filter. Therefore, the input signal to the loop filter can be considered as

$$e(t) = \frac{A_c A_v}{2} \sin[\emptyset_2(t) - \emptyset_1(t)]$$

The difference $\emptyset_2(t) - \emptyset_1(t) = \emptyset_e(t)$ constitutes the phase error. Let us assume that the PLL is in phase lock so that the phase error is very small. Then,

$$\sin[\emptyset_2(t) - \emptyset_1(t)] \approx \emptyset_2(t) - \emptyset_1(t)$$

$$\emptyset_e(t) = 2\pi k_v \int_0^t v(t)dt - \emptyset_1(t)$$

$$e(t) = \frac{A_c A_v}{2} \emptyset_e(t)$$

Differentiating Eq. (5.48) with respect to time, we get

$$\frac{d\emptyset_e(t)}{dt} = 2\pi k_v v(t) - \frac{d\emptyset_1(t)}{dt}$$

Since

$$v(t) = e(t) * h(t) = \frac{A_c A_v}{2} [\emptyset_e(t) * h(t)]$$

Eq. (5.50) becomes

$$\frac{d\emptyset_e(t)}{dt} = 2\pi k_v \frac{A_c A_v}{2} [\emptyset_e(t) * h(t)] - \frac{d\emptyset_1(t)}{dt}$$

$$or, \pi k_v A_c A_v [\emptyset_e(t) * h(t)] - \frac{d\emptyset_e(t)}{dt} = \frac{d\emptyset_1(t)}{dt}$$

Taking Fourier transform of Eq. (5.52), we get

$$\pi k_v A_c A_v \emptyset_e(f) H(f) - j2\pi f \emptyset_e(f) = j2\pi f \emptyset_1(f)$$

$$or, \emptyset_e(f) = \frac{j2\pi f}{\pi k_v A_c A_v H(f) - j2\pi f} \emptyset_1(f)$$

$$or, \emptyset_e(f) = \frac{1}{\frac{\pi k_v A_c A_v}{j2\pi f} H(f) - 1} \emptyset_1(f)$$

Fourier transform of Eq. (5.51) is

$$V(f) = \frac{A_c A_v}{2} \emptyset_e(f) H(f)$$

Substituting Eq. (5.53) into (5.54), we get

$$V(f) = \frac{A_c A_v}{2} \frac{1}{\frac{\pi k_v A_c A_v}{j2\pi f} H(f) - 1} \emptyset_1(f) H(f)$$

We design $H(f)$ such that

$$\left| \frac{\pi k_v A_c A_v}{j2\pi f} H(f) \right| \gg 1$$

in the frequency band $|f| < W$ of the message signal.

$$\therefore V(f) = \frac{A_c A_v}{2} \frac{1}{\frac{\pi k_v A_c A_v}{j2\pi f} H(f)} \emptyset_1(f) H(f)$$

$$or, V(f) = \frac{1}{2\pi k_v} j2\pi f \emptyset_1(f)$$

Taking inverse Fourier transform of Eq. (4.56), we get

$$v(t) = \frac{1}{2\pi k_v} \frac{d\emptyset_1(t)}{dt}$$

$$= \frac{1}{2\pi k_v} \frac{d}{dt} \left\{ 2\pi k_f \int_0^t m(t)dt \right\}$$

$$= \frac{1}{2\pi k_v} 2\pi k_f m(t)$$

$$\therefore v(t) = \frac{k_f}{k_v} m(t)$$

Since the control voltage of the VCO is proportional to the message signal, $v(t)$ is the demodulated signal.

We observe that the output of the loop filter with frequency response H(f) is the desired message signal. Hence the bandwidth of H(f) should be the same as the bandwidth W of the message signal. Consequently, the noise at the output of the loop filter is also limited to the bandwidth W. On the other hand, the output from the VCO is a wideband FM signal with an instantaneous frequency that follows the instantaneous frequency of the received FM signal.

Comparison of AM and FM:

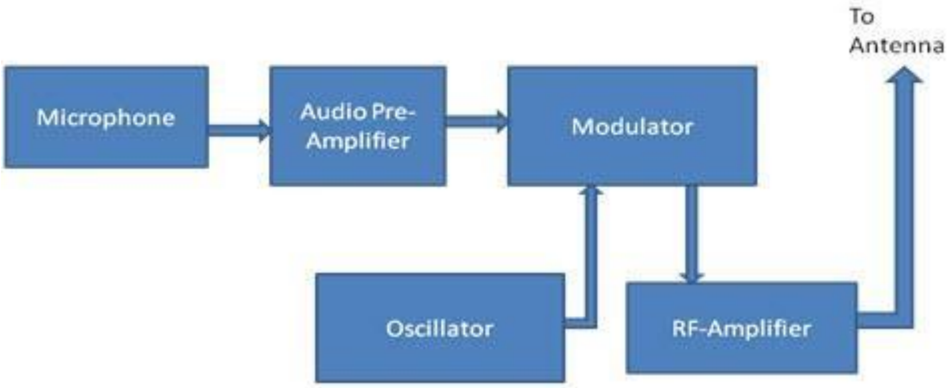
S.NO	AMPLITUDE MODULATION	FREQUENCY MODULATION
1.	Band width is very small which is one of the biggest advantage	It requires much wider channel (7 to 15 times) as compared to AM.
2.	The amplitude of AM signal varies depending on modulation index.	The amplitude of FM signal is constant and independent of depth of the modulation.
3.	Area of reception is large	The are of reception is small since it is limited to line of sight.
4.	Transmitters are relatively simple & cheap.	Transmitters are complex and hence expensive.
5.	The average power in modulated wave is greater than carrier power. This added power is provided by modulating source.	The average power in frequency modulated wave is same as contained in un-modulated wave.
6.	More susceptible to noise interference and has low signal to noise ratio, it is more difficult to eliminate effects of noise.	Noise can be easily minimized amplitude variations can be eliminated by using limiter.
7.	it is not possible to operate without interference.	it is possible to operate several independent transmitters on same frequency.
8.	The maximum value of modulation index = 1, other wise over-modulation would result in distortions.	No restriction is placed on modulation index.

MODULE III

Radio Transmitters &Receivers

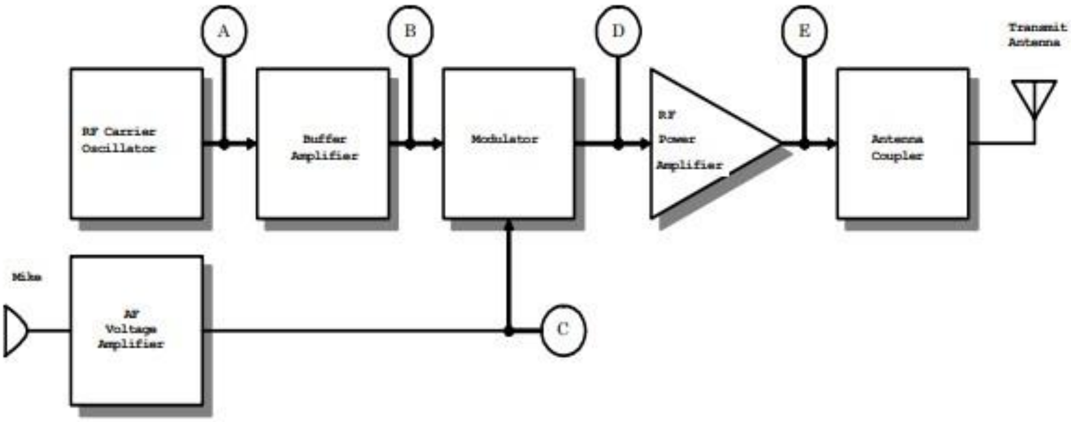
Transmitters:

Block diagram of AM Transmitter and FM Transmitter:

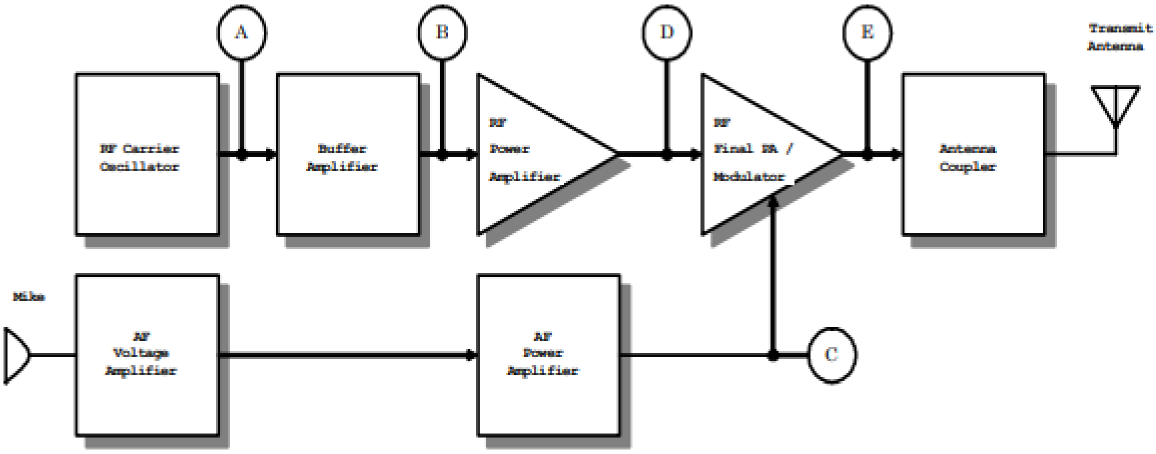


The FM transmitter is a single transistor circuit. In the telecommunication, the frequency modulation (FM)transfers the information by varying the frequency of carrier wave according to the message signal. Generally, the FM transmitter uses VHF radio frequencies of 87.5 to 108.0 MHz to transmit & receive the FM signal. This transmitter accomplishes the most excellent range with less power. The performance and working of the wireless audio transmitter circuit is depends on the induction coil & variable capacitor. This article will explain about the working of the FM transmitter circuit with its applications.

The FM transmitter is a low power transmitter and it uses FM waves for transmitting the sound, this transmitter transmits the audio signals through the carrier wave by the difference of frequency. The carrier wave frequency is equivalent to the audio signal of the amplitude and the FM transmitter produce VHF band of 88 to 108MHZ.Plese follow the below link for: Know all About Power Amplifiers for FM Transmitter



Low level



High Level

The RF signal is created in the RF carrier oscillator. At test point A the oscillator's output signal is present. The output of the carrier oscillator is a fairly small AC voltage, perhaps 200 to 400 mV RMS. The oscillator is a critical stage in any transmitter. It must produce an accurate and steady frequency. Every radio station is assigned a different carrier frequency. The dial (or display) of a receiver displays the carrier frequency. If the oscillator drifts off frequency, the receiver will be unable to receive the transmitted signal without being readjusted. Worse yet, if the oscillator drifts onto the frequency being used by another radio station, interference will occur. Two circuit techniques are commonly used to stabilize the oscillator, buffering and voltage regulation.

The buffer amplifier has something to do with buffering or protecting the oscillator. An oscillator is a little like an engine (with the speed of the engine being similar to the oscillator's frequency). If the load on the engine is increased (the engine is asked to do more work), the engine will respond by slowing down. An oscillator acts in a very similar fashion. If the current drawn from the oscillator's output is increased or decreased, the oscillator may speed up or slow down slightly.

Buffer amplifier is a relatively low-gain amplifier that follows the oscillator. It has a constant input impedance (resistance). Therefore, it always draws the same amount of current from the oscillator. This helps to prevent "pulling" of the oscillator frequency. The buffer amplifier is needed because of what's happening "downstream" of the oscillator. Right after this stage is the modulator. Because the modulator is a nonlinear amplifier, it may not have a constant input resistance -- especially when information is passing into it. But since there is a buffer amplifier between the oscillator and modulator, the oscillator sees a steady load resistance, regardless of what the modulator stage is doing.

Voltage Regulation: An oscillator can also be pulled off frequency if its power supply voltage isn't held constant. In most transmitters, the supply voltage to the oscillator is regulated at a constant value. The regulated voltage value is often between 5 and 9 volts; zener diodes and three-terminal regulator ICs are commonly used voltage regulators. Voltage regulation is especially important when a transmitter is being powered by batteries or an automobile's electrical system. As a battery discharges, its terminal voltage falls. The DC supply voltage in a car can be anywhere between 12 and 16 volts, depending on engine RPM and other electrical load conditions within the vehicle.

Modulator: The stabilized RF carrier signal feeds one input of the modulator stage. The modulator is a variable-gain (nonlinear) amplifier. To work, it must have an RF carrier signal and an AF information signal. In a low-level transmitter, the power levels are low in the oscillator, buffer, and modulator stages; typically, the modulator output is around 10 mW (700 mV RMS into 50 ohms) or less.

AF Voltage Amplifier: In order for the modulator to function, it needs an information signal. A microphone is one way of developing the intelligence signal, however, it only produces a few millivolts of signal. This simply isn't enough to operate the modulator, so a voltage amplifier is used to boost the microphone's signal. The signal level at the output of the AF voltage amplifier is usually at least 1 volt RMS; it is highly dependent upon the transmitter's design. Notice that the AF amplifier in the transmitter is only providing a voltage gain, and not necessarily a current gain for the microphone's signal. The power levels are quite small at the output of this amplifier; a few mW at best.

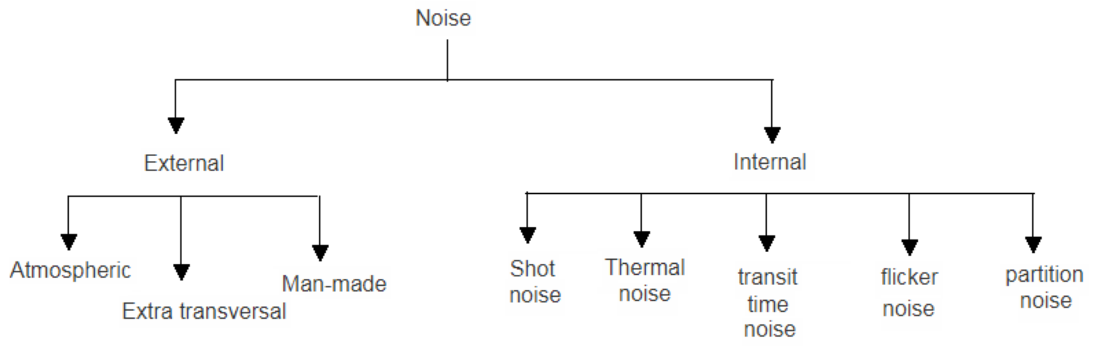
Types of noise:

Noise is an unwanted electrical disturbance which gives rise to audioble or visual disturbance in communication systems and errors in digital system, the noise gets super impaired on the signal and makes it impossible to separate can be signal from noise.

Noise can be divided into two categories:

1] External Noise.

2] Internal Noise.



Thermal noise:

Thermal noise is produced by the motion of free electrons in a resistance due to temperature. It is generated even when the resistance is not connected to a circuit but is due to the random fluctuations in charge at either end of the resistance. Thermal noise is often called Johnson noise. The noise power in thermal noise is constant per unit of bandwidth across the usable electronic spectrum. Because of this, it is a form of white noise. The maximum noise power available from a thermal noise source is given by the equation:

Pn = kTB

Pn= noise power,

k= Boltzmann's constant, 1.38x10 -23 J/K

T= absolute temperature B=bandwidth

Shot noise

The flow of current is not continuous in a circuit but rather is associated with random variations in the number of charge carriers passing some voltage boundary. Charge is limited by the smallest unit of charge available-that of the charge on an electron. Shot noise, like thermal noise, has the same power per unit of bandwidth; hence it is a type of white noise. When amplified, it sounds something like lead shot raining on a metal roof-hence the term shot noise. Shot noise is given in terms of a current and is found from the equation

ish = 2eIDCB ish=rms shot noise current

e=electron charge, 1.6x10 -19C

Shot noise occurs in virtually all active devices. The shot noise depends on a number of variables, so it is convenient to represent noise sources by assuming a noise –free device with external noise sources connected to it. One way of specifying the random noise contribution is an active device is to assign an "effective noise temperature" to the input. This temperature, labeled Te, is added to the effective input noise temperature Tin to obtain an equivalent operating temperature of an active device. The noise temperature of a device does not mean that the device is actual operating at that temperature; rather, it gives an equivalent temperature of a thermal source with the same noise power. The output noise power of a transistor can be written as:

Pn = Gk(Te + Tin)B

G=transistor gain

Te=effective noise temperature, K

Tin=effective input noise temperature, K

EXTRINSIC NOISE

Extrinsic noise is induced from an external source and can cause unsatisfactory operation of a circuit (interference). The source of noise may be from another circuit on the same circuit board (often referred to as cross talk), or it may be external to the equipment. For interference to occur, there needs to be a source of noise and a means of coupling it into the circuit. The external source may come from conduction, capacitive coupling, magnetic coupling, or radiation. To reduce the effects of interference, the interference can be suppressed at the source, the source can be isolated by shielding or filtering, the coupling path can be reduced, or the receiving circuit can be made less sensitive to noise.

White noise

One of the very important random processes is the *white noise* process. Noises in many practical situations are approximated by the white noise process. Most importantly, the white noise plays an important role in modelling of WSS signals.

A white noise process $\{W(t)\}$ is a random process that has constant power spectral density at all frequencies. Thus

$$S_W(\omega) = \frac{N_0}{2} \qquad -\infty < \omega < \infty$$

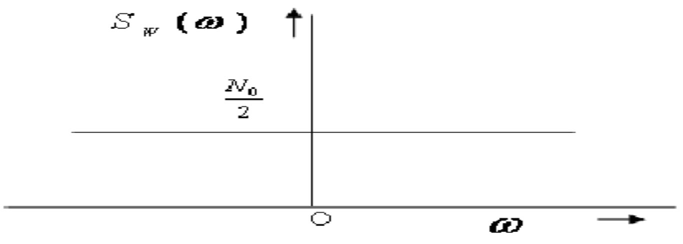
where N_0 is a real constant and called the *intensity* of the white noise. The corresponding autocorrelation function is given by

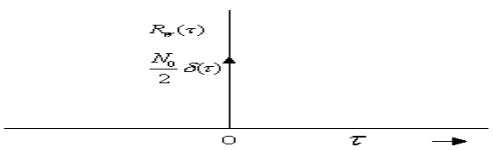
$R_W(\tau) = \frac{N}{2}\delta(\tau)$ where $\delta(\tau)$ is the Dirac delta.

The average power of white noise

$$P_{avg} = EW^2(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty}\frac{N}{2}d\omega \rightarrow \infty$$

The autocorrelation function and the PSD of a white noise process is shown in Figure 1 below.





NARROWBAND NOISE (NBN)

In most communication systems, we are often dealing with band-pass filtering of signals. Wideband noise will be shaped into band limited noise. If the bandwidth of the band limited noise is relatively small compared to the carrier frequency, we refer to this as *narrowband noise*.the narrowband noise is expressed as as

$$n(t) = x(t)\cos 2\pi f_C t - y(t)\sin 2\pi f_C t$$

where f_C is the carrier frequency within the band occupied by the noise. $x(t)$ and $y(t)$ are known as the *quadrature components* of the noise $n(t)$. The Hibert transform of $n(t)$ is

Proof.

The Fourier transform of $n(t)$ is

$$N(f) = \frac{1}{2}X(f-f_C) + \frac{1}{2}X(f+f_C) + \frac{1}{2}jY(f-f_C) - \frac{1}{2}jY(f+f_C)$$

Let $\hat{N}(f)$ be the Fourier transform of $\hat{n}(t)$. In the frequency domain, $\hat{N}(f) = N(f)[-j\ \mathrm{sgn}(f)]$. We simply multiply all positive frequency components of $N(f)$ by $-j$ and all negative frequency components of $N(f)$ by j. Thus

$$\hat{n}(t) = H[n(t)] = x(t)\sin 2\pi f_C t + y(t)\cos 2\pi f_C t$$

$$\hat{N}(f) = -j\frac{1}{2}X(f-f_C) + j\frac{1}{2}X(f+f_C) - j\frac{1}{2}jY(f-f_C) - j\frac{1}{2}jY(f+f_C)$$
$$\hat{N}(f) = -j\frac{1}{2}X(f-f_C) + j\frac{1}{2}X(f+f_C) + \frac{1}{2}Y(f-f_C) + \frac{1}{2}Y(f+f_C)$$

and the inverse Fourier transform of $\hat{N}(f)$ is

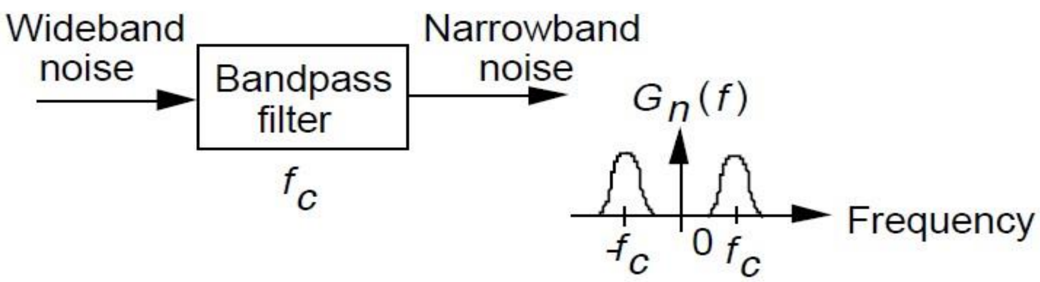
$$\hat{n}(t) = x(t)\sin 2\pi f_C t + y(t)\cos 2\pi f_C t$$

The quadrature components $x(t)$ and $y(t)$ can now be derived from equations

$x(t) = n(t)\mathrm{co}2\pi f_C t + n\hat{\ }(t)\sin 2\pi f_C t$ \

and

$y(t) = n(t)\cos 2\pi f_C t - n\hat{\ }(t)\sin 2\pi f_C t$



Definition of Noise Factor

Noise factor of a network is defined as the ratio of signal-to-noise ratio (SNR) at the input to the SNR at the output.

F=SNRi/SNRo

Where SNRi is the signal-to-noise ratio at the input of the network and SNRo is the signal-to-noise ratio at the output of the network.

This can also be specified as noise figure (NF) in logarithmic units as:

NF=SNRi-SNR0

From Equation 8, it can be inferred that NF is a measure of degradation of the SNR from input to output. We can rewrite Equation 7 as division of respective signal power density to noise power density at input and output as:

F=Si/Ni/So/No

Where: Si is the signal power density at the input of the network.

Ni is the noise power density at the input of the network.

So is the signal power density at the output of the network.

No is the noise power density at the output of the network.

To understand further, it is assumed that a signal with a power density of Si is input to the system which amplifies it by a power gain of G and adds a certain amount of noise Nx to the input.

Noise Figure of cascade of two stages

Consider a system with cascade of two gain stages having gains of G_1, G_2 and with noise figure of N_{F1} and N_{F2} respectively.

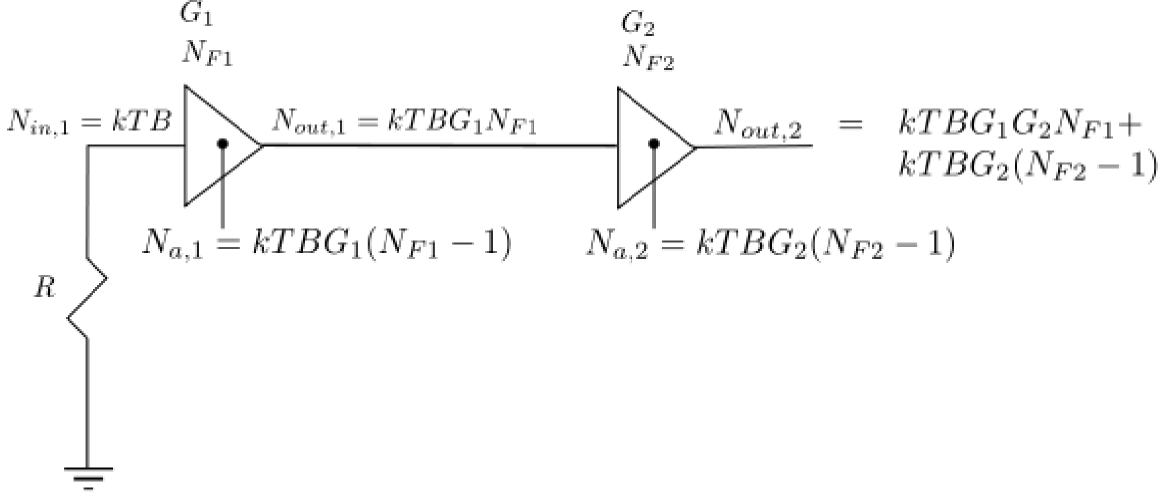


Figure : Cascaded gain stages with noise power output
The noise power at the output of the first stage is the sum of
a) Noise power at the input terminal of first gain stage $N_{in,1}=kTB$ which is amplified by gain G_1.

b) Noise power generated by the first gain stage $N_{a,1}=kTBG_1(N_{F1}-1)$

Summing them up, the noise seen at the output of first gain stage is,

$$\begin{aligned} N_{out,1} &= kTBG_1+kTBG_1(N_{F1}-1) \\ &= kTBG_1N_{F1} \end{aligned}$$

*** Note :** The sum of powers is possible because it is assumed that the input noise and the noise added by the system are uncorrelated i.e.

$$\begin{aligned} E\{(x+y)^2\} &= E\{x^2\}+E\{y^2\}+2E\{xy\} \\ &= E\{x^2\}+E\{y^2\} \end{aligned}$$

, where

$E\{xy\}=0$ when x and y are uncorrelated.

Similarly the noise power seen at the output of the second stage is the sum of

a) Noise power at the input terminal of second gain stage $N_{out,1}=kTBG_1N_{F1}$ which is amplified by gain G_2.

b) Noise power generated by the second gain stage $N_{a,2}=kTBG_2(N_{F2}-1)$

Adding them up, the noise seen at the output of second gain stage is,

.

Equivalent Noise Figure of 2 gain stages

Summarizing, the equivalent noise figure of the two cascaded stage is,

$$N_{F12} = \left(N_{F1} + \frac{N_{F2} - 1}{G_1} \right)$$

and the equivalent gain is,

$$G_{12} = G_1 G_2$$

Note : All the expressions are in linear terms

Equivalent Noise Figure of n gain stages

Extending this to cascade of n stages, the equivalent noise figure is,

$$N_{F12...n} = N_{F1} + \frac{N_{F2} - 1}{G_1} + \frac{N_{F3} - 1}{G_1 G_2} + ... + \frac{N_{Fn} - 1}{G_1 G_2 ... G_{n-1}}$$

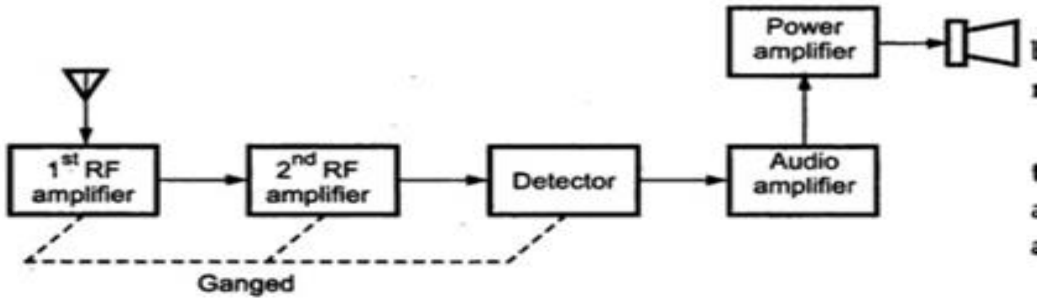
RADIO RECEIVERS:

Introduction to Radio Receivers: In radio communications, a **radio receiver** (**receiver** or simply **radio**) is an electronic device that receives **radio** waves and converts the information carried by them to a usable form.

Types of Receivers:

The TRF (Tuned Radio Frequency) Receiver and Superheterodyne Receiver are the two main configurations of the receivers, they have real practical or commercial significance. Most of the present day receivers use superheterodyne configuration. But the TRF receivers are simple and easy to understand.

Tuned Radio Frequency Receiver:



power amplifier. The RF amplifier stages placed between the antenna and detector are used to increase the strength of the received signal before it is applied to the detector. These RF amplifiers are tuned to fix frequency, amplify the desired band of frequencies. Therefore, they provide amplification for selected band of frequencies and rejection for all others. As selection and amplification process is carried out in two or three stages and each stage must amplify the same band of frequencies, the ganged tuning is provided.

The amplified signal is then demodulated using detector to recover the modulating signal. The recovered signal is amplified further by the audio amplifier followed by power amplifier which provides sufficient gain to operate a loudspeaker. The TRF receivers suffered from number of annoying problems. These are listed in the next section.

Problems in TRF Receivers:

1. Tracking of Tuned Circuit

In a receiver, tuned circuits are made variable so that they can be set to the frequency of the desired signal. In most of the receivers, the capacitors in the tuned circuits are made variable. These capacitors are 'ganged' between the stages so that they all can be changed simultaneously when the tuning knob is rotated. To have perfect tuning the capacitor values between the stages must be exactly same but this is not the case. The differences in the capacitors cause the resonant frequency of each tuned circuit to be slightly different.

3. Variable Bandwidth

TRF receivers suffer from a variation in bandwidth over the tuning range. Consider a medium wave receiver required to tune over 535 kHz to 1640 kHz and it provides the necessary bandwidth of 10 kHz at 535 kHz. Let us calculate Q of this circuit.

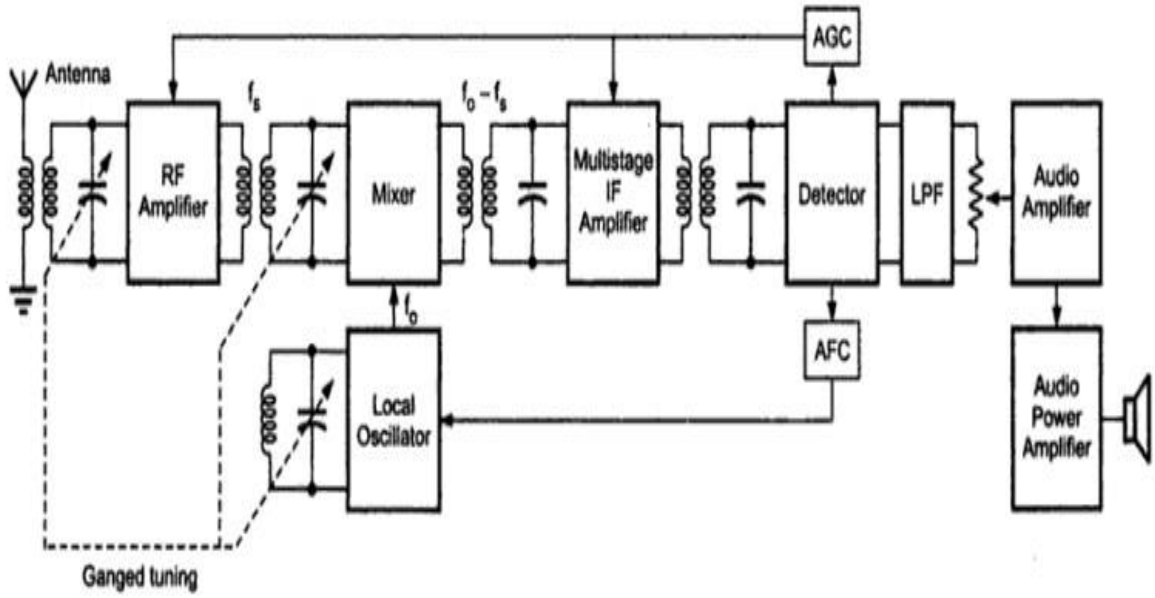
$$Q = \frac{f}{\text{Bandwidth}} = \frac{535\,\text{kHz}}{10\,\text{kHz}} = 53.5$$

Now consider the frequency at the other end of the broadcast band, i.e. 1640 kHz. At 1640 kHz, Q of the coil should be 164 (1640 kHz / 10 kHz). However, in practice due to various losses depending on frequency, we will not set so large increase in Q. Let us assume that at 1640 kHz frequency Q is increased to value 100 instead of 164. With this Q of the tuned circuit bandwidth can be calculated as follows

$$\text{Bandwidth} = \frac{f}{Q} = \frac{1640\,\text{kHz}}{100} = 16.4\ \text{kHz}$$

We know, necessary bandwidth is 10 kHz. This increase in bandwidth of tuned circuit, pick up the adjacent stations along with station it is tuned for, providing insufficient adjacent frequency rejection. In other words we can say that in TRF receivers the bandwidth of the tuned circuit varies over the frequency range, resulting in poor selectivity of the receiver.

Because of the problems of tracking, instability and bandwidth variation, the TRF receivers have almost been replaced by superheterodyne receivers.



Superheterodyne Receivers

To solve basic problems of TRF receivers, in these receivers, first all the incoming RF frequencies are converted to a fix lower frequency called **intermediate frequency** (IF). Then this fix intermediate frequency is amplified and detected to reproduce the original information. Since the characteristics of the IF amplifier are independent of the frequency to which the receiver is tuned, the selectivity and sensitivity of superheterodyne receivers are fairly uniform throughout its tuning range.

Mixer circuit is used to produce the frequency translation of the incoming signal down to the IF. The incoming signals are mixed with the local oscillator frequency signal in such a way that a constant frequency difference is maintained between the local oscillator and the incoming signals. This is achieved by using ganged tuning capacitors.

Fig. 2 shows the block diagram of superheterodyne receiver. As shown in the Fig. 2 antenna picks up the weak radio signal and feeds it to the RF amplifier. The RF amplifier provides some initial gain and selectivity. The output of the RF amplifier is applied to the input of the mixer. The mixer also receives an input from local oscillator.

The output of the mixer circuit is difference frequency ($f_o - f_s$) commonly known as IF (Intermediate Frequency). The signal at this intermediate frequency contains the same modulation as the original carrier. This signal is amplified by one or more IF amplifier stages, and most of the receiver gain is obtained in these IF stages.

The highly amplified IF signal is applied to detector circuits to recover the original modulating information. Finally, the output of detector circuit is fed to audio and power amplifier which provides a sufficient gain to operate a speaker.

Another important circuit in the superheterodyne receiver are AGC and AFC circuit. AGC is used to maintain a constant output voltage level over a wide range of RF input signal levels.

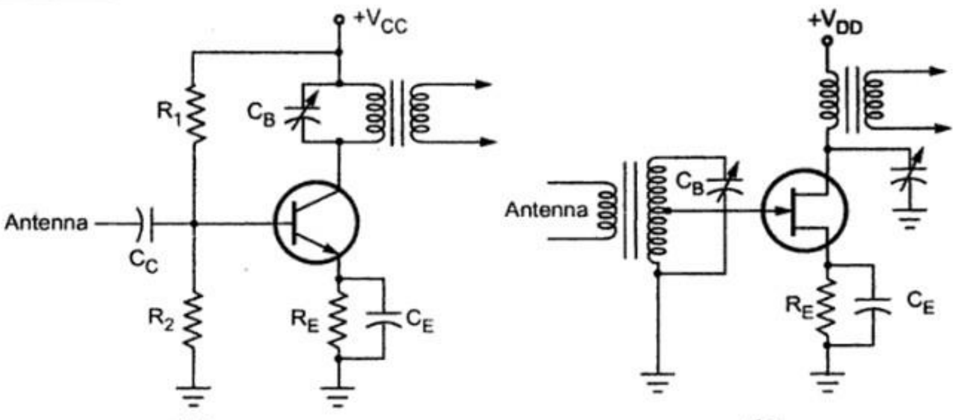
It derives the dc bias voltage from the output of detector which is proportional to the amplitude of the received signal. This dc bias voltage is feedback to the IF amplifiers, and sometimes to the RF amplifier, to control the gain of the receiver. As a result, it provides a constant output voltage level over a wide range of RF input signal levels. AFC circuit generates AFC signal which is used to adjust and stabilize the frequency of the local oscillator.

Blocks in Super heterodyne Receiver:

- **Basic principle**
  - Mixing
  - Intermediate frequency of 455 KHz
  - Ganged tuning
- **RF section**
  - Tuning circuits – reject interference and reduce noise figure
  - Wide band RF amplifier
- **Local Oscillator**
  - 995 KHz to 2105 KHz
  - Tracking
- **IF amplifier**
  - Very narrow band width Class A amplifier – selects 455 KHz only
  - Provides much of the gain
  - Double tuned circuits
- **Detector**
  - RF is filtered to ground

1. RF Amplifier:

RF amplifier provides initial gain and selectivity. Fig. 4 shows the RF amplifier circuits. It is a tuned circuit followed by an amplifier The RF amplifier is usually a simple class A circuit. A typical bipolar circuit is shown in Fig. 4. (a) , and a typical FET circuit is shown in Fig. 4. (b).



The values of resistors R_1 and R_2 in the bipolar circuit are adjusted such that the amplifier works as class A amplifier. The antenna is connected through coupling capacitor to the base of the transistor. This makes the circuit very broad band as the transistor will amplify virtually any signal picked up by the antenna. However the collector is tuned with a parallel resonant circuit to provide the initial selectivity for the mixer input.

The FET circuit shown in Fig. 4 (b) is more effective than the transistor circuit. Their high input impedance minimizes the loading on tuned circuits, thereby permitting the Q of the circuit to be higher and selectivity to be sharper.

The receiver having an RF amplifier stage has following advantages :

1. It provides greater gain, i.e. better sensitivity.
2. It improves image-frequency rejection.
3. It improves signal to noise ratio.
4. It improves rejection of adjacent unwanted signals, providing better selectivity.
5. It provides better coupling of the receiver to the antenna.
6. It prevents spurious frequencies from entering the mixer and heterodyning there to produce an interfering frequency equal to the IF from the desired signal.
7. It also prevents reradiation of the local oscillator through the antenna of the receiver.

1. Mixer

The frequency converter is a nonlinear resistance having two sets of input terminals and one set of output terminal. The two inputs to the frequency converter are the input signal along with any modulation and the input from a local oscillator (LO). The output contains several frequencies including the difference between the input frequencies. The difference frequency is called intermediate frequency and output circuit of the mixer is tuned for the intermediate frequency.

Separately Excited Mixer:

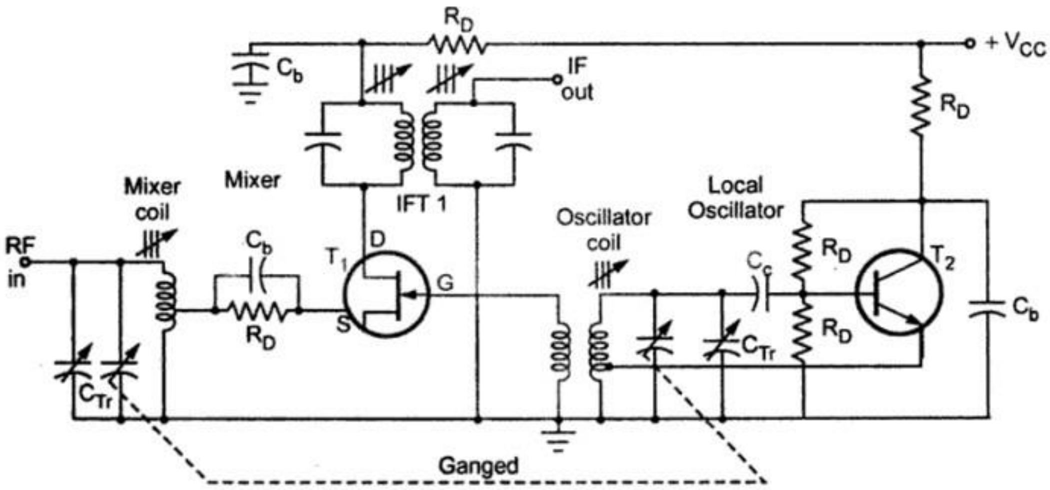


Fig.5 Separately Excited FET Mixer

Fig. 5 shows the separately excited mixer using FET. Here, one device acts as a mixer while the other supplies the necessary oscillations. The bipolar transistor T_2, forms the Hartley oscillator circuit. It oscillates with local frequency (f_o). FET T_1, is a mixer, whose gate is fed with the output of local oscillator and its bias is adjusted such that it operates in a nonlinear portion of its characteristic. The local oscillator varies the gate bias of the FET to vary its transconductance in a nonlinear manner, resulting intermediate frequency (IF) at the output. The output is taken through double tuned transformer in the drain of the mixer and fed to the IF amplifier. The ganged tuning capacitor allows simultaneous tuning of mixer and local oscillator.

The C_{Tr}, a small trimmer capacitors across each of the tuning capacitors are used for fine adjustments.

Self Excited Mixer:

It is possible to combine the function of the mixer and local oscillator in one circuit. The circuit is commonly known as self excited mixer. Fig. 6 shows self excited bipolar transistor mixer. The circuit oscillates and the transconductance of the transistor is varied

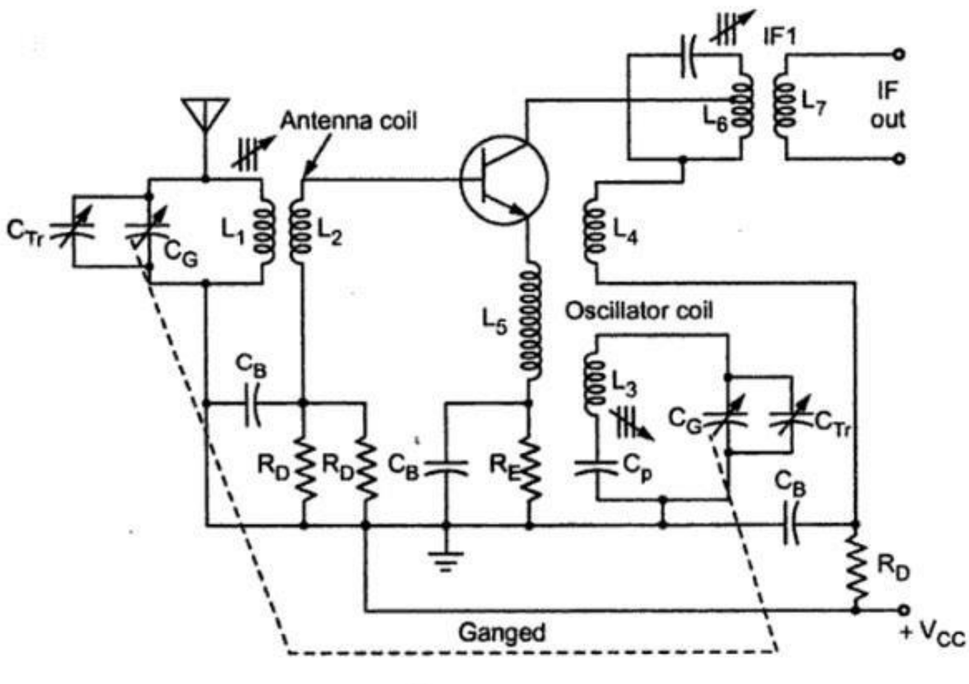


Fig.6. Self Excited Mixer

2. Tracking

The superheterodyne receiver has number of tunable circuits which must all be tuned correctly if any given station is to be received. The ganged tuning is employed to do this work, which mechanically couples all tuning circuits so that only one tuning control or dial is required. Usually, there are three tuned circuits : Antenna or RF tuned circuit, mixer tuned circuit and local oscillator tuned circuit. All these circuits must be tuned to get proper RF input and to get IF frequency at the output of mixer. The process of tuning circuits to get the desired output is called **Tracking**. Any error that exists in the frequency difference will result in an incorrect frequency being fed to the IF amplifier. Such errors are known as **'Tracking Errors'** and these must be avoided.

To avoid tracking errors standard capacitors are not used, and ganged capacitors with identical sections are used. A different value of inductance and special extra capacitors called trimmers and padders are used to adjust the capacitance of the oscillator to the proper range. There are three common methods used for tracking. These are

- Padder tracking
- Trimmer tracking

1. Local Oscillator

In shortwave broadcasting, the operating limit for receivers is 36 MHz. For such operating limit local oscillators such as Armstrong, Hartley, Colpitts, Clapp or ultra-audion are used. The Colpitts, Clapp and ultra-audion oscillators are used at the top of the operating limit, whereas Hartley oscillator is used for frequencies below 120 MHz. All these oscillators are LC oscillators and each employs only one tuned circuit to defermine its frequency. When higher frequency stability of local oscillator is required, the circuits like AFC (Automatic Frequency Control) are used.

2. IF Amplifier

IF amplifiers are tuned voltage amplifiers tuned for the fixed frequency. Its important function is to amplify only tuned frequency signal and reject all others. As we know, most of the receiver gain is provided by the IF amplifiers, to obtain required gain, usually two or more stages of IF amplifiers are required.

Fig. 7 shows the two stage IF amplifier. Two stages are transformer coupled and all IF transformers are single tuned, i.e. tuned for single frequency.

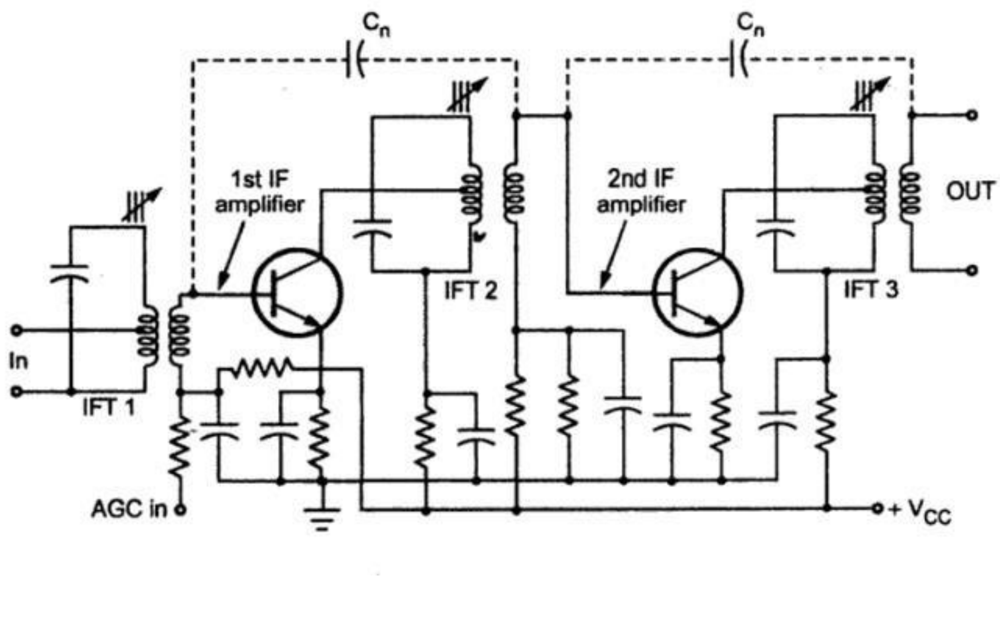


Fig.7 Two Stage IF Amplifier

Choice of Intermediate Frequency

Selection of the intermediate frequency depends on various factors. While choosing the intermediate frequency it is necessary to consider following factors.

1. Very high intermediate frequency will result in poor selectivity and poor adjacent channel rejection.
2. A high value of intermediate frequency increases tracking difficulties.
3. At low values of intermediate frequency, image frequency rejection is poor.
4. At very low values of intermediate frequency, selectivity is too sharp. Cutting off the sidebands.
5. At very low IF, the frequency stability of the local oscillator must be correspondingly high because any frequency drift is now a larger proportion of the low IF than of a high IF.
6. The IF must not fall in the tuning range of the receiver, otherwise instability will occur and heterodyne whistles will be heard, making it impossible to tune to the frequency band immediately adjacent to the intermediate frequency.

With the above considerations the standard broadcast AM receivers [tuning to 540 to 1650 kHz] use an IF within the 438 kHz to 465 kHz range. The 465 kHz IF is most commonly used.

6 Automatic Gain Control

Automatic Gain Control is a system by means of which the overall gain of a radio receiver is varied automatically with the variations in the strength of the receiver signal, to maintain the output substantially constant. AGC circuitry derives the dc bias voltage from the output of the detector. It applies this derived dc bias voltage to a selected number of RF, IF and mixer stages to control their gains. When the average signal level increases, the size of the AGC bias increases, and the gain of the controlled stages decreases. When there is no signal, there is a minimum AGC bias, and the amplifiers produce maximum gain. There are two types of AGC circuits in use : Simple AGC and Delayed AGC

Simple AGC

In simple AGC receivers the AGC bias starts to increase as soon as the received signal level exceeds the background noise level. As a result receiver gain starts falling down, reducing the sensitivity of the receiver.

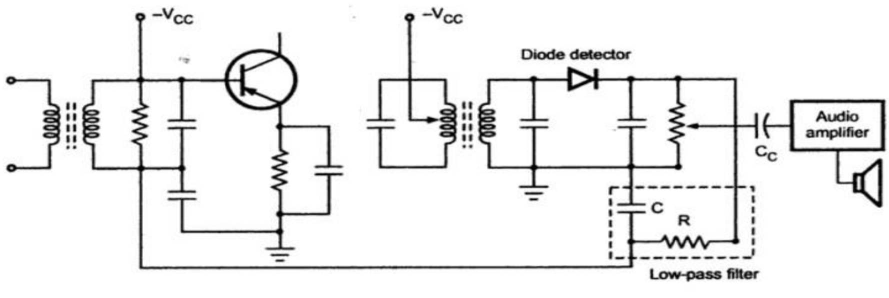
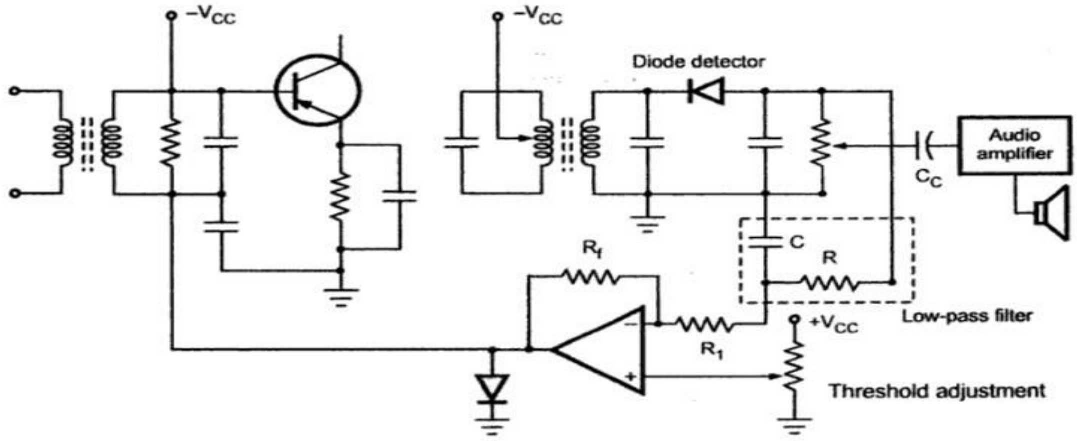


Fig. 8 shows the simple AGC circuit. In this circuit, dc bias produced by half wave rectifier as a AM detector, is used to control the gain of RF or IF amplifier. Before application of this voltage to the base of the RF and / or IF stage amplifier the audio signal is removed by the lowpass filter. The time constant of the filter is kept at least 10 times longer than the period of the lowest modulation frequency received. If the time constant is kept longer, it will give better filtering, but it will cause an annoying delay in the application of the AGC control when tuning from one signal to another. The recovered signal is then passed through C_C to remove the dc. The resulting ac signal is further amplified and applied to the loudspeaker.

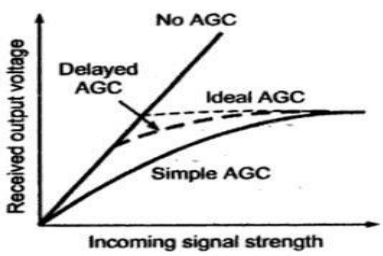
Delayed AGC

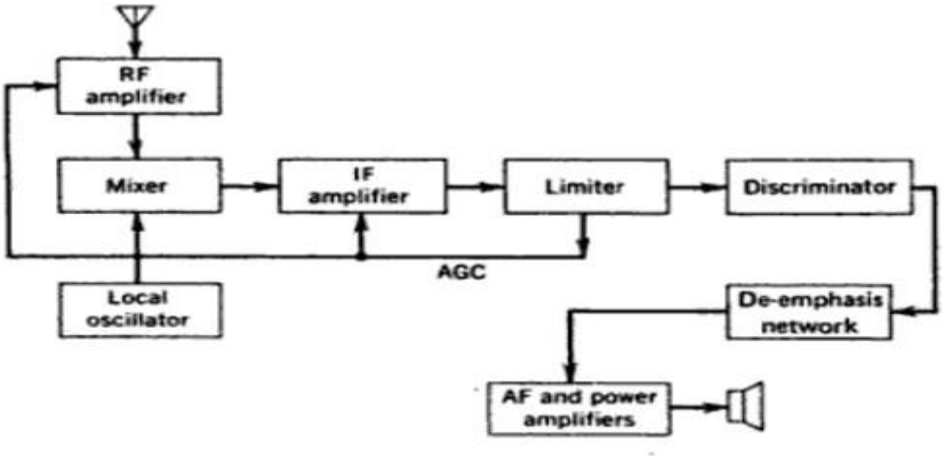
Simple AGC is clearly an improvement over no AGC at all. Unfortunately, in simple AGC circuit, the unwanted weak signals(noise signals) are amplified with high gain. To



avoid this, in delayed AGC circuits, AGC bias is not applied to amplifiers until signal strength has reached a predetermined level, after which AGC bias is applied as with simple AGC, but more strongly.

Here, AGC output is applied to the difference amplifier. It gives negative dc AGC only when AGC output generated by diode detector is above certain dc threshold voltage. This threshold voltage can be adjusted by adjusting the voltage at the positive input of the operational amplifier.





The FM receiver is a superheterodyne receiver, and the block diagram of Figure 11 shows just how similar it is to an AM receiver. The basic differences are as follows:

1. Generally much higher operating frequencies in FM
2. Need for limiting and de-emphasis in FM
3. Totally different methods of demodulation
4. Different methods of obtaining AGC

Comparison of FM and AM

Sr. No.	FM	AM
1	FM receivers are immune to noise	AM receivers are not immune to noise
2	It is possible to decrease noise by increasing deviation	This feature is absent in AM
3	Bandwidth is higher and depends on modulation index	Bandwidth is lower compared to AM but independent of modulation index
4	FM transmission and reception equipment are more complex	FM transmission and reception equipment are less complex
5	All transmitted power is useful	Carrier power and one sideband power is useless

Amplitude Limiter:

In order to make full use of the advantages offered by FM, a demodulator must be preceded by an amplitude limiter, on the grounds that any amplitude changes in the signal fed to the FM demodulator are spurious

They must therefore be removed if distortion is to be avoided. The point is significant, since most FM demodulators react to amplitude changes as well as frequency changes. The limiter is a form of clipping device, a circuit whose output tends to remain constant despite changes in the input signal. Most limiters behave in this fashion, provided that the input voltage remains within a certain range. The common type of limiter uses two separate electrical effects to provide a relatively constant output. There are leak-type bias and early (collector) saturation.

Operation of the amplitude limiter Figure shows a typical FET amplitude limiter. Examination of the dc conditions shows that the drain supply voltage has been dropped through resistor R_D. Also, the bias on the gate is leak-type bias supplied by the parallel $R_g - C_g$ combination. Finally, the FET is shown neutralized by means of capacitor C_N, in consideration of the high frequency of operation.

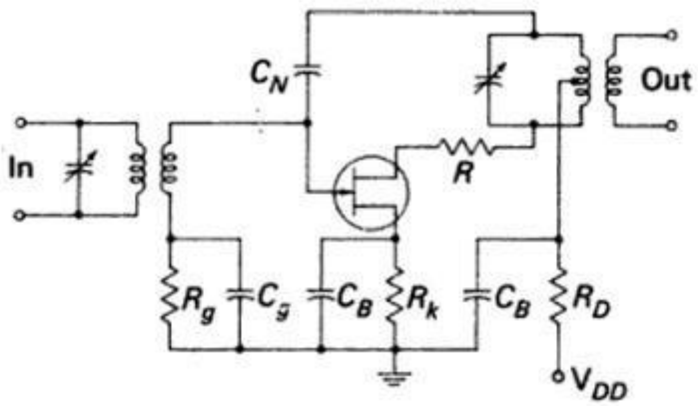


FIGURE **Amplitude limiter.**

Leak-type bias provides limiting, as shown in Figure . When input signal voltage rises, current flows in the $R_g - C_g$ bias circuit, and a negative voltage is developed across the capacitor. It is seen that the bias on the FET is increased in proportion to the size of the input voltage. As a result, the gain of the amplifier is lowered, and the output voltage tends to remain constant.

Although some limiting is achieved by this process, it is insufficient by itself, the action just described would occur only with rather large input voltages. To overcome this, early saturation of the output current is used, achieved by means of a low drain supply voltage. This is the reason for the drain dropping resistor of Figure . The supply voltage for a limiter is typically one-half of the normal dc drain voltage. The result of early saturation is to ensure limiting for conveniently low input voltages.

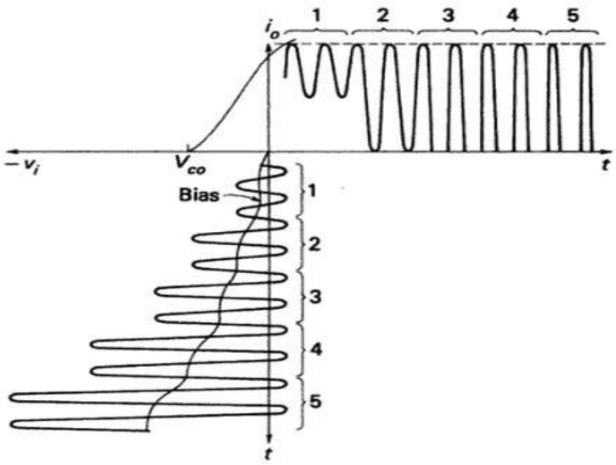


FIGURE Amplitude limiter transfer characteristic.

It is possible for the gate-drain section to become forward-biased under saturation conditions, causing a short circuit between input and output. To avert this, a resistance of a few hundred ohms is placed between the drain and its tank. This is R of Figure

MODULE-IV

Elements of Digital Communication Systems

Elements of Digital Communication Systems:

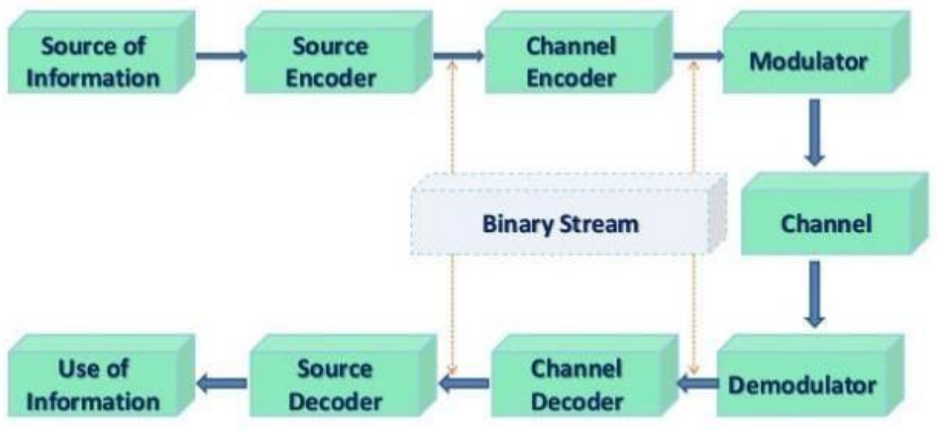


Fig. 1 Elements of Digital Communication Systems

1. **Information Source and Input Transducer:**

   The source of information can be analog or digital, e.g. analog: audio or video signal, digital: like teletype signal. In digital communication the signal produced by this source is converted into digital signal which consists of 1's and 0's. For this we need a source encoder.

2. **Source Encoder:**

   In digital communication we convert the signal from source into digital signal as mentioned above. The point to remember is we should like to use as few binary digits as possible to represent the signal. In such a way this efficient representation of the source output results in little or no redundancy. This sequence of binary digits is called ***information sequence.***

   Source Encoding or Data Compression: the process of efficiently converting the output of whether analog or digital source into a sequence of binary digits is known as source encoding.

3. **Channel Encoder:**

    The information sequence is passed through the channel encoder. The purpose of the channel encoder is to introduce, in controlled manner, some redundancy in the binary information sequence that can be used at the receiver to overcome the effects of noise and interference encountered in the transmission on the signal through the channel.

    For example take k bits of the information sequence and map that k bits to unique n bit sequence called code word. The amount of redundancy introduced is measured by the ratio *n/k* and the reciprocal of this ratio (k/n) is known as *rate of code or code rate.*

4. **Digital Modulator:**

    The binary sequence is passed to digital modulator which in turns convert the sequence into electric signals so that we can transmit them on channel (we will see channel later). The digital modulator maps the binary sequences into signal wave forms , for example if we represent 1 by sin x and 0 by cos x then we will transmit sin x for 1 and cos x for 0. (a case similar to BPSK)

5. **Channel:**

    The communication channel is the physical medium that is used for transmitting signals from transmitter to receiver. In wireless system, this channel consists of atmosphere , for traditional telephony, this channel is wired , there are optical channels, under water acoustic channels etc.We further discriminate this channels on the basis of their property and characteristics, like AWGN channel etc.

6. **Digital Demodulator:**

    The digital demodulator processes the channel corrupted transmitted waveform and reduces the waveform to the sequence of numbers that represents estimates of the transmitted data symbols.

7. **Channel Decoder:**

    This sequence of numbers then passed through the channel decoder which attempts to reconstruct the original information sequence from the knowledge of the code used by the channel encoder and the redundancy contained in the received data

Note: The average probability of a bit error at the output of the decoder is a measure of the performance of the demodulator – decoder combination.

8. **Source Decoder:**

    At the end, if an analog signal is desired then source decoder tries to decode the sequence from the knowledge of the encoding algorithm. And which results in the approximate replica of the input at the transmitter end

9. **Output Transducer:**

Finally we get the desired signal in desired format analog or digital.

Advantages of digital communication:

- Can withstand channel noise and distortion much better as long as the noise and the distortion are within limits.
- Regenerative repeaters prevent accumulation of noise along the path.
- Digital hardware implementation is flexible.
- Digital signals can be coded to yield extremely low error rates, high fidelity and well as privacy.
- Digital communication is inherently more efficient than analog in realizing the exchange of SNR for bandwidth.
- It is easier and more efficient to multiplex several digital signals.

- Digital signal storage is relatively easy and inexpensive.
- Reproduction with digital messages is extremely reliable without deterioration.
- The cost of digital hardware continues to halve every two or three years, while performance or capacity doubles over the same time period.

Disadvantages

- TDM digital transmission is not compatible with the FDM
- A Digital system requires large bandwidth.

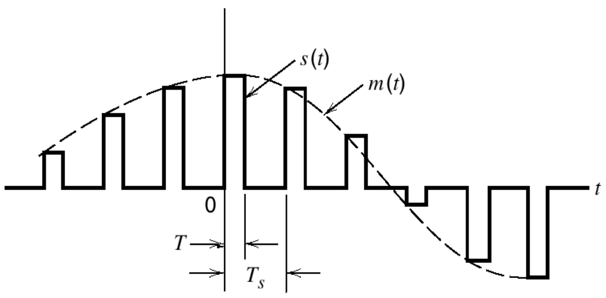
- Carrier is a train of pulses
- Example: Pulse Amplitude Modulation (PAM), Pulse width modulation (PWM) , Pulse Position Modulation (PPM)

Types of Pulse Modulation:

- The immediate result of sampling is a pulse-amplitude modulation (PAM) signal
- PAM is an analog scheme in which the amplitude of the pulse is proportional to the amplitude of the signal at the instant of sampling
- Another analog pulse-forming technique is known as **pulse-duration modulation (PDM).** This is also known as **pulse-width modulation (PWM)**
- **Pulse-position modulation** is closely related to PDM

Pulse Amplitude Modulation:

In PAM, amplitude of pulses is varied in accordance with instantaneous value of modulating signal.



PAM Generation:

The carrier is in the form of narrow pulses having frequency fc. The uniform sampling takes place in multiplier to generate PAM signal. Samples are placed Ts sec away from each other.

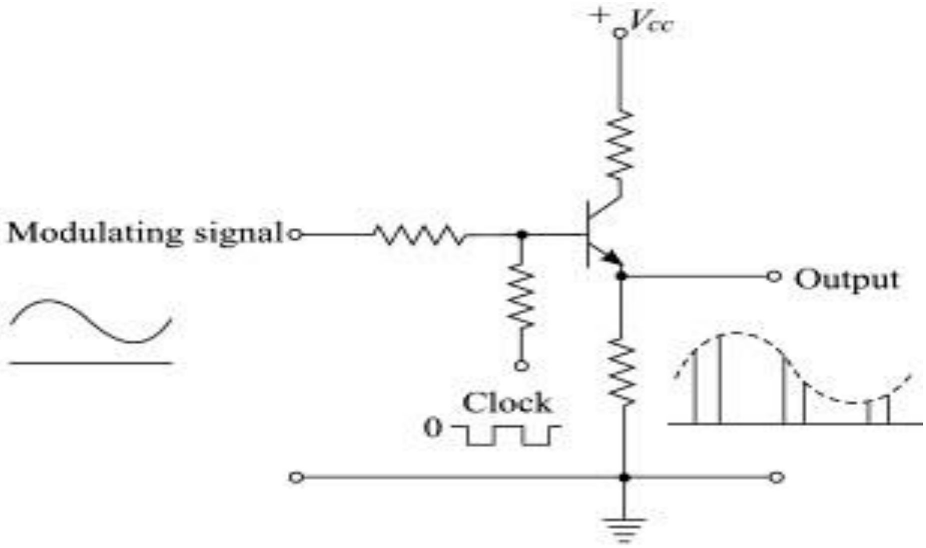


Fig.12. PAM Modulator

- The circuit is simple emitter follower.
- In the absence of the clock signal, the output follows input.
- The modulating signal is applied as the input signal.
- Another input to the base of the transistor is the clock signal.
- The frequency of the clock signal is made equal to the desired carrier pulse train frequency.
- The amplitude of the clock signal is chosen the high level is at ground level(0v) and low level at some negative voltage sufficient to bring the transistor in cutoff region.
- When clock is high, circuit operates as emitter follower and the output follows in the input modulating signal.
- When clock signal is low, transistor is cutoff and output is zero.
- Thus the output is the desired PAM signal.

PAM Demodulator:

- The PAM demodulator circuit which is just an envelope detector followed by a second order op-amp low pass filter (to have good filtering characteristics) is as shown below

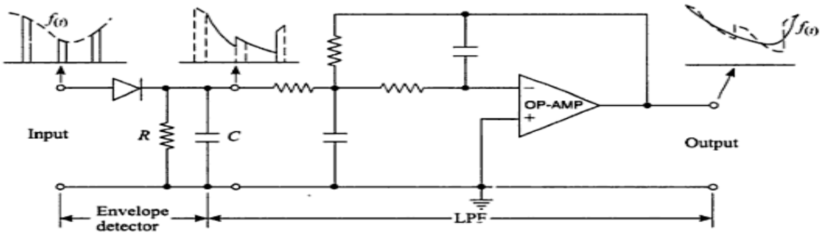
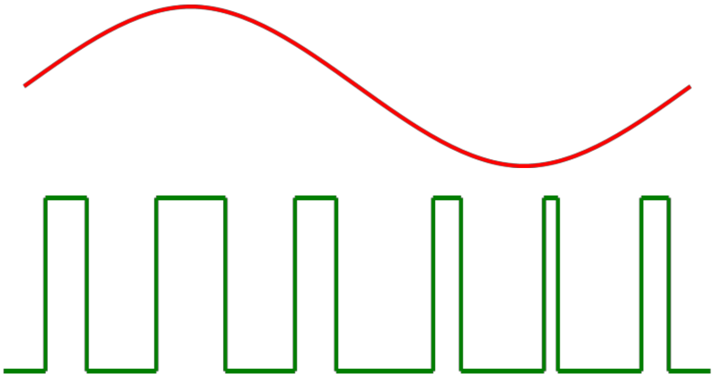


Fig.13. PAM Demodulator

Pulse Width Modulation:

- In this type, the amplitude is maintained constant but the width of each pulse is varied in accordance with instantaneous value of the analog signal.



- In PWM information is contained in width variation. This is similar to FM.
- In pulse width modulation (PWM), the width of each pulse is made directly proportional to the amplitude of the information signal.

Pulse Position Modulation:

- In this type, the sampled waveform has fixed amplitude and width whereas the position of each pulse is varied as per instantaneous value of the analog signal.
- PPM signal is further modification of a PWM signal.

PPM & PWM Modulator:

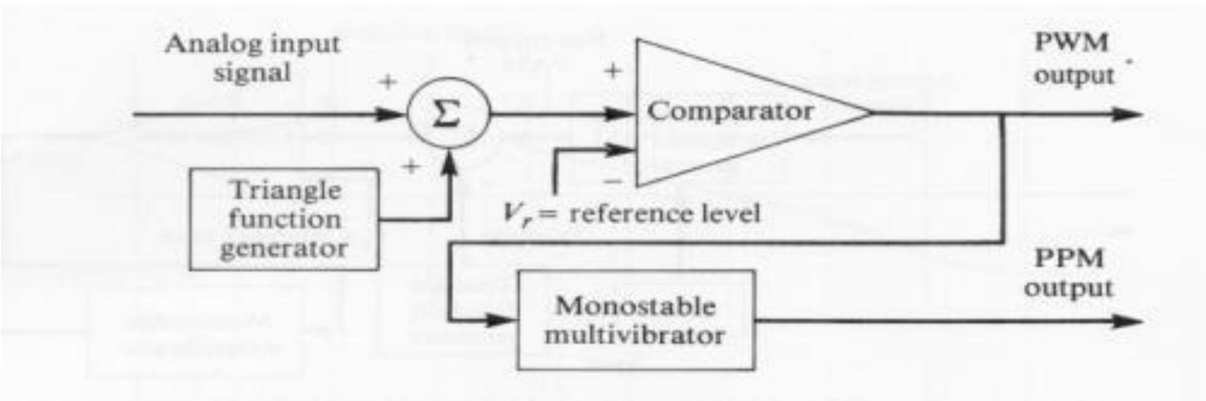


Fig.14. PWM & PPM Modulator

- The PPM signal can be generated from PWM signal.
- The PWM pulses obtained at the comparator output are applied to a mono stable multi vibrator which is negative edge triggered.

- Hence for each trailing edge of PWM signal, the monostable output goes high. It remains high for a fixed time decided by its RC components.
- Thus as the trailing edges of the PWM signal keeps shifting in proportion with the modulating signal, the PPM pulses also keep shifting.
- Therefore all the PPM pulses have the same amplitude and width. The information is conveyed via changing position of pulses.

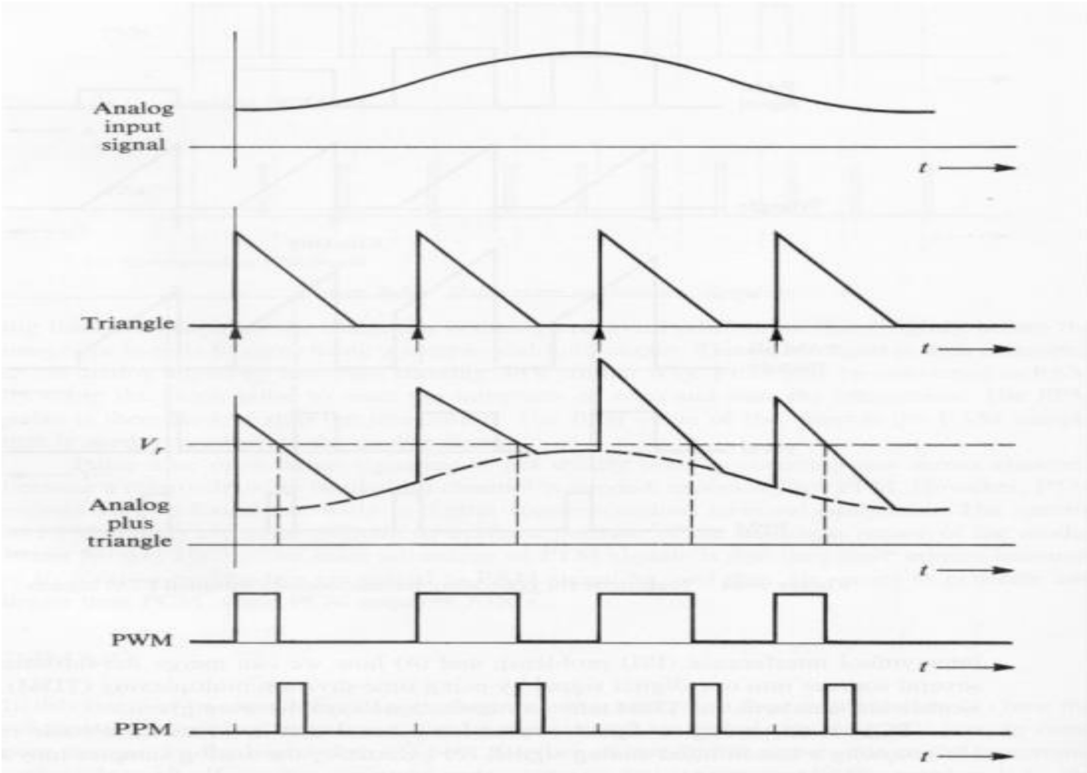


Fig.15. PWM & PPM Modulation waveforms

PWM Demodulator:

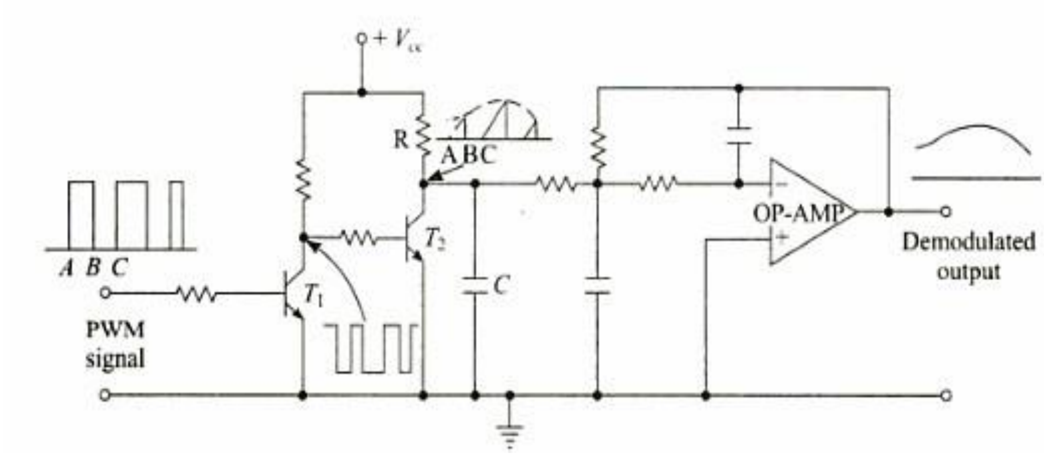


Fig.16. PWM Demodulator

- Transistor T1 works as an inverter.
- During time interval A-B when the PWM signal is high the input to transistor T2 is low.
- Therefore, during this time interval T2 is cut-off and capacitor C is charged through an R-C combination.
- During time interval B-C when PWM signal is low, the input to transistor T2 is high, and it gets saturated.
- The capacitor C discharges rapidly through T2.The collector voltage of T2 during B-C is low.
- Thus, the waveform at the collector of T2is similar to saw-tooth waveform whose envelope is the modulating signal.
- Passing it through 2nd order op-amp Low Pass Filter, gives demodulated signal.

PPM Demodulator:

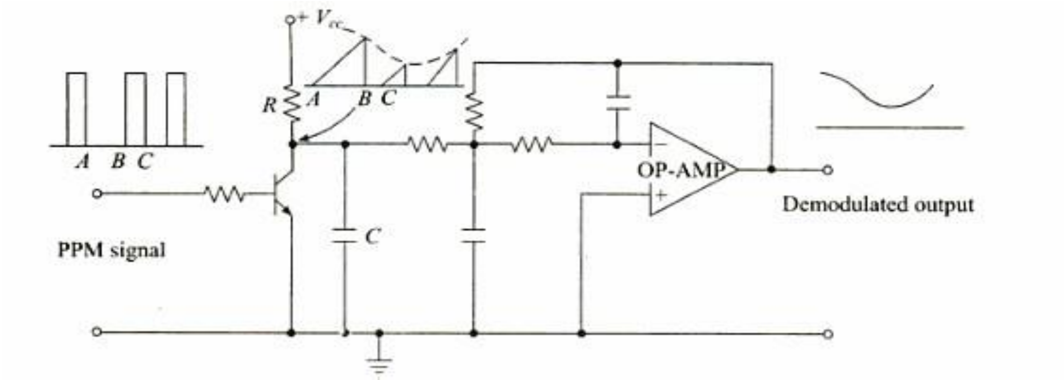


Fig.17. PPM Demodulator

- The gaps between the pulses of a PPM signal contain the information regarding the modulating signal.
- During gap A-B between the pulses the transistor is cut-off and the capacitor C gets charged through R-C combination.
- During the pulse duration B-C the capacitor discharges through transistor and the collector voltage becomes low.
- Thus, waveform across collector is saw-tooth waveform whose envelope is the modulating signal.
- Passing it through 2nd order op-amp Low Pass Filter, gives demodulated signal.

Fig. 3 Elements of PCM System

Sampling:

- Process of converting analog signal into discrete signal.
- Sampling is common in all pulse modulation techniques
- The signal is sampled at regular intervals such that each sample is proportional to amplitude of signal at that instant
- Analog signal is sampled every T_s $Secs$, called sampling interval. f_s=1/T_S is called sampling rate or sampling frequency.
- f_s=2f_m is Min. sampling rate called **Nyquist rate.** Sampled spectrum (ω) is repeating periodically without overlapping.
- Original spectrum is centered at ω=0 and having bandwidth of ω_m. Spectrum can be recovered by passing through low pass filter with cut-off ω_m.
- For f_s<2f_m sampled spectrum will overlap and cannot be recovered back. This is called **aliasing.**

Sampling methods:

- Ideal – An impulse at each sampling instant.
- Natural – A pulse of Short width with varying amplitude.
- Flat Top – Uses sample and hold, like natural but with single amplitude value.

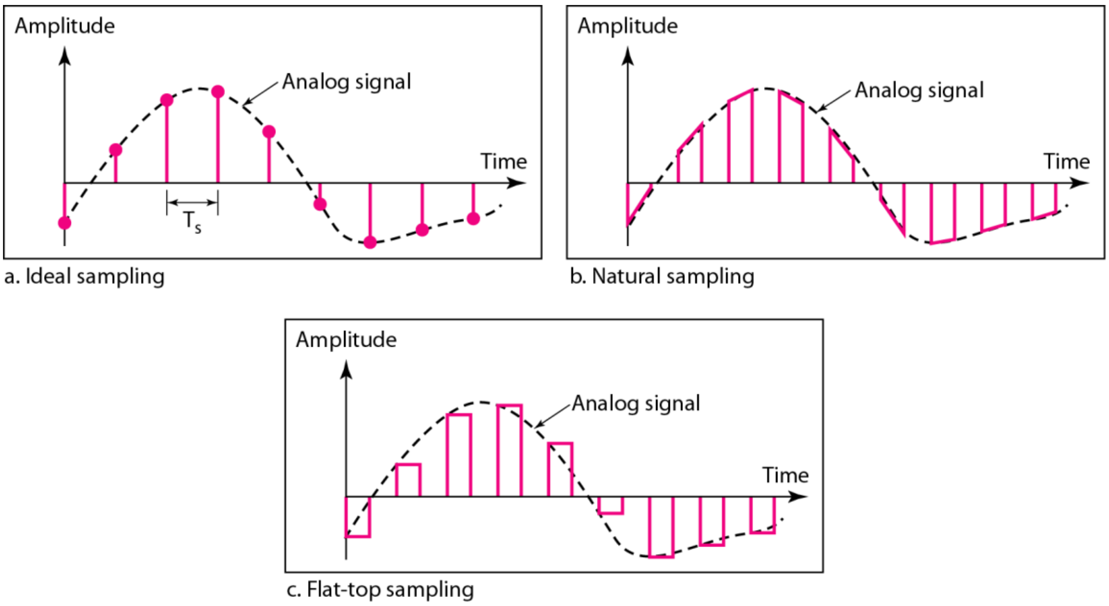


Fig. 4 Types of Sampling

PCM Generator:

The pulse code modulator technique samples the input signal $x(t)$ at frequency $f_s \geq 2W$. This sampled 'Variable amplitude' pulse is then digitized by the analog to digital converter. The parallel bits obtained are converted to a serial bit stream. Fig. 8 shows the PCM generator.

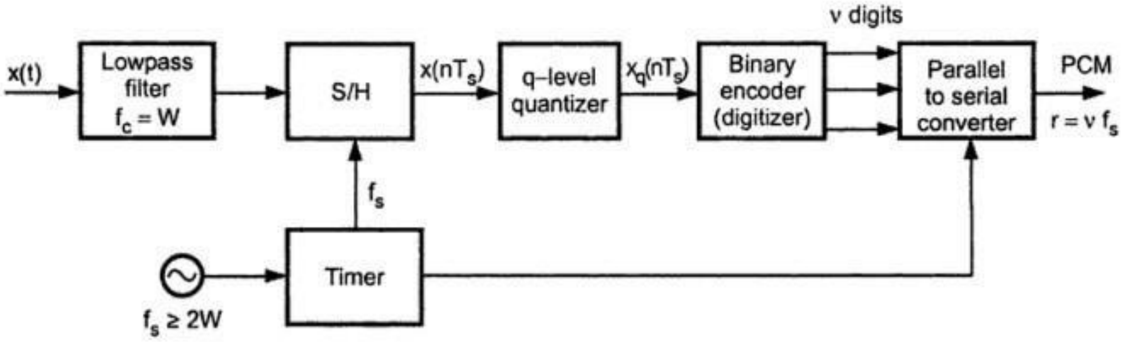


Fig. 8 PCM generator

In the PCM generator of above figure, the signal $x(t)$ is first passed through the lowpass filter of cutoff frequency 'W' Hz. This lowpass filter blocks all the frequency components above 'W' Hz. Thus $x(t)$ is bandlimited to 'W' Hz. The sample and hold circuit then samples this signal at the rate of f_s. Sampling frequency f_s is selected sufficiently above Nyquist rate to avoid aliasing i.e.,

$$f_s \geq 2W$$

In Fig. 8 output of sample and hold is called $x(nT_s)$. This $x(nT_s)$ is discrete in time and continuous in amplitude. A q-level quantizer compares input $x(nT_s)$ with its fixed digital levels. It assigns any one of the digital level to $x(nT_s)$ with its fixed digital levels. It then assigns any one of the digital level to $x(nT_s)$ which results in minimum distortion or error. This error is called *quantization error*. Thus output of quantizer is a digital level called $x_q(nT_s)$.

Now coming back to our discussion of PCM generation, the quantized signal level $x_q(nT_s)$ is given to binary encoder. This encoder converts input signal to 'v' digits binary word. Thus $x_q(nT_s)$ is converted to 'V' binary bits. The encoder is also called digitizer.

It is not possible to transmit each bit of the binary word separately on transmission line. Therefore '*v*' binary digits are converted to serial bit stream to generate single baseband signal. In a parallel to serial converter, normally a shift register does this job. The output of PCM generator is thus a single baseband signal of binary bits.

An oscillator generates the clocks for sample and hold an parallel to serial converter. In the pulse code modulation generator discussed above ; sample and hold, quantizer and encoder combinely form an analog to digital converter.

Transmission BW in PCM:

Let the quantizer use 'v' number of binary digits to represent each level. Then the number of levels that can be represented by 'v' digits will be,

$$q = 2^v \quad \ldots \quad 1$$

Here 'q' represents total number of digital levels of q-level quantizer.

For example if $v = 3$ bits, then total number of levels will be,

$$q = 2^3 = 8 \text{ levels}$$

Each sample is converted to 'v' binary bits. i.e. Number of bits per sample = v

We know that, Number of samples per second = f_s

∴ Number of bits per second is given by,

$$\begin{aligned}(\text{Number of bits per second}) &= (\text{Number of bits per samples}) \\ &\quad \times (\text{Number of samples per second}) \\ &= v \text{ bits per sample} \times f_s \text{ samples per second} \quad \ldots \quad 2\end{aligned}$$

The number of bits per second is also called signaling rate of PCM and is denoted by 'r' i.e.,

$$\boxed{\textbf{Signaling rate in PCM} : r = v f_s} \quad \ldots \quad 3$$

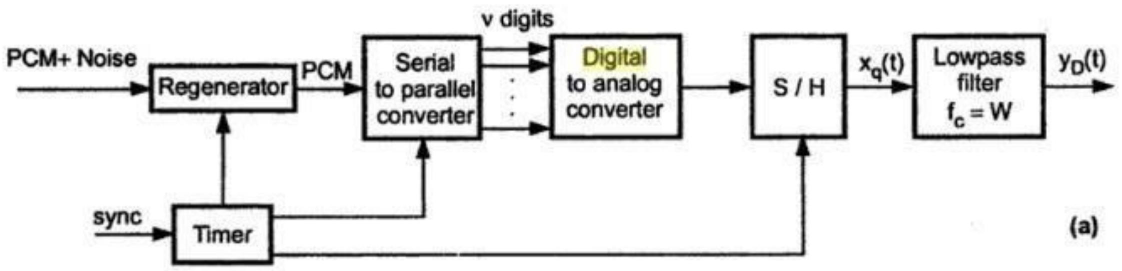
Here $f_s \geq 2W$.

Bandwidth needed for PCM transmission will be given by half of the signaling rate i.e.,

$$\textbf{Transmission Bandwidth of PCM} : \begin{cases} B_T \geq \frac{1}{2} r & \ldots \quad 4 \\ B_T \geq \frac{1}{2} v f_s \quad \text{Since } f_s \geq 2W & \ldots \quad 5 \\ B_T \geq v W & \ldots \quad 6 \end{cases}$$

PCM Receiver:

Fig. 9 (a) shows the block diagram of PCM receiver and Fig. 9 (b) shows the reconstructed signal. The regenerator at the start of PCM receiver reshapes the pulses and removes the noise. This signal is then converted to parallel digital words for each sample.



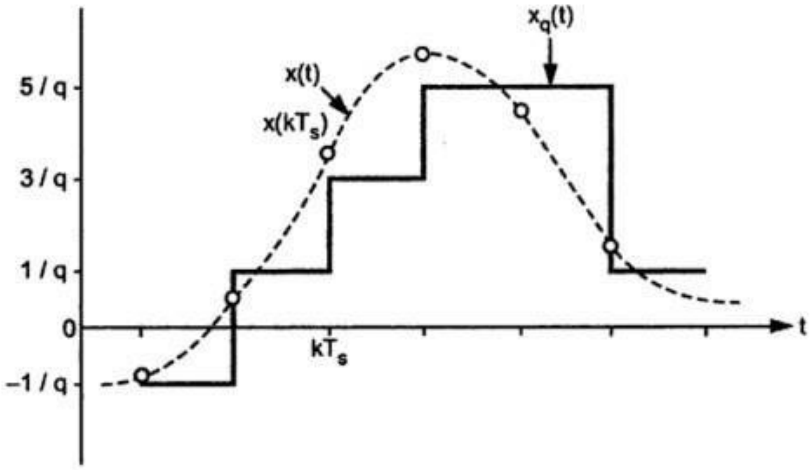


Fig. 9 (a) PCM receiver
(b) Reconstructed waveform

The digital word is converted to its analog value $x_q(t)$ along with sample and hold. This signal, at the output of S/H is passed through lowpass reconstruction filter to get $y_D(t)$. As shown in reconstructed signal of Fig. 9 (b), it is impossible to reconstruct exact original signal $x(t)$ because of permanent quantization error introduced during quantization at the transmitter. This quantization error can be reduced by increasing the binary levels. This is equivalent to increasing binary digits (bits) per sample. But increasing bits 'v' increases the signaling rate as well as transmission bandwidth as we have seen in equation 3 and equation 6. Therefore the choice of these parameters is made, such that noise due to quantization error (called as quantization noise) is in tolerable limits.

Quantization

- The quantizing of an analog signal is done by discretizing the signal with a number of quantization levels.

- **Quantization** is representing the sampled values of the amplitude by a finite set of levels, which means converting a continuous-amplitude sample into a discrete-time signal
- Both sampling and quantization result in the loss of information.
- The quality of a Quantizer output depends upon the number of quantization levels used.
- The discrete amplitudes of the quantized output are called as **representation levels** or **reconstruction levels**.
- The spacing between the two adjacent representation levels is called a **quantum** or **step-size**.
- There are two types of Quantization
    - Uniform Quantization
    - Non-uniform Quantization.
- The type of quantization in which the quantization levels are uniformly spaced is termed as a **Uniform Quantization**.
- The type of quantization in which the quantization levels are unequal and mostly the relation between them is logarithmic, is termed as a **Non-uniform Quantization.**

Uniform Quantization:

- There are two types of uniform quantization.
    - Mid-Rise type
    - Mid-Tread type.
- The following figures represent the two types of uniform quantization.

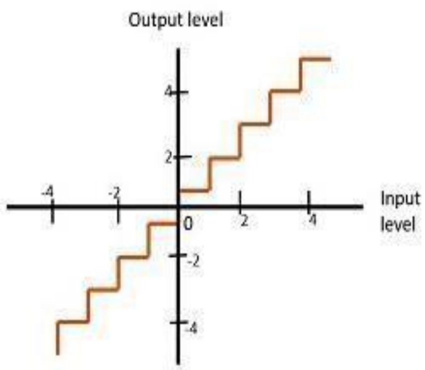


Fig 1 : Mid-Rise type Uniform Quantization

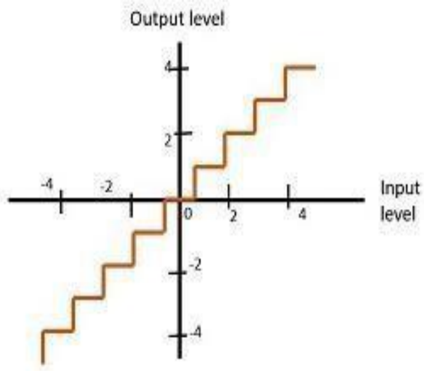


Fig 2 : Mid-Tread type Uniform Quantization

- The **Mid-Rise** type is so called because the origin lies in the middle of a raising part of the stair-case like graph. The quantization levels in this type are even in number.
- The **Mid-tread** type is so called because the origin lies in the middle of a tread of the stair-case like graph. The quantization levels in this type are odd in number.
- Both the mid-rise and mid-tread type of uniform quantizer is symmetric about the origin.

Quantization Noise and Signal to Noise ratio in PCM System:

Derivation of Quantization Error/Noise or Noise Power for Uniform (Linear) Quantization

Step 1 : Quantization Error

Because of quantization, inherent errors are introduced in the signal. This error is called *quantization error*. We have defined quantization error as,

$$\varepsilon = x_q(nT_s) - x(nT_s) \qquad \text{..........} (1)$$

Step 2 : Step size

Let an input $x(nT_s)$ be of continuous amplitude in the range $-x_{max}$ to $+x_{max}$.

Therefore the total amplitude range becomes,

$$\text{Total amplitude range} = x_{max} - (-x_{max}) = 2x_{max} \qquad \text{..........} (2)$$

If this amplitude range is divided into 'q' levels of quantizer, then the step size 'δ' is given as,

$$\delta = \frac{x_{max} - (-x_{max})}{q}$$

$$= \frac{2x_{max}}{q} \qquad \text{..........} (3)$$

If signal $x(t)$ is normalized to minimum and maximum values equal to 1, then

$$x_{max} = 1$$
$$-x_{max} = -1 \qquad \text{..........} (4)$$

Therefore step size will be,

$$\delta = \frac{2}{q} \quad \text{(for normalized signal)} \qquad \text{..........} (5)$$

Step 3 : Pdf of Quantization error

If step size 'δ' is sufficiently small, then it is reasonable to assume that the quantization error 'ε' will be uniformly distributed random variable. The maximum quantization error is given by

$$\varepsilon_{max} = \left|\frac{\delta}{2}\right| \qquad \text{..........} (6)$$

i.e.

$$-\frac{\delta}{2} \geq \varepsilon_{max} \geq \frac{\delta}{2} \qquad \text{..........} (7)$$

Thus over the interval $\left(-\frac{\delta}{2}, \frac{\delta}{2}\right)$ quantization error is uniformly distributed random variable.

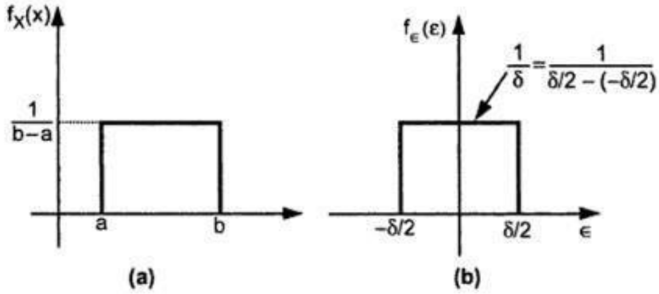


Fig. 10 (a) Uniform distribution
(b) Uniform distribution for quantization error

In above figure, a random variable is said to be uniformly distributed over an interval (a, b). Then PDF of 'X' is given by, (from equation of Uniform PDF).

$$f_X(x) = \begin{cases} 0 & \text{for } x \le a \\ \dfrac{1}{b-a} & \text{for } a < x \le b \\ 0 & \text{for } x > b \end{cases} \quad \text{.......................(8)}$$

Thus with the help of above equation we can define the probability density function for quantization error 'ε' as,

$$f_\varepsilon(\varepsilon) = \begin{cases} 0 & \text{for } \varepsilon \le \dfrac{\delta}{2} \\ \dfrac{1}{\delta} & \text{for } -\dfrac{\delta}{2} < \varepsilon \le \dfrac{\delta}{2} \\ 0 & \text{for } \varepsilon > \dfrac{\delta}{2} \end{cases} \quad \text{..................................(9)}$$

Step 4 : Noise Power

quantization error 'ε' has zero average value. That is mean 'm_ε' of the quantization error is zero.

The signal to quantization noise ratio of the quantizer is defined as,

$$\frac{S}{N} = \frac{\text{Signal power (normalized)}}{\text{Noise power (normalized)}} \quad \cdots 10$$

If type of signal at input i.e., $x(t)$ is known, then it is possible to calculate signal power.

The noise power is given as,

$$\text{Noise power} = \frac{V_{noise}^2}{R} \quad \cdots (11)$$

Here V_{noise}^2 is the mean square value of noise voltage. Since noise is defined by random variable 'ε' and PDF $f_\varepsilon(\varepsilon)$, its mean square value is given as,

$$\text{mean square value} = E[\varepsilon^2] = \bar{\varepsilon}^2 \quad \cdots (12)$$

The mean square value of a random variable 'X' is given as,

$$\bar{X}^2 = E[X^2] = \int_{-\infty}^{\infty} x^2 f_X(x)\,dx \quad \text{By definition} \quad \cdots (13)$$

Here $$E[\varepsilon^2] = \int_{-\infty}^{\infty} \varepsilon^2 f_\varepsilon(\varepsilon)\,d\varepsilon \quad \cdots (14)$$

From equation 9 we can write above equation as,

$$E[\varepsilon^2] = \int_{-\delta/2}^{\delta/2} \varepsilon^2 \times \frac{1}{\delta}\,d\varepsilon$$

$$= \frac{1}{\delta}\left[\frac{\varepsilon^3}{3}\right]_{-\delta/2}^{\delta/2} = \frac{1}{\delta}\left[\frac{(\delta/2)^3}{3} + \frac{(\delta/2)^3}{3}\right]$$

$$= \frac{1}{3\delta}\left[\frac{\delta^3}{8} + \frac{\delta^3}{8}\right] = \frac{\delta^2}{12} \quad \cdots (15)$$

∴ From equation 1.8.25, the mean square value of noise voltage is,

$$V_{noise}^2 = \text{mean square value} = \frac{\delta^2}{12}$$

When load resistance, $R = 1$ ohm, then the noise power is normalized i.e.,

$$\text{Noise power (normalized)} = \frac{V_{noise}^2}{1} \qquad [\text{with } R = 1 \text{ in equation 11 }]$$

$$= \frac{\delta^2 / 12}{1} = \frac{\delta^2}{12}$$

Thus we have,

Normalized noise power

or Quantization noise power $= \dfrac{\delta^2}{12}$ **; For linear quantization.**

or Quantization error (in terms of power) ... (16)

Derivation of Maximum Signal to Quantization Noise Ratio for Linear Quantization:

signal to quantization noise ratio is given as,

$$\frac{S}{N} = \frac{\text{Normalized signal power}}{\text{Normalized noise power}}$$

$$= \frac{\text{Normalized signal power}}{(\delta^2 / 12)} \qquad \text{... (17)}$$

The number of bits 'v' and quantization levels 'q' are related as,

$$q = 2^v \qquad \text{... (18)}$$

Putting this value in equation (3) we have,

$$\delta = \frac{2\,x_{\max}}{2^v} \qquad \text{... (19)}$$

Putting this value in equation 1.8.30 we get,

$$\frac{S}{N} = \frac{\text{Normalized signal power}}{\left(\frac{2\,x_{\max}}{2^v}\right)^2 + 12}$$

Let normalized signal power be denoted as 'P'.

$$\frac{S}{N} = \frac{P}{\frac{4\,x_{\max}^2}{2^{2v}} \times \frac{1}{12}} = \frac{3P}{x_{\max}^2} \cdot 2^{2v}$$

This is the required relation for maximum signal to quantization noise ratio. Thus,

$$\textbf{Maximum signal to quantization noise ratio : } \frac{S}{N} = \frac{3P}{x_{max}^2} \cdot 2^{2v} \qquad \cdots (20)$$

This equation shows that signal to noise power ratio of quantizer increases exponentially with increasing bits per sample.

If we assume that input $x(t)$ is normalized, i.e.,

$$x_{max} = 1 \qquad \cdots (21)$$

Then signal to quantization noise ratio will be,

$$\frac{S}{N} = 3 \times 2^{2v} \times P \qquad \cdots (22)$$

If the destination signal power 'P' is normalized, i.e.,

$$P \leq 1 \qquad \cdots (23)$$

Then the signal to noise ratio is given as,

$$\frac{S}{N} \leq 3 \times 2^{2v} \qquad \cdots (24)$$

Since $x_{max} = 1$ and $P \leq 1$, the signal to noise ratio given by above equation is normalized.

Expressing the signal to noise ratio in decibels,

$$\left(\frac{S}{N}\right) dB = 10 \log_{10} \left(\frac{S}{N}\right) dB \qquad \text{since power ratio.}$$

$$\leq 10 \log_{10} [3 \times 2^{2v}]$$

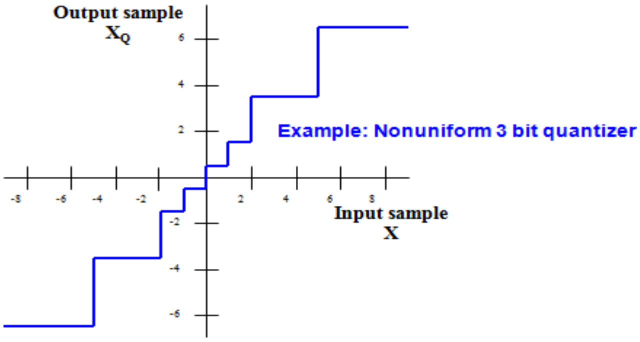
$$\leq (4.8 + 6v)\, dB$$

Thus,

Signal to Quantization noise ratio for normalized values of power 'P' and amplitude of input $x(t)$: $\left(\frac{S}{N}\right) dB \leq (4.8 + 6v)\, dB$ $\cdots (25)$

Non-Uniform Quantization:

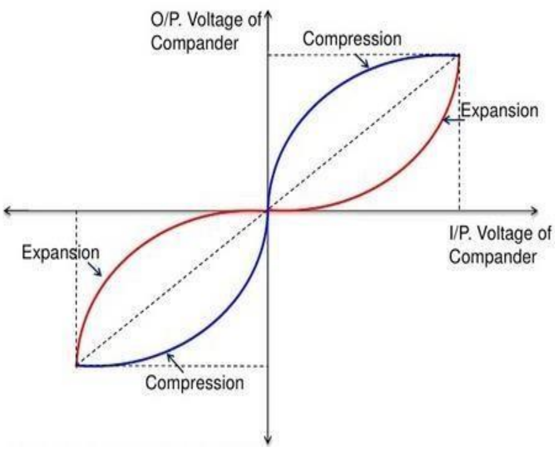
In non-uniform quantization, the step size is not fixed. It varies according to certain law or as per input signal amplitude. The following fig shows the characteristics of Non uniform quantizer.

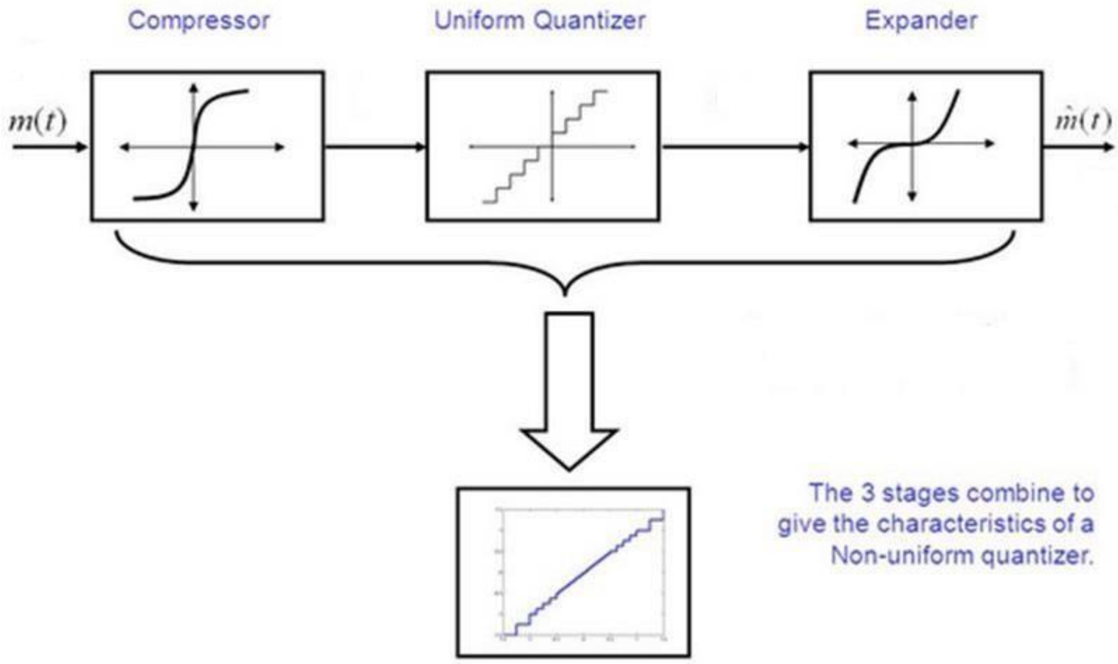


In this figure observe that step size is small at low input signal levels. Hence quantization error is also small at these inputs. Therefore signal to quantization noise power ratio is improved at low signal levels. Stepsize is higher at high input levels. Hence signal to noise power ratio remains almost same throughout the dynamic range of quantizer.

Companding PCM System:

- Non-uniform quantizers are difficult to make and expensive.
- An alternative is to first pass the speech signal through nonlinearity before quantizing with a uniform quantizer.
- The nonlinearity causes the signal amplitude to be ***compressed***.
  - The input to the quantizer will have a more uniform distribution.
- At the receiver, the signal is ***expanded*** by an inverse to the nonlinearity.
- The process of compressing and expanding is called ***Companding***.





μ - Law Companding for Speech Signals

Normally for speech and music signals a μ - law compression is used. This compression is defined by the following equation,

$$Z(x) = (\text{Sgn } x) \frac{\ln(1+\mu|x|)}{\ln(1+\mu)} \quad |x| \leq 1 \qquad \cdots (1)$$

Below Fig shows the variation of signal to noise ratio with respect to signal level without companding and with companding.

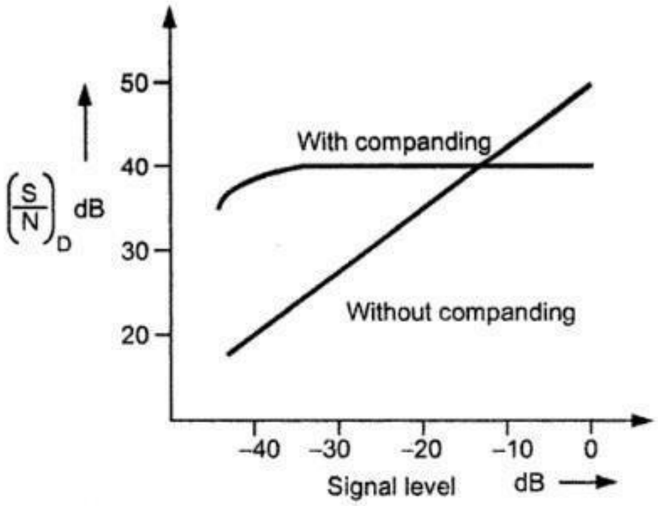


Fig. 11 PCM performance with μ - law companding

It can be observed from above figure that signal to noise ratio of PCM remains almost constant with companding.

A-Law for Companding

The A law provides piecewise compressor characteristic. It has linear segment for low level inputs and logarithmic segment for high level inputs. It is defined as,

$$Z(x) = \begin{cases} \dfrac{A|x|}{1+\ln A} & \text{for } 0 \le |x| \le \dfrac{1}{A} \\ \dfrac{1+\ln(A|x|)}{1+\ln A} & \text{for } \dfrac{1}{A} \le |x| \le 1 \end{cases} \quad \cdots (2)$$

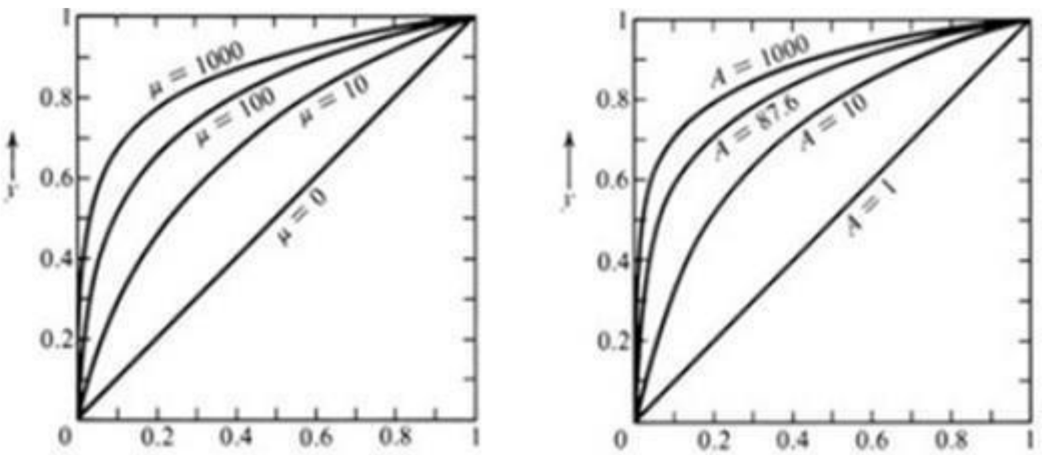
When A = 1, we get uniform quantization. The practical value for A is 87.56. Both A-law and μ-law companding is used for PCM telephone systems.

Signal to Noise Ratio of Companded PCM

The signal to noise ratio of companded PCM is given as,

$$\frac{S}{N} = \frac{3q^2}{[\ln(1+\mu)]^2} \quad \cdots (3)$$

Here $q = 2^v$ is number of quantization levels.



Differential Pulse Code Modulation (DPCM):

Redundant Information in PCM:

The samples of a signal are highly corrected with each other. This is because any signal does not change fast. That is its value from present sample to next sample does not differ by large amount. The adjacent samples of the signal carry the same information with little difference. When these samples are encoded by standard PCM system, the resulting encoded signal contains redundant information.

Fig. shows a continuous time signal $x(t)$ by dotted line. This signal is sampled by flat top sampling at intervals $T_s, 2T_s, 3T_s \dots nT_s$. The sampling frequency is selected to be higher than nyquist rate. The samples are encoded by using 3 bit (7 levels) PCM. The sample is quantized to the nearest digital level as shown by small

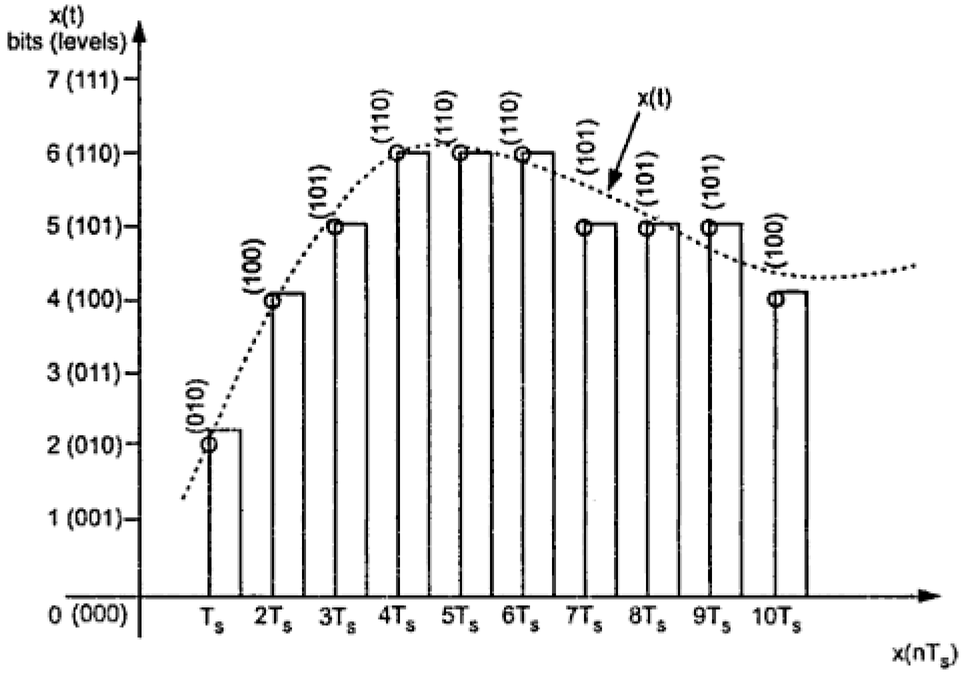


Fig. Redundant information in PCM

circles in the diagram. The encoded binary value of each sample is written on the top of the samples. We can see from Fig. that the samples taken at $4T_s, 5T_s$ and $6T_s$ are encoded to same value of (110). This information can be carried only by one sample. But three samples are carrying the same information means it is redundant. Consider another example of samples taken at $9T_s$ and $10T_s$. The difference between these samples is only due to last bit and first two bits are redundant, since they do not change.

Principle of DPCM

If this redundancy is reduced, then overall bit rate will decrease and number of bits required to transmit one sample will also be reduced. This type of digital pulse modulation scheme is called Differential Pulse Code Modulation.

DPCM Transmitter

The differential pulse code modulation works on the principle of prediction. The value of the present sample is predicted from the past samples. The prediction may not be exact but it is very close to the actual sample value. Fig. shows the transmitter of Differential Pulse Code Modulation (DPCM) system. The sampled signal is denoted by $x(nT_s)$ and the predicted signal is denoted by $\hat{x}(nT_s)$. The comparator finds out the difference between the actual sample value $x(nT_s)$ and predicted sample value $\hat{x}(nT_s)$. This is called error and it is denoted by $e(nT_s)$. It can be defined as,

$$e(nT_s) = x(nT_s) - \hat{x}(nT_s) \qquad \text{..................(1)}$$

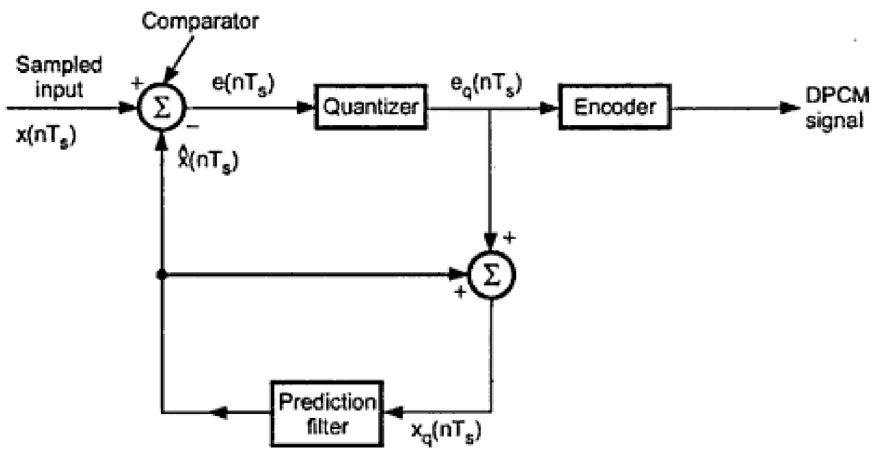


Fig. Differential pulse code modulation transmitter

Thus error is the difference between unquantized input sample $x(nT_s)$ and prediction of it $\hat{x}(nT_s)$. The predicted value is produced by using a prediction filter. The quantizer output signal $e_q(nT_s)$ and previous prediction is added and given as input to the prediction filter. This signal is called $x_q(nT_s)$. This makes the prediction more and more close to the actual sampled signal. We can see that the quantized error signal $e_q(nT_s)$ is very small and can be encoded by using small number of bits. Thus number of bits per sample are reduced in DPCM.

The quantizer output can be written as,

$$e_q(nT_s) = e(nT_s) + q(nT_s) \quad \text{.................(2)}$$

Here $q(nT_s)$ is the quantization error. As shown in Fig. the prediction filter input $x_q(nT_s)$ is obtained by sum $\hat{x}(nT_s)$ and quantizer output i.e.,

$$x_q(nT_s) = \hat{x}(nT_s) + e_q(nT_s) \quad \text{...................(3)}$$

Putting the value of $e_q(nT_s)$ from equation 2 in the above equation we get,

$$x_q(nT_s) = \hat{x}(nT_s) + e(nT_s) + q(nT_s) \quad \text{................(4)}$$

Equation 1 is written as,

$$e(nT_s) = x(nT_s) - \hat{x}(nT_s)$$

$$\therefore \quad e(nT_s) + \hat{x}(nT_s) = x(nT_s) \quad \text{...............(5)}$$

$\therefore$ Putting the value of $e(nT_s) + \hat{x}(nT_s)$ from above equation into equation 4 we get,

$$x_q(nT_s) = x(nT_s) + q(nT_s) \quad \text{..................(6)}$$

Thus the quantized version of the signal $x_q(nT_s)$ is the sum of original sample value and quantization error $q(nT_s)$. The quantization error can be positive or negative. Thus equation 6 does not depend on the prediction filter characteristics.

Reconstruction of DPCM Signal

Fig. shows the block diagram of DPCM receiver.

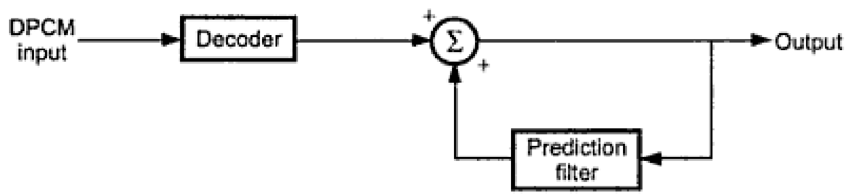


Fig. DPCM receiver

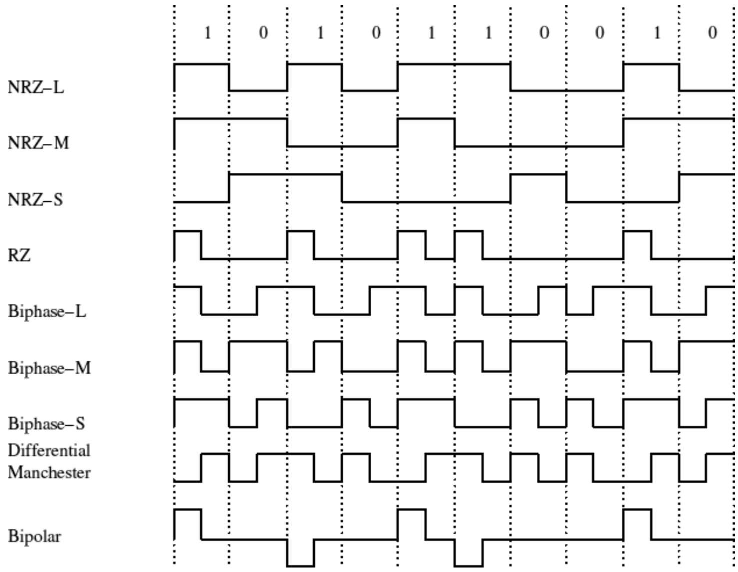
The decoder first reconstructs the quantized error signal from incoming binary signal. The prediction filter output and quantized error signals are summed up to give the quantized version of the original signal. Thus the signal at the receiver differs from actual signal by quantization error $q(nT_s)$, which is introduced permanently in the reconstructed signal.

Line Coding:

In telecommunication, a line code is a code chosen for use within a communications system for transmitting a digital signal down a transmission line. Line coding is often used for digital data transport.

The waveform pattern of voltage or current used to represent the 1s and 0s of a digital signal on a transmission link is called **line** encoding. The common types of

line encoding are unipolar, polar, bipolar and Manchester encoding. **Line codes** are used commonly in computer communication networks over short distances.



Signal	Comments
NRZ–L	Non-return to zero level. This is the standard positive logic signal format used in digital circuits. 1 forces a high level 0 forces a low level
NRZ–M	Non return to zero mark 1 forces a transition 0 does nothing
NRZ–S	Non return to zero space 1 does nothing 0 forces a transition
RZ	Return to zero 1 goes high for half the bit period 0 does nothing
Biphase–L	Manchester. Two consecutive bits of the same type force a transition at the beginning of a bit period. 1 forces a negative transition in the middle of the bit 0 forces a positive transition in the middle of the bit
Biphase–M	There is always a transition at the beginning of a bit period. 1 forces a transition in the middle of the bit 0 does nothing
Biphase–S	There is always a transition at the beginning of a bit period. 1 does nothing 0 forces a transition in the middle of the bit
Differential Manchester	There is always a transition in the middle of a bit period. 1 does nothing 0 forces a transition at the beginning of the bit
Bipolar	The positive and negative pulses alternate. 1 forces a positive or negative pulse for half the bit period 0 does nothing

Introduction to Delta Modulation

PCM transmits all the bits which are used to code the sample. Hence signaling rate and transmission channel bandwidth are large in PCM. To overcome this problem Delta Modulation is used.

Delta Modulation

Operating Principle of DM

Delta modulation transmits only one bit per sample. That is the present sample value is compared with the previous sample value and the indication,whether the amplitude is increased or decreased is sent. Input signal $x(t)$ is approximated to step signal by the delta modulator. This step size is fixed. The difference between the input signal $x(t)$ and staircase approximated signal confined to two levels, i.e. $+\delta$ and $-\delta$. If the difference is positive, then approximated signal is increased by one step i.e. 'δ'. If the difference is negative, then approximated signal is reduced by 'δ'. When the step is reduced, '0' is transmitted and if the step is increased, '1' is transmitted. Thus for each sample, only one binary bit is transmitted. Fig. shows the analog signal $x(t)$ and its staircase approximated signal by the delta modulator.

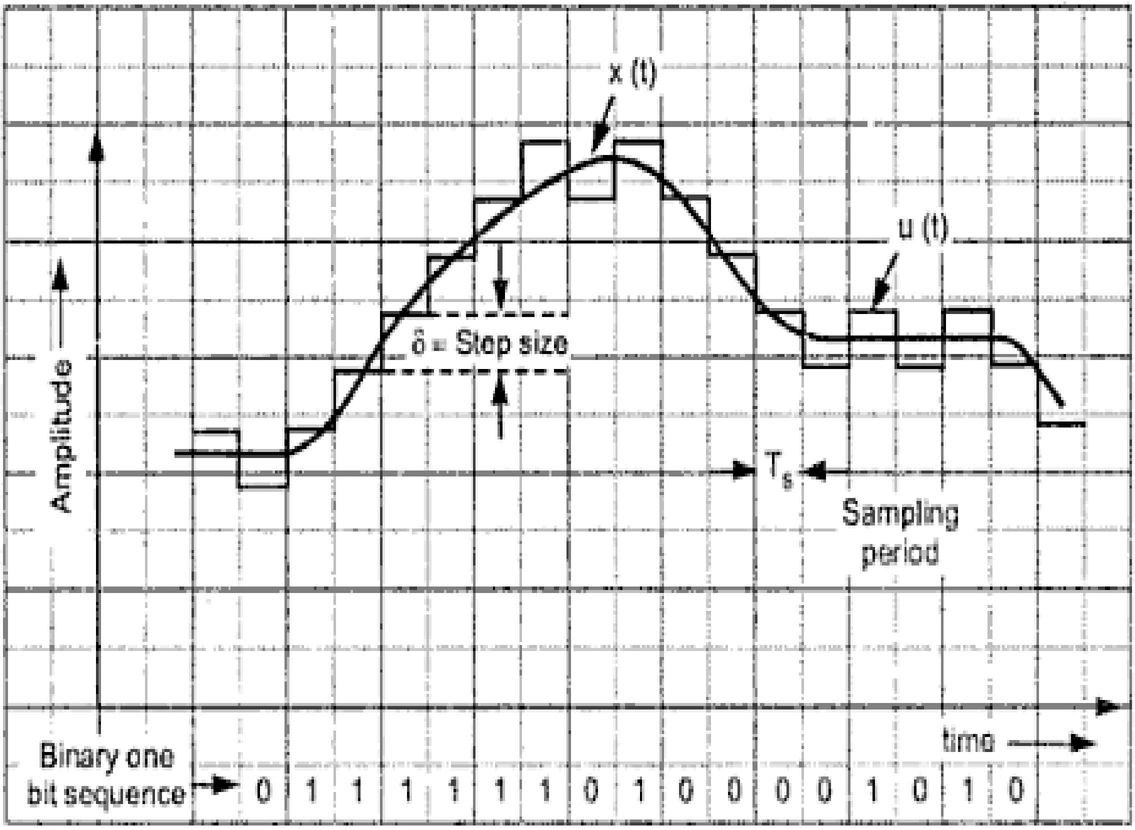


Fig. Delta modulation waveform

The principle of delta modulation can be explained by the following set of equations. The error between the sampled value of $x(t)$ and last approximated sample is given as,

$$e(nT_s) = x(nT_s) - \hat{x}(nT_s) \qquad \ldots (1)$$

Here, $e(nT_s)$ = Error at present sample

$x(nT_s)$ = Sampled signal of $x(t)$

$\hat{x}(nT_s)$ = Last sample approximation of the staircase waveform.

We can call $u(nT_s)$ as the present sample approximation of staircase output.

Then, $u[(n-1)T_s] = \hat{x}(nT_s) \qquad \ldots (2)$

= Last sample approximation of staircase waveform.

Let the quantity $b(nT_s)$ be defined as,

$$b(nT_s) = \delta\, sgn[e(nT_s)] \qquad \ldots (3)$$

That is depending on the sign of error $e(nT_s)$ the sign of step size δ will be decided. In other words,

$$b(nT_s) = +\delta \quad \text{if} \quad x(nT_s) \geq \hat{x}(nT_s)$$

$$= -\delta \quad \text{if} \quad x(nT_s) < \hat{x}(nT_s) \qquad \ldots (4)$$

If $b(nT_s) = +\delta$; binary '1' is transmitted

and if $b(nT_s) = -\delta$; binary '0' is transmitted.

T_s = Sampling interval.

DM Transmitter

Fig. (a) shows the transmitter based on equations 3 to 5.

The summer in the accumulator adds quantizer output ($\pm\delta$) with the previous sample approximation. This gives present sample approximation. i.e.,

$$u(nT_s) = u(nT_s - T_s) + [\pm\delta] \quad \text{or}$$

$$= u[(n-1)T_s] + b(nT_s) \qquad \ldots (5)$$

The previous sample approximation $u[(n-1)T_s]$ is restored by delaying one sample period T_s. The sampled input signal $x(nT_s)$ and staircase approximated signal $\hat{x}(nT_s)$ are subtracted to get error signal $e(nT_s)$.

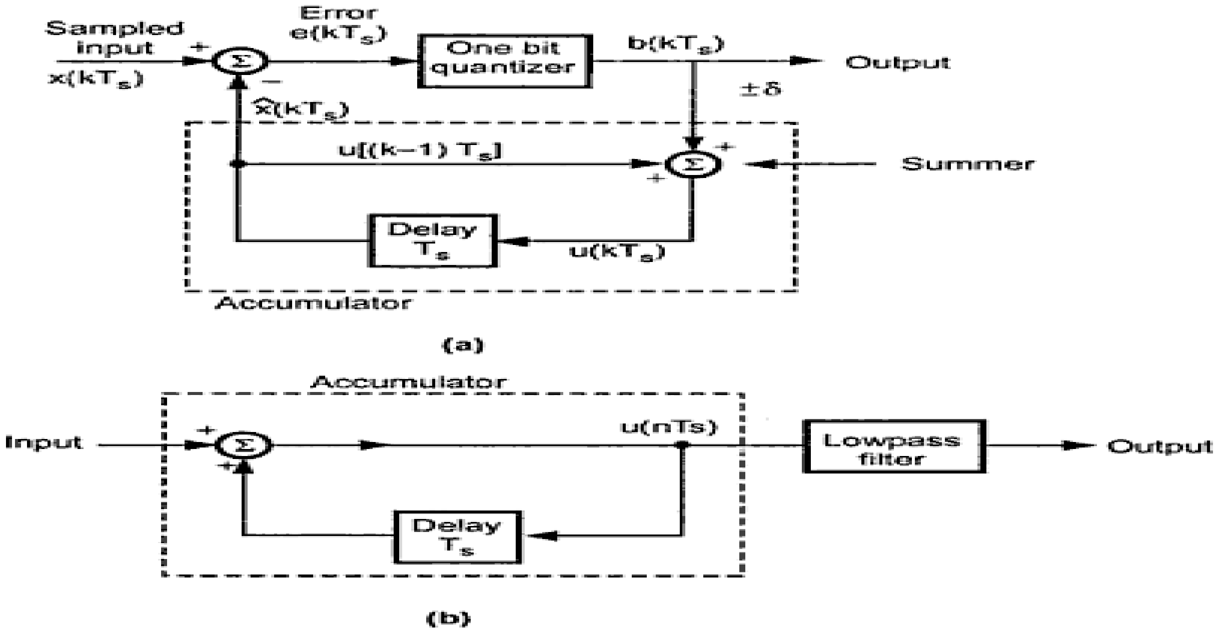


Fig. (a) Delta modulation transmitter and
(b) Delta modulation receiver

Depending on the sign of $e(nT_s)$ one bit quantizer produces an output step of $+\delta$ or $-\delta$. If the step size is $+\delta$, then binary '1' is transmitted and if it is $-\delta$, then binary '0' is transmitted.

DM Receiver

At the receiver shown in Fig. (b), the accumulator and low-pass filter are used. The accumulator generates the staircase approximated signal output and is delayed by one sampling period T_s. It is then added to the input signal. If input is binary '1' then it adds $+\delta$ step to the previous output (which is delayed). If input is binary '0' then one step 'δ' is subtracted from the delayed signal. The low-pass filter has the cutoff frequency equal to highest frequency in $x(t)$. This filter smoothen the staircase signal to reconstruct $x(t)$.

Advantages and Disadvantages of Delta Modulation

Advantages of Delta Modulation

The delta modulation has following advantages over PCM,

1. Delta modulation transmits only one bit for one sample. Thus the signaling rate and transmission channel bandwidth is quite small for delta modulation.
2. The transmitter and receiver implementation is very much simple for delta modulation. There is no analog to digital converter involved in delta modulation.

Disadvantages of Delta Modulation

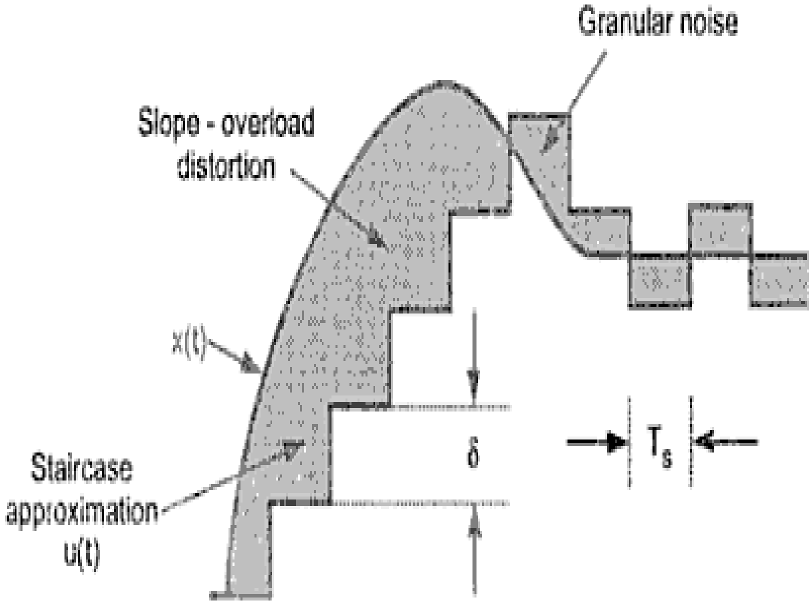


Fig. Quantization errors in delta modulation

The delta modulation has two drawbacks -

Slope Overload Distortion (Startup Error)

This distortion arises because of the large dynamic range of the input signal.

As can be seen from Fig. the rate of rise of input signal $x(t)$ is so high that the staircase signal cannot approximate it, the step size 'δ' becomes too small for staircase signal $u(t)$ to follow the steep segment of $x(t)$. Thus there is a large error between the staircase approximated signal and the original input signal $x(t)$. This error is called *slope overload distortion*. To reduce this error, the step size should be increased when slope of signal of $x(t)$ is high.

Since the step size of delta modulator remains fixed, its maximum or minimum slopes occur along straight lines. Therefore this modulator is also called Linear Delta Modulator (LDM).

Granular Noise (Hunting)

Granular noise occurs when the step size is too large compared to small variations in the input signal. That is for very small variations in the input signal, the staircase signal is changed by large amount (δ) because of large step size. Fig shows that when the input signal is almost flat, the staircase signal $u(t)$ keeps on oscillating by $\pm\delta$ around the signal. The error between the input and approximated signal is called *granular noise*. The solution to this problem is to make step size small.

Thus large step size is required to accommodate wide dynamic range of the input signal (to reduce slope overload distortion) and small steps are required to reduce granular noise. Adaptive delta modulation is the modification to overcome these errors.

Adaptive Delta Modulation

Operating Principle

To overcome the quantization errors due to slope overload and granular noise, the step size (δ) is made adaptive to variations in the input signal $x(t)$. Particularly in the steep segment of the signal $x(t)$, the step size is increased. When the input is varying slowly, the step size is reduced. Then the method is called *Adaptive Delta Modulation (ADM)*.

The adaptive delta modulators can take continuous changes in step size or discrete changes in step size.

Transmitter and Receiver

Fig. (a) shows the transmitter and (b) shows receiver of adaptive delta modulator. The logic for step size control is added in the diagram. The step size increases or decreases according to certain rule depending on one bit quantizer output.

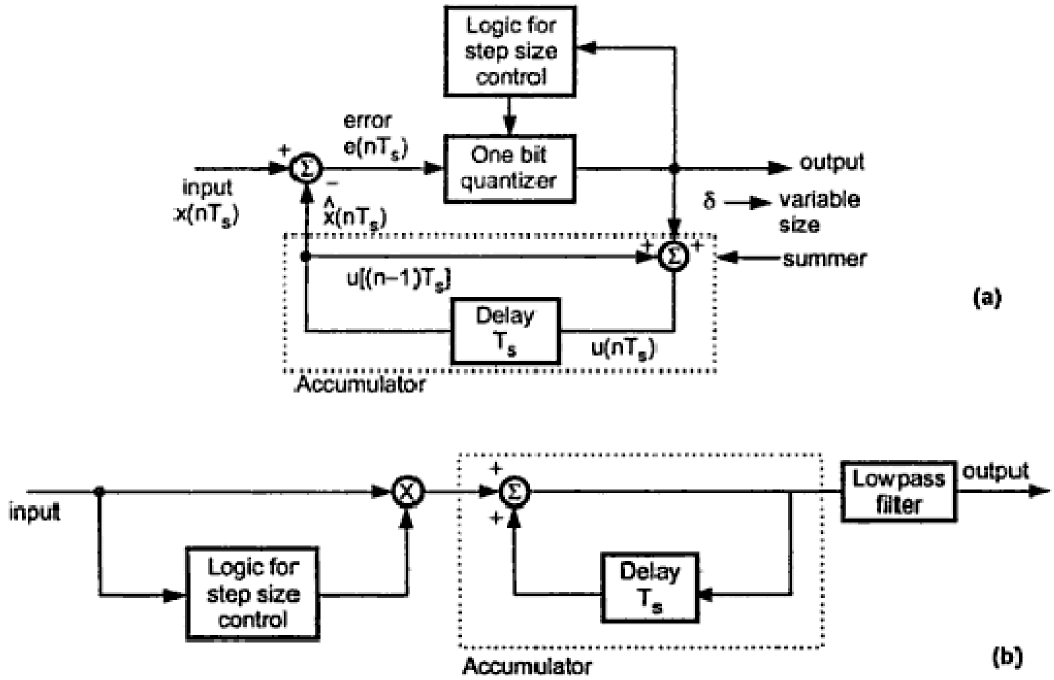


Fig. Adaptive delta modulator (a) Transmitter (b) Receiver

For example if one bit quantizer output is high (1), then step size may be doubled for next sample. If one bit quantizer output is low, then step size may be reduced by one step. Fig. shows the waveforms of adaptive delta modulator and sequence of bits transmitted.

In the receiver of adaptive delta modulator shown in Fig. (b) the first part generates the step size from each incoming bit. Exactly the same process is followed as that in transmitter. The previous input and present input decides the step size. It is then given to an accumulator which builds up staircase waveform. The low-pass filter then smoothens out the staircase waveform to reconstruct the smooth signal.

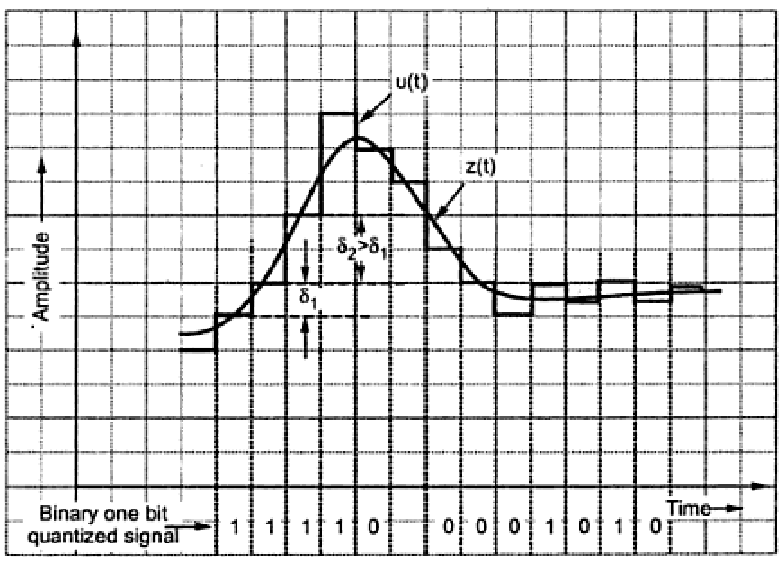


Fig. Waveforms of adaptive delta modulation

Advantages of Adaptive Delta Modulation

Adaptive delta modulation has certain advantages over delta modulation. i.e.,

1. The signal to noise ratio is better than ordinary delta modulation because of the reduction in slope overload distortion and granular noise.
2. Because of the variable step size, the dynamic range of ADM is wide.
3. Utilization of bandwidth is better than delta modulation.

Plus other advantages of delta modulation are, only one bit per sample is required and simplicity of implementation of transmitter and receiver.

Condition for Slope overload distortion occurrence:

Slope overload distortion will occur if

$$A_m > \frac{\delta}{2\pi f_m T_s}$$

where T_s is the sampling period.

Let the sine wave be represented as,

$$x(t) = A_m \sin(2\pi f_m t)$$

Slope of $x(t)$ will be maximum when derivative of $x(t)$ with respect to 't' will be maximum. The maximum slope of delta modulator is given

$$\text{Max. slope} = \frac{\text{Step size}}{\text{Sampling period}}$$

$$= \frac{\delta}{T_s} \qquad \text{.............(1)}$$

Slope overload distortion will take place if slope of sine wave is greater than slope of delta modulator i.e.

$$\max\left|\frac{d}{dt}x(t)\right| > \frac{\delta}{T_s}$$

$$\max\left|\frac{d}{dt}A_m \sin(2\pi f_m t)\right| > \frac{\delta}{T_s}$$

$$\max\left|A_m 2\pi f_m \cos(2\pi f_m t)\right| > \frac{\delta}{T_s}$$

$$A_m 2\pi f_m > \frac{\delta}{T_s}$$

or

$$\boxed{A_m > \frac{\delta}{2\pi f_m T_s}} \qquad \text{...............(2)}$$

Expression for Signal to Quantization Noise power ratio for Delta Modulation:

To obtain signal power :

slope overload distortion will not occur if

$$A_m \leq \frac{\delta}{2\pi f_m T_s}$$

Here A_m is peak amplitude of sinusoided signal

δ is the step size

f_m is the signal frequency and

T_s is the sampling period.

From above equation, the maximum signal amplitude will be,

$$A_m = \frac{\delta}{2\pi f_m T_s} \qquad \text{..................(1)}$$

Signal power is given as,

$$P = \frac{V^2}{R}$$

Here V is the rms value of the signal. Here $V = \frac{A_m}{\sqrt{2}}$. Hence above equation becomes,

$$P = \left(\frac{A_m}{\sqrt{2}}\right)^2 / R$$

Normalized signal power is obtained by taking R = 1. Hence,

$$P = \frac{A_m^2}{2}$$

Putting for A_m from equation 1

$$P = \frac{\delta^2}{8\pi^2 f_m^2 T_s^2} \qquad \text{.................(2)}$$

This is an expression for signal power in delta modulation.

(ii) To obtain noise power

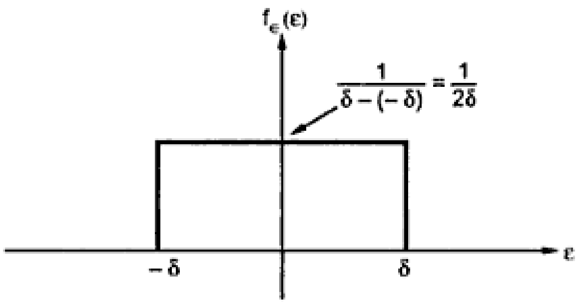


Fig. Uniform distribution of quantization error

We know that the maximum quantization error in delta modulation is equal to step size 'δ'. Let the quantization error be uniformly distributed over an interval $[-\delta, \delta]$. This is shown in Fig. From this figure the PDF of quantization error can be expressed as,

$$f_\epsilon(\varepsilon) = \begin{cases} 0 & for \quad \varepsilon < \delta \\ \frac{1}{2\delta} & for \quad -\delta < \varepsilon < \delta \\ 0 & for \quad \varepsilon > \delta \end{cases} \quad \text{................(3)}$$

The noise power is given as,

$$\text{Noise power} = \frac{V_{noise}^2}{R}$$

Here V_{noise}^2 is the mean square value of noise voltage. Since noise is defined by random variable 'ε' and PDF $f_\epsilon(\varepsilon)$, its mean square value is given as,

$$\text{mean square value} = E[\varepsilon^2] = \overline{\varepsilon^2}$$

mean square value is given as,

$$E[\varepsilon^2] = \int_{-\infty}^{\infty} \varepsilon^2 f_\epsilon(\varepsilon)\, d\varepsilon$$

From equation 3

$$E[\varepsilon^2] = \int_{-\delta}^{\delta} \varepsilon^2 \cdot \frac{1}{2\delta}\, d\varepsilon$$

$$= \frac{1}{2\delta}\left[\frac{\varepsilon^3}{3}\right]_{-\delta}^{\delta}$$

$$= \frac{1}{2\delta}\left[\frac{\delta^3}{3} + \frac{\delta^3}{3}\right] = \frac{\delta^2}{3} \quad \text{....................(4)}$$

Hence noise power will be,

$$\text{noise power} = \left(\frac{\delta^2}{3}\right) / R$$

Normalized noise power can be obtained with R = 1. Hence,

$$\text{noise power} = \frac{\delta^2}{3} \quad \text{...............(5)}$$

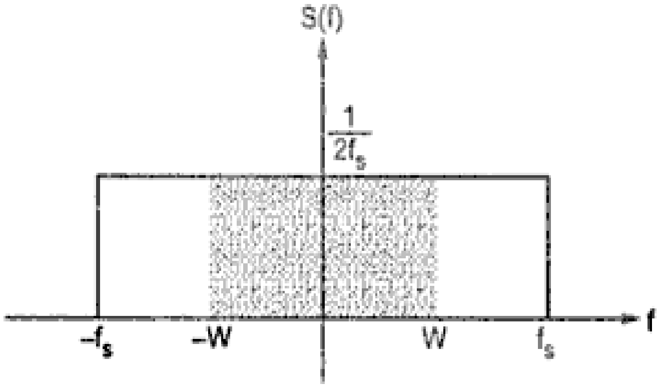


Fig. PSD of noise

This noise power is uniformly distributed over $-f_s$ to f_s range. This is illustrated in Fig. At the output of delta modulator receiver there is lowpass reconstruction filter whose cutoff frequency is 'W'. This cutoff frequency is equal to highest signal frequency. The reconstruction filter passes part of the noise power at the output as Fig. From the geometry of Fig. output noise power will be,

$$\text{Output noise power} = \frac{W}{f_s} \times \text{noise power} = \frac{W}{f_s} \times \frac{\delta^2}{3}$$

We know that $f_s = \frac{1}{T_s}$, hence above equation becomes,

$$\text{Output noise power} = \frac{W T_s \delta^2}{3} \qquad \text{.................(6)}$$

(iii) To obtain signal to noise power ratio

Signal to noise power ratio at the output of delta modulation receiver is given as,

$$\frac{S}{N} = \frac{\textit{Normalized signal power}}{\textit{Normalized noise power}}$$

From equation 2. and equation 6

$$\frac{S}{N} = \frac{\frac{\delta^2}{8\pi^2 f_m^2 T_s^2}}{\frac{W T_s \delta^2}{3}}$$

$$\boxed{\frac{S}{N} = \frac{3}{8\pi^2 W f_m^2 T_s^3}} \qquad \text{.............(7)}$$

This is an expression for signal to noise power ratio in delta modulation.

MODULE-V

DIGITAL BINARY CARRIER MODULATION SCHEMES

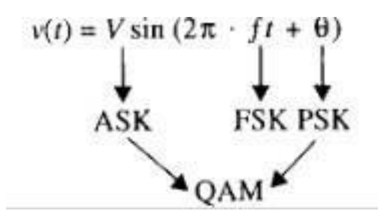
Digital Modulation provides more information capacity, high data security, quicker system availability with great quality communication. Hence, digital modulation techniques have a greater demand, for their capacity to convey larger amounts of data than analog ones.

There are many types of digital modulation techniques and we can even use a combination of these techniques as well. In this chapter, we will be discussing the most prominent digital modulation techniques.

if the information signal is digital and the amplitude (lV of the carrier is varied proportional to the information signal, a digitally modulated signal called amplitude shift keying (ASK) is produced.

If the frequency (f) is varied proportional to the information signal, frequency shift keying (FSK) is produced, and if the phase of the carrier (0) is varied proportional to the information signal,

phase shift keying (PSK) is produced. If both the amplitude and the phase are varied proportional to the information signal, quadrature amplitude modulation (QAM) results. ASK, FSK, PSK, and QAM are all forms of digital modulation:



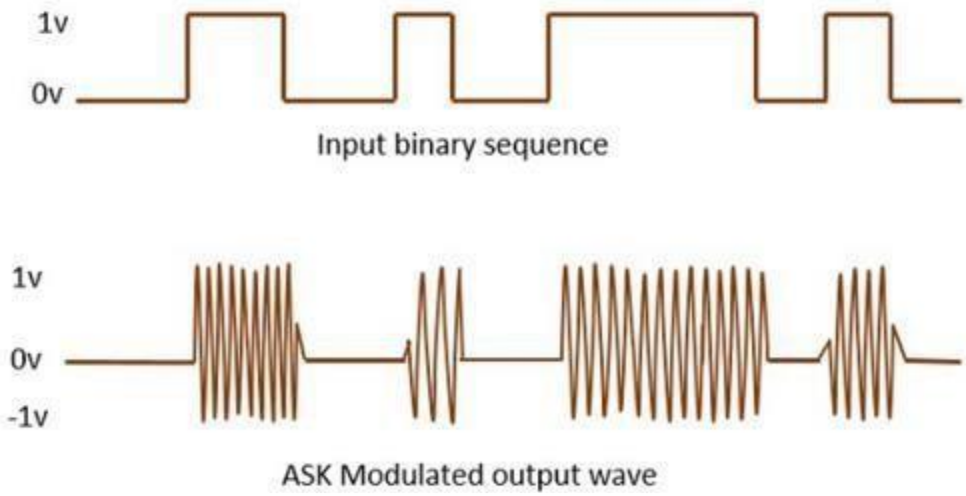
a simplified block diagram for a digital modulation system.

Amplitude Shift Keying

The amplitude of the resultant output depends upon the input data whether it should be a zero level or a variation of positive and negative, depending upon the carrier frequency.

Amplitude Shift Keying (ASK) is a type of Amplitude Modulation which represents the binary data in the form of variations in the amplitude of a signal.

Following is the diagram for ASK modulated waveform along with its input.



Any modulated signal has a high frequency carrier. The binary signal when ASK is modulated, gives a zero value for LOW input and gives the carrier output for HIGH input.
Mathematically, amplitude-shift keying is

$$v_{(ask)}(t) = [1 + v_m(t)]\left[\frac{A}{2}\cos(\omega_c t)\right]$$

where vask(t) = amplitude-shift keying wave

vm(t) = digital information (modulating) signal (volts)

A/2 = unmodulated carrier amplitude (volts)

ωc= analog carrier radian frequency (radians per second, 2πfct)

In above Equation, the modulating signal [vm(t)] is a normalized binary waveform, where + 1 V = logic 1 and -1 V = logic 0. Therefore, for a logic 1 input, vm(t) = + 1 V, Equation 2.12 reduces to

$$v_{(ask)}(t) = [1 + 1]\left[\frac{A}{2}\cos(\omega_c t)\right]$$

$$= A\cos(\omega_c t)$$

Mathematically, amplitude-shift keying is (2.12) where vask(t) = amplitude-shift keying wave vm(t) = digital information (modulating) signal (volts) A/2 = unmodulated carrier amplitude (volts) ωc= analog carrier radian frequency (radians per second, 2πfct) In Equation 2.12, the modulating signal [vm(t)] is a normalized binary waveform, where + 1 V = logic 1 and -1 V = logic 0. Therefore, for a logic 1 input, vm(t) = + 1 V, Equation 2.12 reduces to and for a logic 0 input, vm(t) = -1 V,Equation reduces to

$$v_{(ask)}(t) = [1 - 1]\left[\frac{A}{2}\cos(\omega_c t)\right]$$

Thus, the modulated wave vask(t),is either A cos(ωct) or 0. Hence, the carrier is either "on “or "off," which is why amplitude-shift keying is sometimes referred to as on-off keying (OOK).

it can be seen that for every change in the input binary data stream, there is one change in the ASK waveform, and the time of one bit (tb) equals the time of one analog signaling element (t,).

B = fb/1 = fb baud = fb/1 = fb

Example :

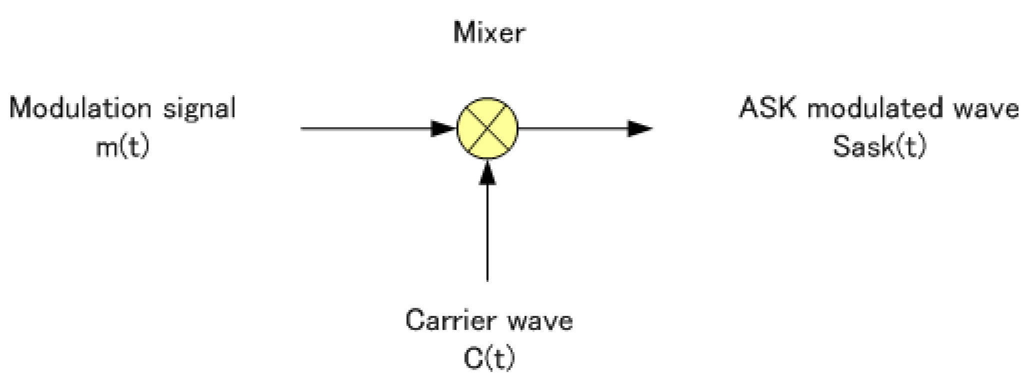
Determine the baud and minimum bandwidth necessary to pass a 10 kbps binary signal using amplitude shift keying. 10Solution For ASK, N = 1, and the baud and minimum bandwidth are determined from Equations 2.11 and 2.10, respectively:

B = 10,000 / 1 = 10,000

baud = 10, 000 /1 = 10,000

The use of amplitude-modulated analog carriers to transport digital information is a relatively low-quality, low-cost type of digital modulation and, therefore, is seldom used except for very low-speed telemetry circuits.

ASK TRANSMITTER:



The input binary sequence is applied to the product modulator. The product modulator amplitude modulates the sinusoidal carrier .it passes the carrier when input bit is ‘1’ .it blocks the carrier when input bit is ‘0.’

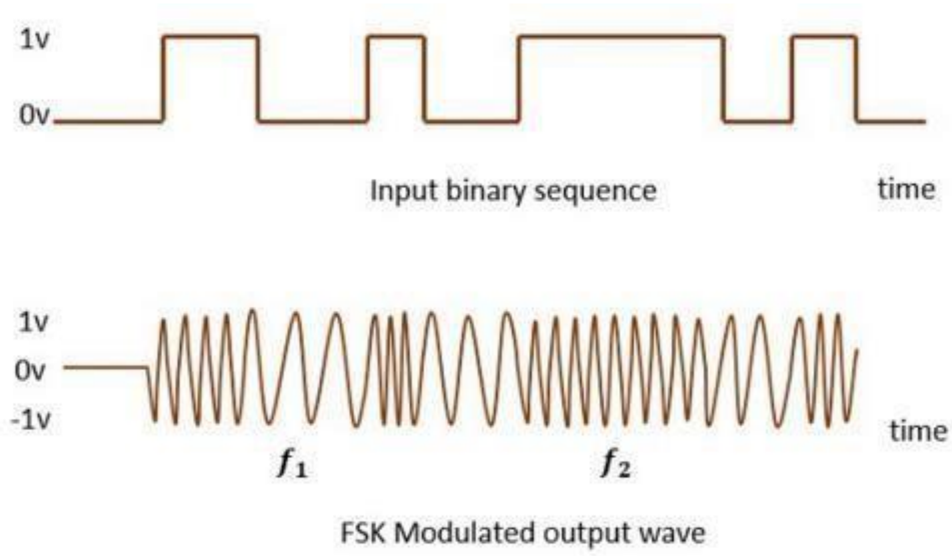
Coherent ASK DETECTOR:

FREQUENCYSHIFT KEYING

The frequency of the output signal will be either high or low, depending upon the input data applied.

Frequency Shift Keying (FSK) is the digital modulation technique in which the frequency of the carrier signal varies according to the discrete digital changes. FSK is a scheme of frequency modulation.

Following is the diagram for FSK modulated waveform along with its input.



The output of a FSK modulated wave is high in frequency for a binary HIGH input and is low in frequency for a binary LOW input. The binary 1s and 0s are called **Mark** and **Space frequencies**.

FSK is a form of constant-amplitude angle modulation similar to standard frequency modulation (FM) except the modulating signal is a binary signal that varies between two discrete voltage levels rather than a continuously changing analog waveform.Consequently, FSK is sometimes called *binary FSK* (BFSK). The general expression for FSK is

$$v_{fsk}(t) = V_c \cos\{2\pi[f_c + v_m(t)\,\Delta f]t\}$$

where

$v_{fsk}(t)$ = binary FSK waveform

V_c = peak analog carrier amplitude (volts)

f_c = analog carrier center frequency(hertz)

f=peak change (shift)in the analog carrier frequency(hertz)

$v_m(t)$ = binary input (modulating) signal (volts)

From Equation 2.13, it can be seen that the peak shift in the carrier frequency (f) is proportional to the amplitude of the binary input signal ($v_m[t]$), and the direction of the shift is determined by the polarity.

The modulating signal is a normalized binary waveform where a logic 1 = + 1 V and a logic 0 = -1 V. Thus, for a logic l input, $v_m(t)$ = + 1, Equation 2.13 can be rewritten as

$$v_{fsk}(t) = V_c \cos[2\pi(f_c + \Delta f)t]$$

For a logic 0 input, $v_m(t)$ = -1, Equation becomes

$$v_{fsk}(t) = V_c \cos[2\pi(f_c - \Delta f)t]$$

With binary FSK, the carrier center frequency (f_c) is shifted (deviated) up and down in the frequency domain by the binary input signal as shown in Figure 2-3.

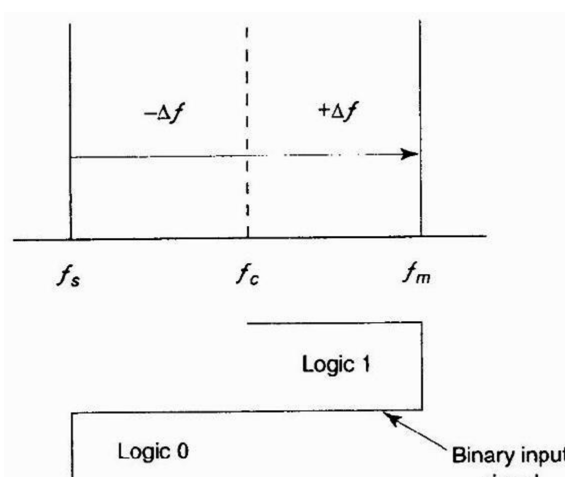


FIGURE: FSK in the frequency domain

As the binary input signal changes from a logic 0 to a logic 1 and vice versa, the output frequency shifts between two frequencies: a mark, or logic 1 frequency (f_m), and a space, or logic 0 frequency (f_s). The mark and space frequencies are separated from the carrier frequency by the peak frequency deviation (f) and from each other by 2 f.

Frequency deviation is illustrated in Figure 2-3 and expressed mathematically as

$$f = |f_m - f_s| / 2 \qquad (2.14)$$

where f = frequency deviation (hertz)

$|f_m - f_s|$ = absolute difference between the mark and space frequencies (hertz)

Figure 2-4a shows in the time domain the binary input to an FSK modulator and the corresponding FSK output.

When the binary input *(f_b)* changes from a logic 1 to a logic 0 and vice versa, the FSK output frequency shifts from a mark (f_m) to a space (f_s) frequency and vice versa.

In Figure 2-4a, the mark frequency is the higher frequency (f_c + f) and the space frequency is the lower frequency (f_c - f), although this relationship could be just the opposite.

Figure 2-4b shows the truth table for a binary FSK modulator. The truth table shows the input and output possibilities for a given digital modulation scheme.

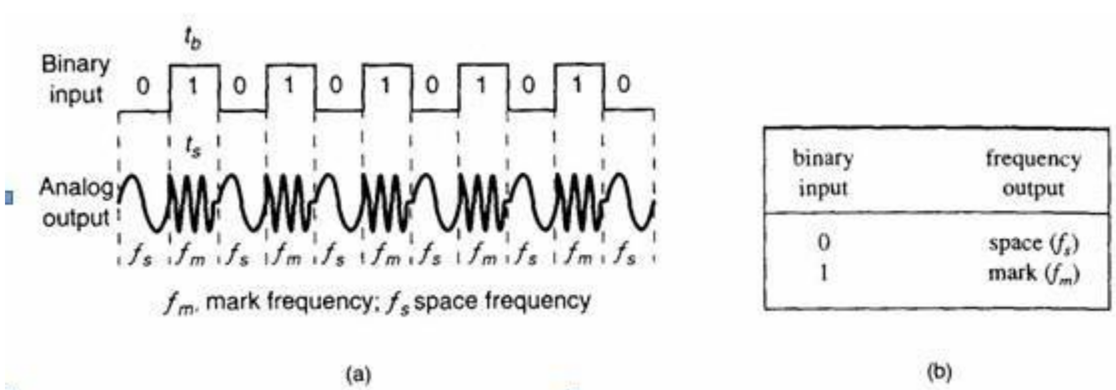


FIGURE 2-4 FSK in the time domain: (a) waveform: (b) truth table

FSK Bit Rate, Baud, and Bandwidth

In Figure 2-4a, it can be seen that the time of one bit *(t_b)* is the same as the time the FSK output is a mark of space frequency *(t_s)*. Thus, the bit time equals the time of an FSK signaling element, and the bit rate equals the baud.

The baud for binary FSK can also be determined by substituting N = 1 in Equation 2.11:

baud = $f_b / 1 = f_b$

The minimum bandwidth for FSK is given as

$$B = |(f_s - f_b) - (f_m - f_b)|$$

$$= |(f_s - f_m)| + 2f_b$$

and since $|(f_s - f_m)|$ equals 2 f, the minimum bandwidth can be approximated as

$$B = 2(f + f_b) \qquad (2.15)$$

where

B= minimum Nyquist bandwidth (hertz)

f= frequency deviation $|(f_m - f_s)|$ (hertz)

f_b = input bit rate (bps)

Example 2-2

Determine (a) the peak frequency deviation, (b) minimum bandwidth, and (c) baud for a binary FSK signal with a mark frequency of 49 kHz, a space frequency of 51 kHz, and an input bit rate of 2 kbps.

Solution

a. The peak frequency deviation is determined from Equation 2.14:

$$f = |149\text{kHz} - 51\text{ kHz}| / 2 = 1\text{ kHz}$$

b. The minimum bandwidth is determined from Equation 2.15:

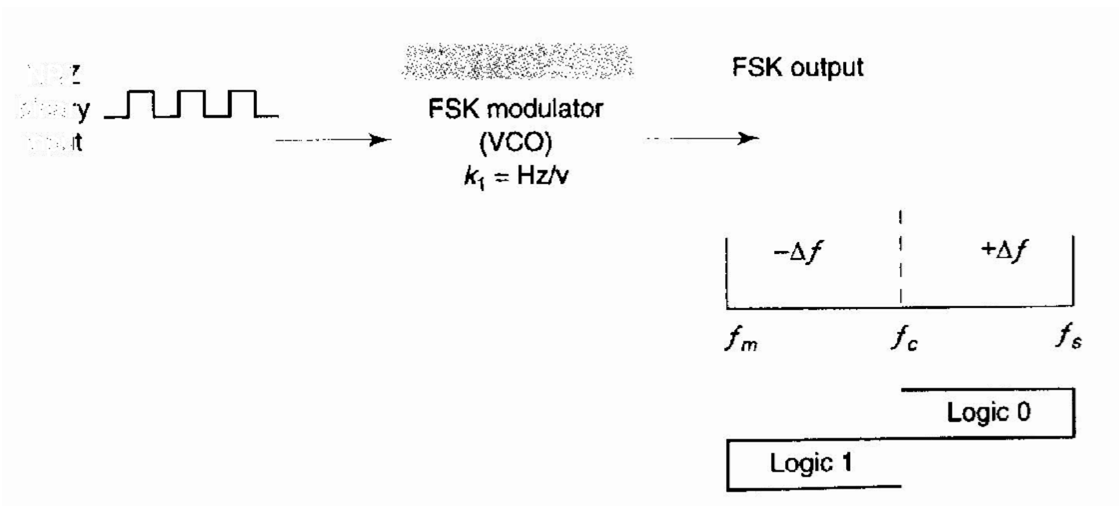
$$B = 2(100 + 2000)$$
$$= 6\text{ kHz}$$

c. For FSK, N = 1, and the baud is determined from Equation 2.11 as

$$\text{baud} = 2000 / 1 = 2000$$

FSK TRANSMITTER:

Figure 2-6 shows a simplified binary FSK modulator, which is very similar to a conventional FM modulator and is very often a voltage-controlled oscillator (VCO).The center frequency (f_c) is chosen such that it falls halfway between the mark and space frequencies.



A logic 1 input shifts the VCO output to the mark frequency, and a logic 0 input shifts the VCO output to the space frequency. Consequently, as the binary input signal changes back and forth between logic 1 and logic 0 conditions, the VCO output shifts or deviates back and forth between the mark and space frequencies.

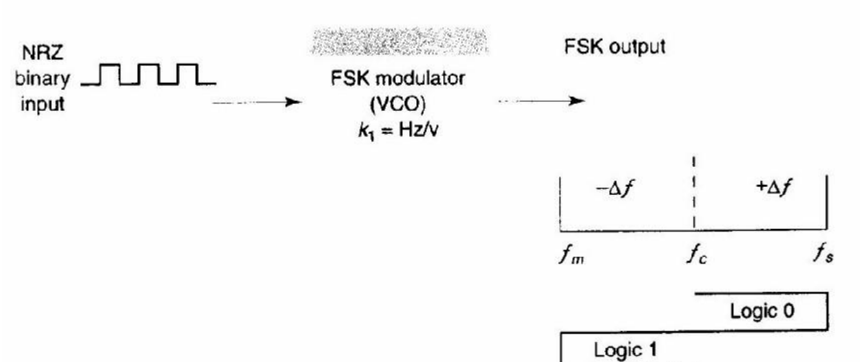


FIGURE 2-6 FSK modulator

A VCO-FSK modulator can be operated in the sweep mode where the peak frequency deviation is simply the product of the binary input voltage and the deviation sensitivity of the VCO.

With the sweep mode of modulation, the frequency deviation is expressed mathematically as

$$f = v_m(t)k_l \qquad (2\text{-}19)$$

$v_m(t)$ = peak binary modulating-signal voltage (volts)

k_l = deviation sensitivity (hertz per volt).

FSK Receiver

FSK demodulation is quite simple with a circuit such as the one shown in Figure 2-7.

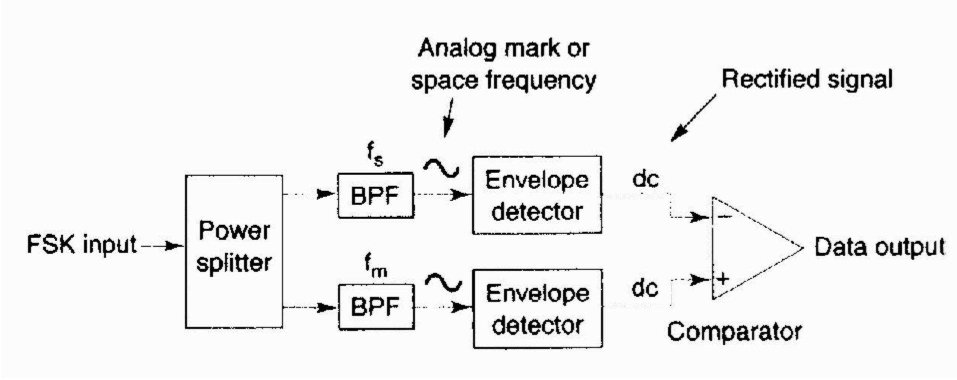


FIGURE 2-7 Noncoherent FSK demodulator

The FSK input signal is simultaneously applied to the inputs of both bandpass filters (BPFs) through a power splitter.The respective filter passes only the mark or only the space frequency on to its respective envelope detector.The envelope detectors, in turn, indicate the total power in each passband, and the comparator responds to the largest of the two powers.This type of FSK detection is referred to as noncoherent detection.

Figure 2-8 shows the block diagram for a coherent FSK receiver.The incoming FSK signal is multiplied by a recovered carrier signal that has the exact same frequency and phase as the transmitter reference.

However, the two transmitted frequencies (the mark and space frequencies) are not generally continuous; it is not practical to reproduce a local reference that is coherent with both of them. Consequently, coherent FSK detection is seldom used.

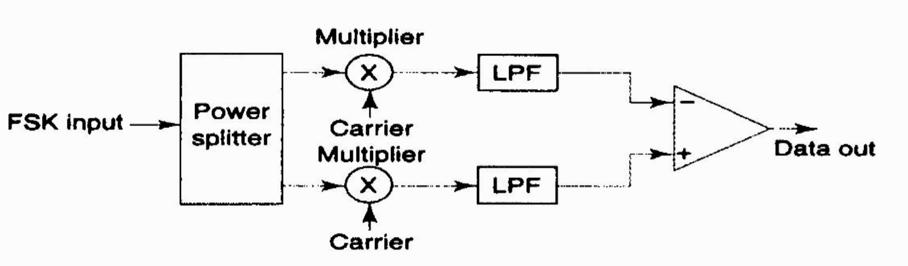
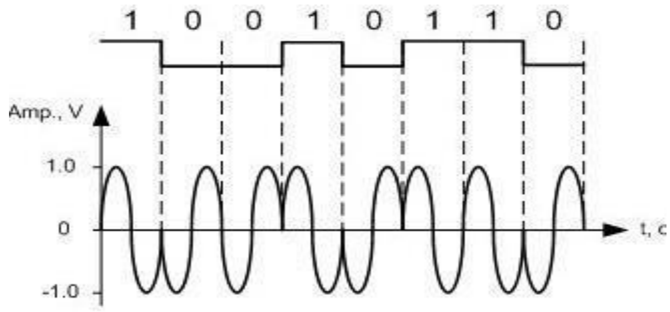


FIGURE 2-8 Coherent FSK demodulator

PHASESHIFT KEYING:

The phase of the output signal gets shifted depending upon the input. These are mainly of two types, namely BPSK and QPSK, according to the number of phase shifts. The other one is DPSK which changes the phase according to the previous value.



Phase shift keying (PSK)

Phase Shift Keying (PSK) is the digital modulation technique in which the phase of the carrier signal is changed by varying the sine and cosine inputs at a particular time. PSK technique is widely used for wireless LANs, bio-metric, contactless operations, along with RFID and Bluetooth communications.

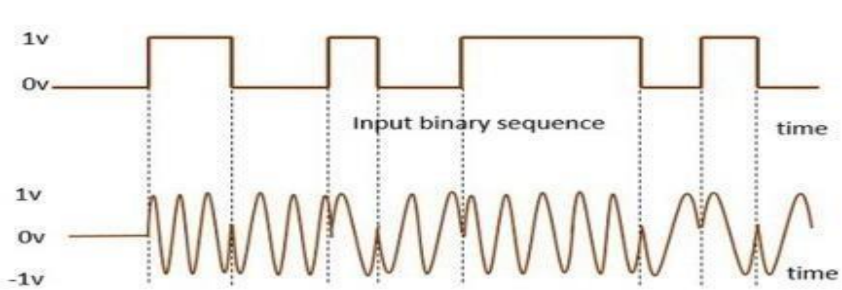
PSK is of two types, depending upon the phases the signal gets shifted. They are −

Binary Phase Shift Keying (BPSK)

This is also called as **2-phase PSK** (or) **Phase Reversal Keying**. In this technique, the sine wave carrier takes two phase reversals such as 0° and 180°.

BPSK is basically a DSB-SC (Double Sideband Suppressed Carrier) modulation scheme, for message being the digital information.

Following is the image of BPSK Modulated output wave along with its input.



BPSK Modulated output wave

Binary Phase-Shift Keying

The simplest form of PSK is *binary phase-shift keying* (BPSK), where N = 1 and *M = 2.*Therefore, with BPSK, two phases ($2^1 = 2$) are possible for the carrier.One phase represents a logic 1, and the other phase represents a logic 0. As the input digital signal changes state (i.e., from a 1 to a 0 or from a 0 to a 1), the phase of the output carrier shifts between two angles that are separated by 180°.

Hence, other names for BPSK are *phase reversal keying* (PRK) and *biphase modulation.* BPSK is a form of square-wave modulation of a *continuous wave (CW)* signal.

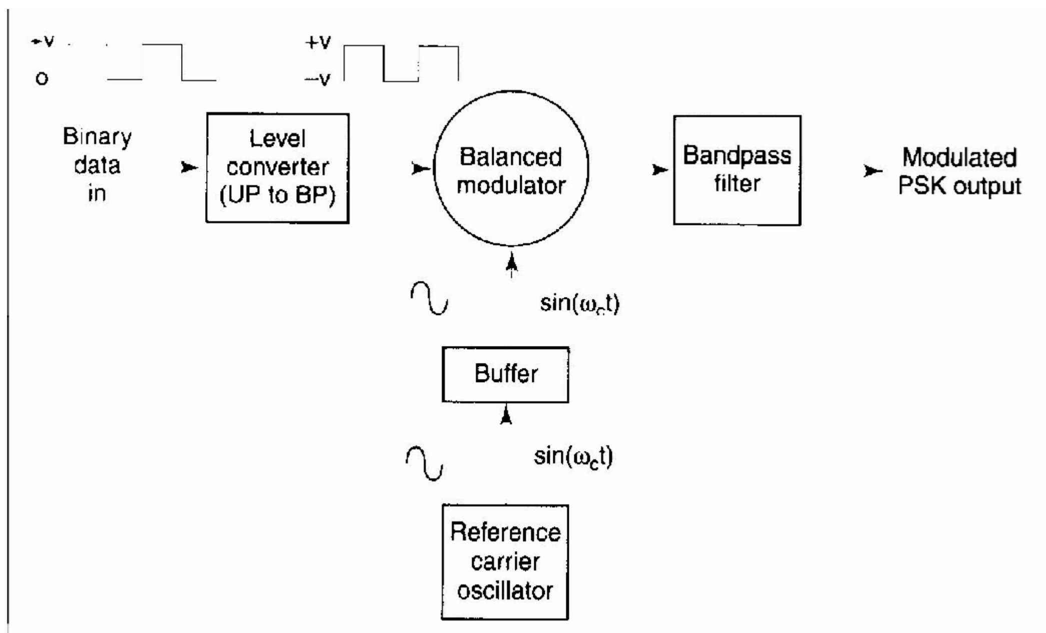


FIGURE 2-12 BPSK transmitter

BPSK TRANSMITTER:

Figure 2-12 shows a simplified block diagram of a BPSK transmitter. The balanced modulator acts as a phase reversing switch. Depending on the logic condition of the digital input, the carrier is transferred to the output either in phase or 180° out of phase with the reference carrier oscillator.

Figure 2-13 shows the schematic diagram of a balanced ring modulator. The balanced modulator has two inputs: a carrier that is in phase with the reference oscillator and the binary digital data. For the balanced modulator to operate properly, the digital input voltage must be much greater than the peak carrier voltage.

This ensures that the digital input controls the on/off state of diodes D1 to D4. If the binary input is a logic 1(positive voltage), diodes D 1 and D2 are forward biased and on, while diodes D3 and D4 are reverse biased and off (Figure 2-13b). With the polarities shown, the carrier voltage is developed across transformer T2 in phase with the carrier voltage across T

1. Consequently, the output signal is in phase with the reference oscillator.

If the binary input is a logic 0 (negative voltage), diodes Dl and D2 are reverse biased and off, while diodes D3 and D4 are forward biased and on (Figure 9-13c). As a result, the carrier voltage is developed across transformer T2 180° out of phase with the carrier voltage across T 1.

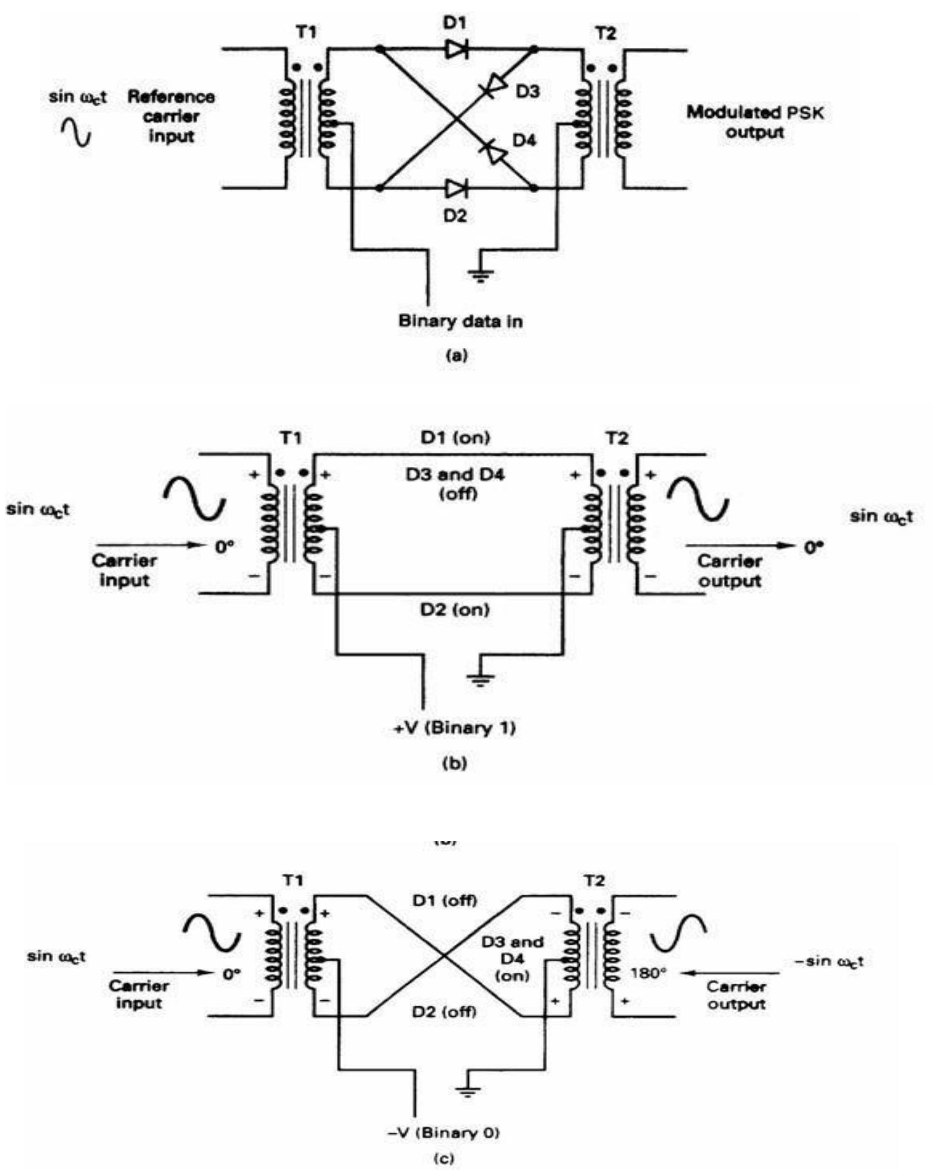


FIGURE 9-13 (a) Balanced ring modulator; (b) logic 1 input; (c) logic 0 input

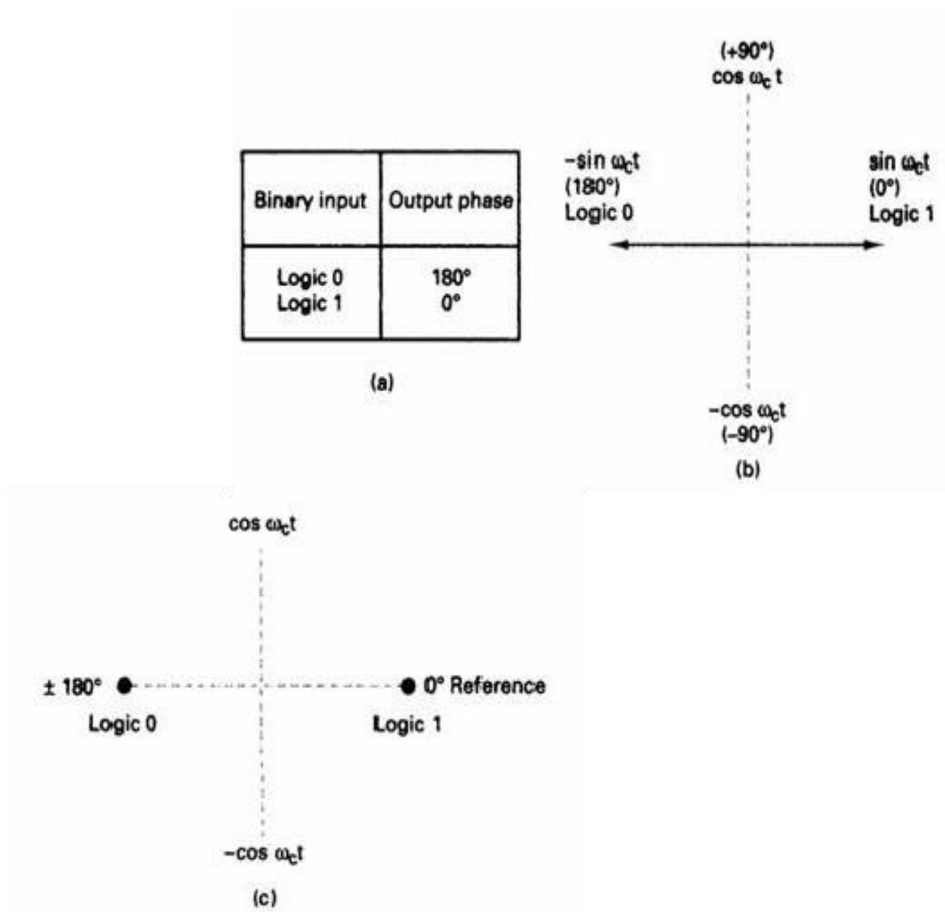


FIGURE 2-14 BPSK modulator: (a) truth table; (b) phasor diagram; (c) constellation diagram

BANDWIDTH CONSIDERATIONS OF BPSK:

In a BPSK modulator. the carrier input signal is multiplied by the binary data.

If + 1 V is assigned to a logic 1 and -1 V is assigned to a logic 0, the input carrier (sin $\omega_c t$) is multiplied by either a + or - 1 .

The output signal is either + 1 sin $\omega_c t$ or -1 sin $\omega_c t$ the first represents a signal that is *in phase* with the reference oscillator, the latter a signal that is 180° out of phase with the reference oscillator.Each time the input logic condition changes, the output phase changes.

Mathematically, the output of a BPSK modulator is proportional to

$$\text{BPSK output} = [\sin(2\pi f_a t)] \times [\sin(2\pi f_c t)] \qquad (2.20)$$

where

f_a = maximum fundamental frequency of binary input (hertz)

f_c = reference carrier frequency (hertz)

Solving for the trig identity for the product of two sine functions,

$0.5\cos[2\pi(f_c - f_a)t] - 0.5\cos[2\pi(f_c + f_a)t]$

Thus, the minimum double-sided Nyquist bandwidth *(B) is*

$$\begin{array}{c} f_c + f_a \\ -(f_c + f_a) \end{array} \quad \text{or} \quad \begin{array}{c} f_c + f_a \\ \underline{-f_c + f_a} \\ 2f_a \end{array}$$

and because $f_a = f_b / 2$, where f_b = input bit rate,

where *B is* the minimum double-sided Nyquist bandwidth**.**

Figure 2-15 shows the output phase-versus-time relationship for a BPSK waveform. Logic 1 input produces an analog output signal with a 0° phase angle, and a logic 0 input produces an analog output signal with a 180° phase angle.

As the binary input shifts between a logic 1 and a logic 0 condition and vice versa, the phase of the BPSK waveform shifts between 0° and 180°, respectively.

BPSK signaling element *(t_s)* is equal to the time of one information bit *(t_b)*, which indicates that the bit rate equals the baud.

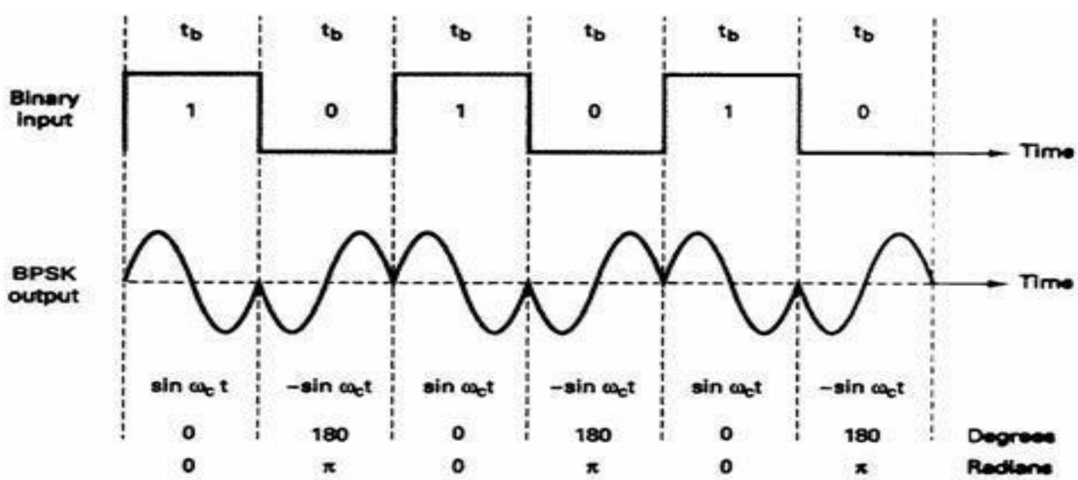


FIGURE 2-15 Output phase-versus-time relationship for a BPSK modulator

Example:

For a BPSK modulator with a carrier frequency of 70 MHz and an input bit rate of 10 Mbps, determine the maximum and minimum upper and lower side frequencies, draw the output spectrum, de-termine the minimum Nyquist bandwidth, and calculate the baud..

Solution

Substituting into Equation 2-20 yields

$$\text{output} = [\sin(2\pi f_a t)] \times [\sin(2\pi f_c t)];\ f_a = f_b / 2 = 5\text{ MHz}$$

$$= [\sin 2\pi(5\text{MHz})t)] \times [\sin 2\pi(70\text{MHz})t)]$$

$$= 0.5\cos[2\pi(70\text{MHz} - 5\text{MHz})t] - 0.5\cos[2\pi(70\text{MHz} + 5\text{MHz})t]$$

lower side frequency upper side frequency

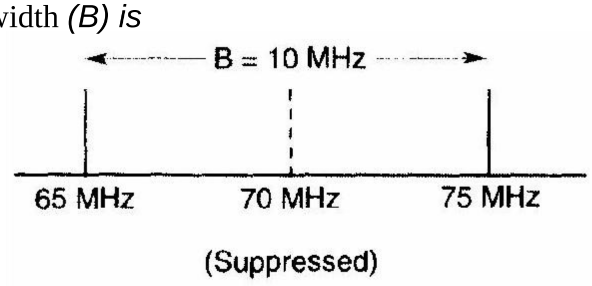
Minimum lower side frequency (LSF):

LSF=70MHz - 5MHz = 65MHz

Maximum upper side frequency (USF):

USF = 70 MHz + 5 MHz = 75 MHz

Therefore, the output spectrum for the worst-case binary input conditions *is* as follows: The minimum Nyquist bandwidth *(B) is*



B = 75 MHz - 65 MHz = 10 MHz

and the baud = f_b or 10 megabaud.

BPSK receiver:.

Figure 2-16 shows the block diagram of a BPSK receiver.

The input signal maybe+ sin $\omega_c t$ or - sin $\omega_c t$ .The coherent carrier recovery circuit detects and regenerates a carrier signal that is both frequency and phase coherent with the original transmit carrier.

The balanced modulator is a product detector; the output is the product d the two inputs (the BPSK signal and the recovered carrier).

The low-pass filter (LPF) operates the recovered binary data from the complex demodulated signal.

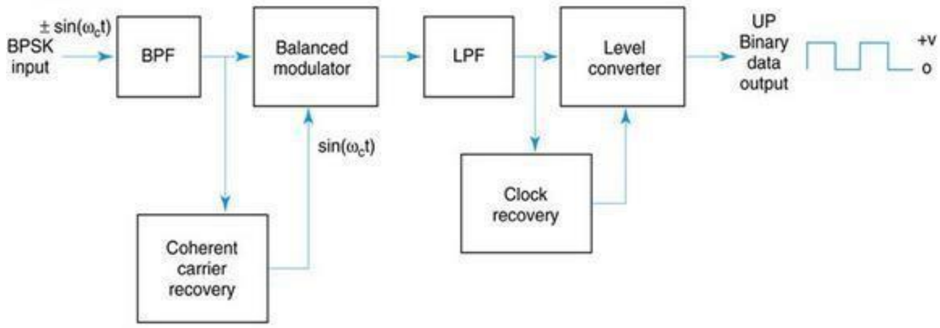


FIGURE 2-16 Block diagram of a BPSK receiver

Mathematically, the demodulation process is as follows.

For a BPSK input signal of + sin $\omega_c t$ (logic 1), the output of the balanced modulator is

$$\text{output} = (\sin \omega_c t)(\sin \omega_c t) = \sin^2\omega_c t \qquad (2.21)$$

or

$$\sin^2\omega_c t = 0.5(1 - \cos 2\omega_c t) = 0.5 - 0.5\cos 2\omega_c t$$

filtered out

leaving output = + 0.5 V = logic 1

It can be seen that the output of the balanced modulator contains a positive voltage (+[1/2]V) and a cosine wave at twice the carrier frequency (2 $\omega_c t$).

The LPF has a cutoff frequency much lower than 2 $\omega_c t$, and, thus, blocks the second harmonic of the carrier and passes only the positive constant component. A positive voltage represents a demodulated logic 1.

For a BPSK input signal of -sin $\omega_c t$ (logic 0), the output of the balanced modulator is

$$\text{output} = (-\sin \omega_c t)(\sin \omega_c t) = \sin^2\omega_c t$$

or

$$\sin^2\omega_c t = -0.5(1 - \cos 2\omega_c t) = 0.5 + 0.5\cos 2\omega_c t$$

filtered out

leaving

output = - 0.5 V = logic 0

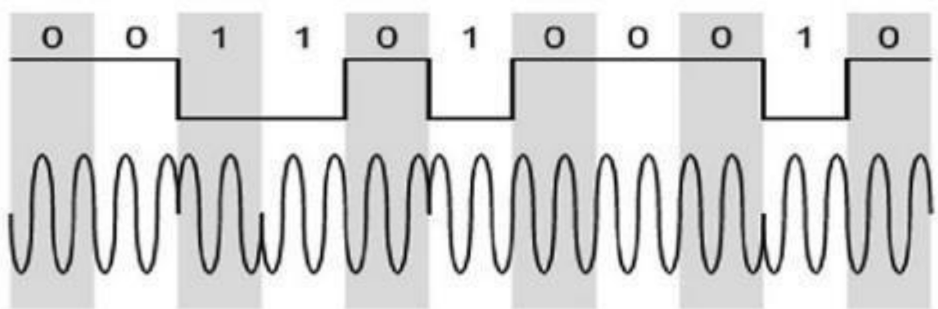
The output of the balanced modulator contains a negative voltage (-[l/2]V) and a cosine wave at twice the carrier frequency ($2\omega_c t$).

Again, the LPF blocks the second harmonic of the carrier and passes only the negative constant component. A negative voltage represents a demodulated logic 0.

DIFFERENTIAL PHASE SHIFT KEYING (DPSK):

In DPSK (Differential Phase Shift Keying) the phase of the modulated signal is shifted relative to the previous signal element. No reference signal is considered here. The signal phase follows the high or low state of the previous element. This DPSK technique doesn't need a reference oscillator.

The following figure represents the model waveform of DPSK.



It is seen from the above figure that, if the data bit is LOW i.e., 0, then the phase of the signal is not reversed, but is continued as it was. If the data is HIGH i.e., 1, then the phase of the signal is reversed, as with NRZI, invert on 1 (a form of differential encoding).

If we observe the above waveform, we can say that the HIGH state represents an **M** in the modulating signal and the LOW state represents a **W** in the modulating signal.

The word binary represents two-bits. **M** simply represents a digit that corresponds to the number of conditions, levels, or combinations possible for a given number of binary variables.

This is the type of digital modulation technique used for data transmission in which instead of one-bit, two or **more bits are transmitted at a time**. As a single signal is used for multiple bit transmission, the channel bandwidth is reduced.

DBPSK TRANSMITTER.:

Figure 2-37a shows a simplified block diagram of a *differential binary phase-shift keying* (DBPSK) transmitter. An incoming information bit is XNORed with the preceding bit prior to entering the BPSK modulator (balanced modulator).

For the first data bit, there is no preceding bit with which to compare it. Therefore, an initial reference bit is assumed. Figure 2-37b shows the relationship between the input data, the XNOR output data, and the phase at the output of the balanced modulator. If the initial reference bit is assumed a logic 1, the output from the XNOR circuit is simply the complement of that shown.

In Figure 2-37b, the first data bit is XNORed with the reference bit. If they are the same, the XNOR output is a logic 1; if they are different, the XNOR output is a logic 0. The balanced modulator operates the same as a conventional BPSK modulator; a logic I produces $+\sin \omega_c t$ at the output, and A logic 0 produces $-\sin \omega_c t$ at the output.

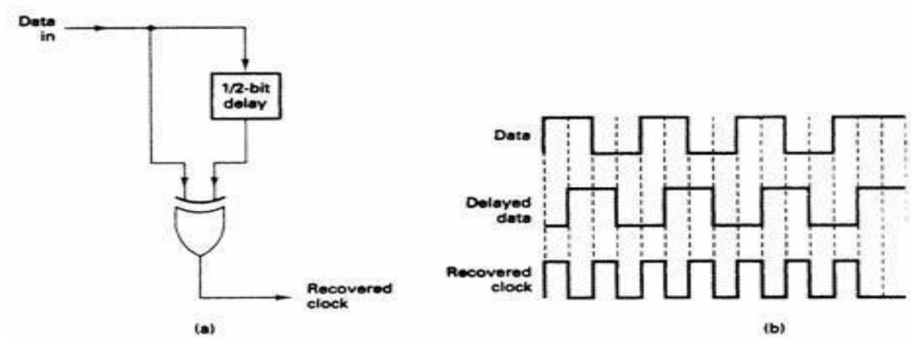


FIGURE 9-40 (a) Clock recovery circuit; (b) timing diagram

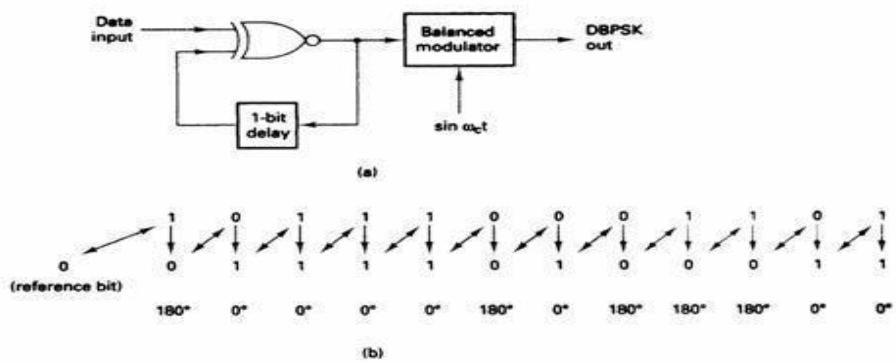


FIGURE 2-37 DBPSK modulator (a) block diagram (b) timing diagram

BPSK RECEIVER:

Figure 9-38 shows the block diagram and timing sequence for a DBPSK receiver. The received signal is delayed by one bit time, then compared with the next signaling element in the balanced modulator. If they are the same. J logic 1(+ voltage) is generated. If they are different, a logic 0 (-voltage) is generated. [f the reference phase is incorrectly assumed, only the first demodulated bit is in error. Differential encoding can be implemented with higher-than-binary digital modulation schemes, although the differential algorithms are much more complicated than for DBPS K.

The primary advantage of DBPSK is the simplicity with which it can be implemented. With DBPSK, no carrier recovery circuit is needed. A disadvantage of DBPSK is, that it requires between 1 dB and 3 dB more signal-to-noise ratio to achieve the same bit error rate as that of absolute PSK.

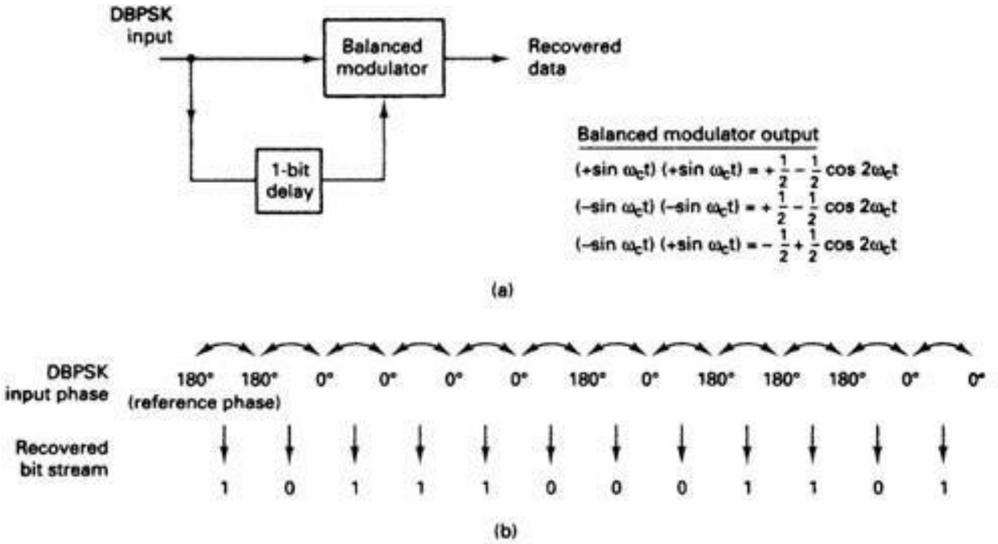


FIGURE 2-38 DBPSK demodulator: (a) block diagram; (b) timing sequence

COHERENT RECEPTION OF FSK:

The coherent demodulator for the coherent FSK signal falls in the general form of coherent demodulators described in Appendix B. The demodulator can be implemented with two correlators as shown in Figure 3.5, where the two reference signals are cos(27r f t) and cos(27r fit). They must be synchronized with the received signal. The receiver is optimum in the sense that it minimizes the error probability for equally likely binary signals. Even though the receiver is rigorously derived in Appendix B, some heuristic explanation here may help understand its operation. When s 1 (t) is transmitted, the upper correlator yields a signal 1 with a positive signal component and a noise component. However, the lower correlator output 12, due to the signals' orthogonality, has only a noise component. Thus the output of the summer is most likely above zero, and the threshold detector will most likely produce a 1. When s2(t) is transmitted, opposite things happen to the two correlators and the threshold detector will most likely produce a 0. However, due to the noise nature that its values range from -00 to m, occasionally the noise amplitude might overpower the signal amplitude, and then detection errors will happen. An alternative to Figure 3.5 is to use just one correlator with the reference signal cos (27r f t) - cos(2s f2t) (Figure 3.6). The correlator in Figure

can be replaced by a matched filter that matches cos(27r fit) - cos(27r f2t) (Figure 3.7). All implementations are equivalent in terms of error performance (see Appendix B). Assuming an AWGN channel, the received signal is

$$r(t) = s_i(t) + n(t), \quad i = 1.2$$

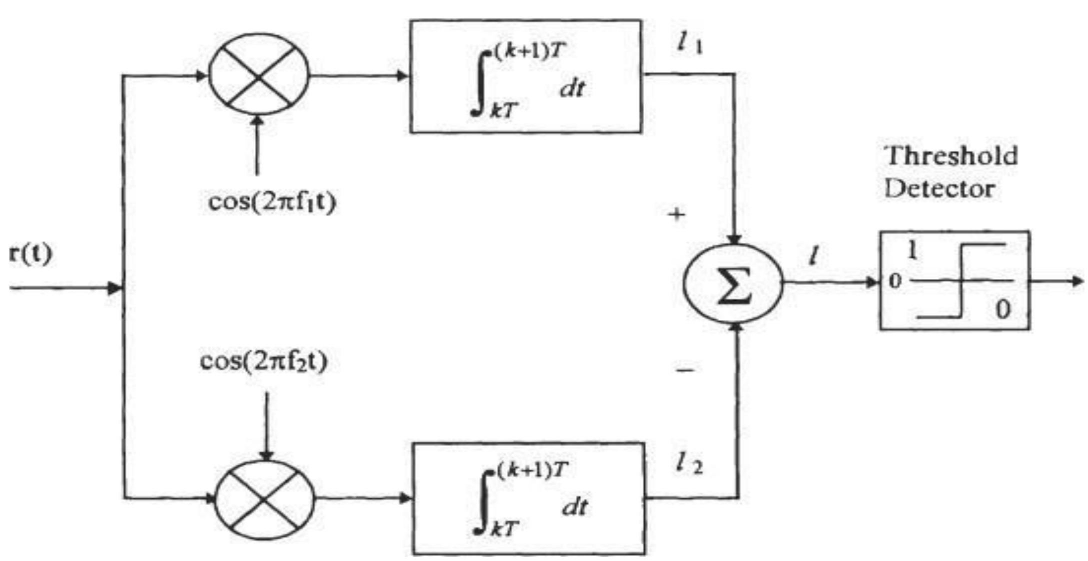
where n(t) is the additive white Gaussian noise with zero mean and a two-sided power spectral density A',/2. From (B.33) the bit error probability for any equally likely binary signals is

$$P_b = Q\left(\sqrt{\frac{E_1 + E_2 - 2\rho_{12}\sqrt{E_1 E_2}}{2N_o}}\right)$$

where No/2 is the two-sided power spectral density of the additive white Gaussian noise. For Sunde's FSK signals El = Ez = Eb, pI2 = 0 (orthogonal). thus the error probability is

$$P_b = Q\left(\sqrt{\frac{E_b}{N_o}}\right)$$

where Eb = A2T/2 is the average bit energy of the FSK signal. The above Pb is plotted in Figure 3.8 where Pb of noncoherently demodulated FSK, whose expression will be given shortly, is also plotted for comparison.



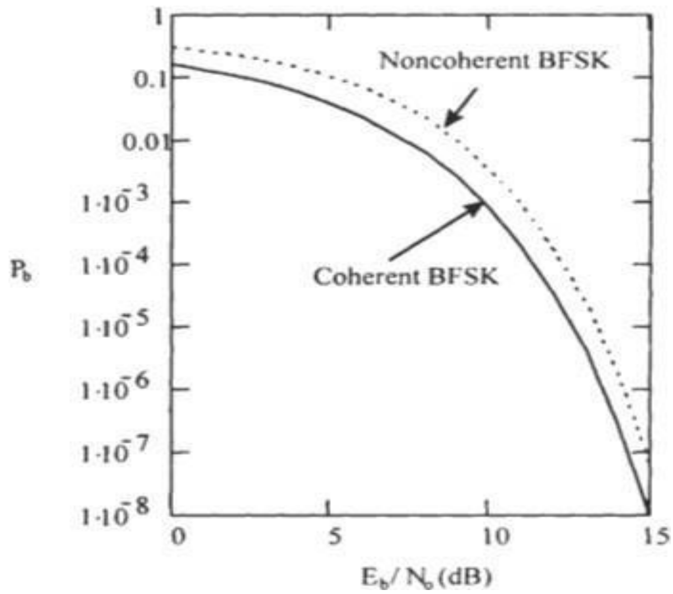


Figure: Pb of coherently and non-coherently demodulated FSK signal.

NONCOHERENT DEMODULATION AND ERROR PERFORMANCE:

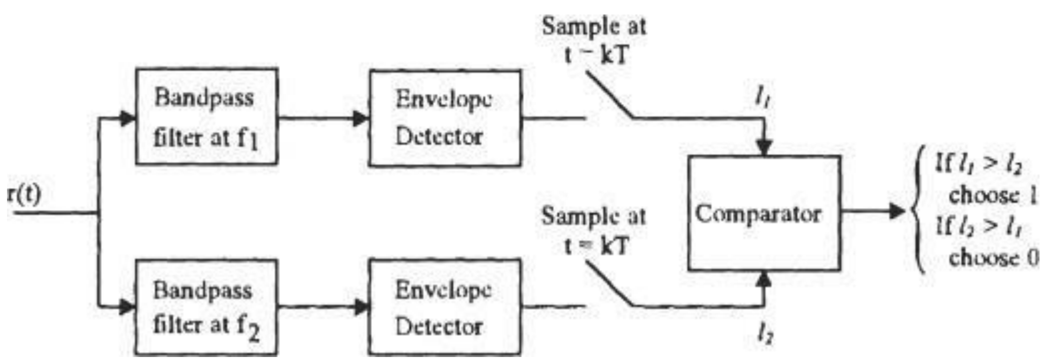
Coherently FSK signals can be noncoherently demodulated to avoid the carrier recovery. Noncoherently generated FSK can only be noncoherently demodulated. We refer to both cases as noncoherent FSK. In both cases the demodulation problem becomes a problem of detecting signals with unknown phases. In Appendix B we have shown that the optimum receiver is a quadrature receiver. It can be implemented using correlators or equivalently, matched filters. Here we assume that the binary noncoherent FSK signals are equally likely and with equal energies. Under these assumptions, the demodulator using correlators is shown in Figure 3.9. Again, like in the coherent case, the optimality of the receiver has been rigorously proved (Appendix B). However, we can easily understand its operation by some heuristic argument as follows. The received signal (ignoring noise for the moment) with an unknown phase can be written as

$$\begin{aligned} s_i(t,\theta) &= A\cos(2\pi f_i t + \theta), \qquad i = 1,2 \\ &= A\cos\theta\cos 2\pi f_i t - A\sin\theta\sin 2\pi f_i t \end{aligned}$$

The signal consists of an in phase component A cos 8 cos 27r f t and a quadrature component A sin 8 sin 2x f,t sin 0. Thus the signal is partially correlated with cos 2s fit and partiah'y correlated with sin 27r fit. Therefore we use two correlators to collect the signal energy in these two parts. The outputs of the in phase and quadrature correlators will be cos 19 and sin 8, respectively. Depending on the value of the unknown phase 8, these two outputs could be anything in (- 5, y). Fortunately the squared sum of these two signals is not dependent on the unknown phase. That is

$$(\frac{AT}{2}\cos\theta)^2 + (\frac{AT}{2}\sin\theta)^2 = \frac{A^2T^2}{2}$$

This quantity is actually the mean value of the statistics I? when signal si (t) is transmitted and noise is taken into consideration. When si (t) is not transmitted the mean value of 1: is 0. The comparator decides which signal is sent by checking these I?. The matched filter equivalence to Figure 3.9 is shown in Figure 3.10 which has the same error performance. For implementation simplicity we can replace the matched filters by bandpass filters centered at f and fi, respectively (Figure 3.1 1).
However, if the bandpass filters are not matched to the FSK signals, degradation to



various extents will result. The bit error probability can be derived using the correlator demodulator (Appendix B). Here we further assume that the FSK signals are orthogonal, then from Appendix B the error probability is

$$P_b = \frac{1}{2}e^{-E_b/2N_o}$$