Unit - 1

Introduction

* Wireless Networking

Wireless Networks are computer networks that are connected by cables of any Kind, that uses wireless data connections between network nodes.

* History of wireless networking

The history of wireless networking goes as far back as the 1800's with the advent of radio waves. The spread of more technology grew throughout the years and expanded to what we communicated with today.

In 1888 a Hamburg, Germany born physicist named Heinrich Rudolf Herz produced the first radio wave ever. By 1894 this radio wave production became a way of communication. Telegraph wires were used to receive the radio waves in signal from. Herz opened the way for radio, television and radar with his discovery of electromagnetic waves.

In 1901, Marconi sent telegraphia signal over a long distance across the Altantic ocean which represents the starting wireless communication. In early years the way of signal modulation was amplitude modulation.

In 1935 Armstrong demonstrated the possibility of frequency modulation (FM), then FM became the main modulation technique.

During second world war radar signal was used for survellience. In late 1940's and 1950's commercial use of one way or two way radio and television was used.

In 1970's the cellular radio and personal communication began to accelerate representing the beginning of first generation (1G) system, successively this system is upgraded in bandwidth, data rate and other facility representing 2G, 3G and 4G.

* Challenges in wireless communication

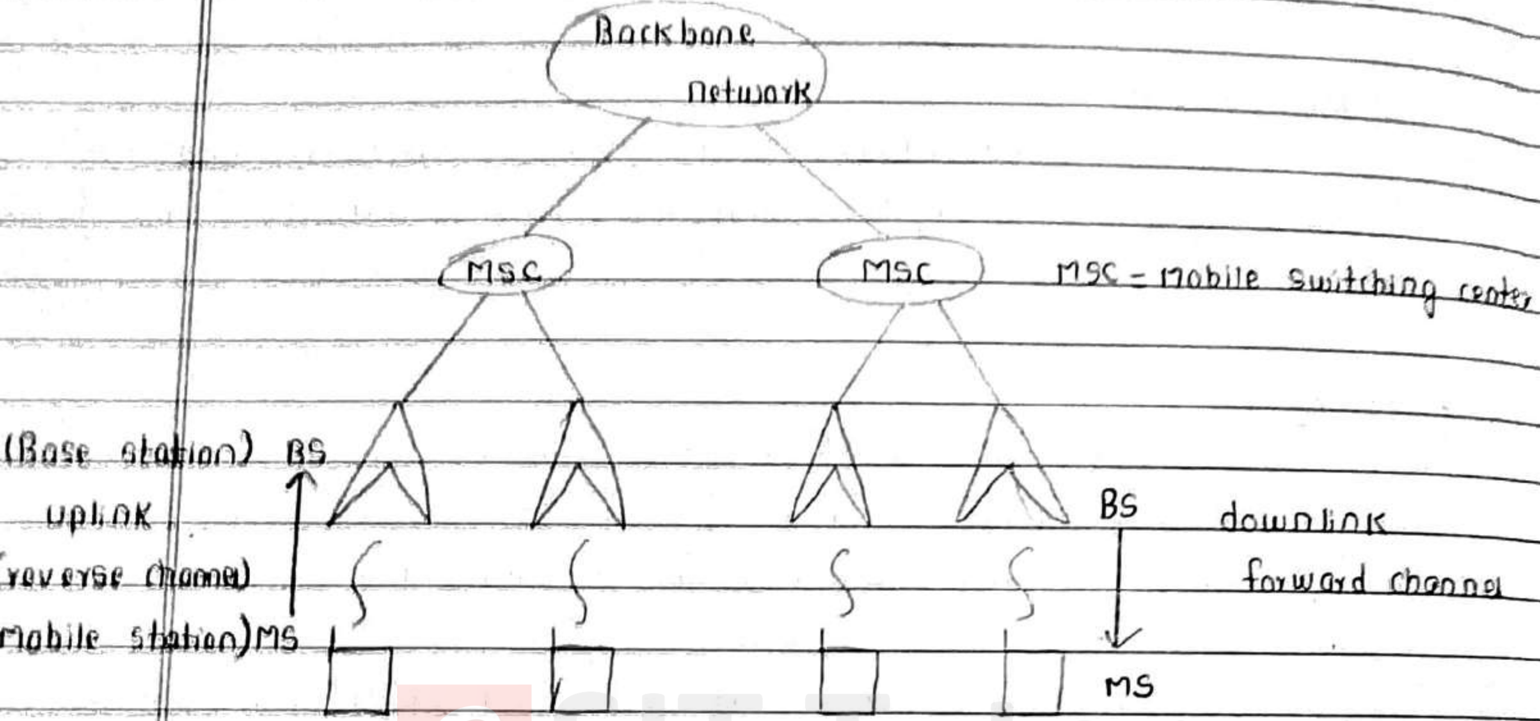

Fig:- wireless communication network

The challenges faced in wireless network are,

(i) Multipath Propagation

For wireless communications, the transmission medium is the radio channel between transmitter Tx and receiver Rx. The signal can get from the Tx to Rx via a number of different propagation paths. In some cases, a line of sight (LOS) connection might exist between Tx and Rx or the signal can get from the Tx ~~to~~ to Rx by being reflected at or diffracted by different interacting objects (IO) in the environment. The number of these possible propagation paths is very large. Each of the paths have a distinct amplitude, delay (runtime of signal), direction of departure from Tx and direction of arrival; most importantly, the components have different phase shifts with respect to each other.

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(ii) Spectrum limitations

The spectrum available for wireless communications services is limited and regulated by international agreements. For this reason, the spectrum has to used in a highly efficient manner. Two approaches are used: regulated spectrum usage where a single network operator has control over the usage of the spectrum and unregulated spectrum where each user can transmit without additional control as long as he complies with certain restrictions on the emission power and bandwidth.

(iii) Limited energy

Truly wireless communications requires not only that the information is sent over the air (not via cables) but also that the MS is powered by one-way or rechargeable batteries. Otherwise, a MS would be tied to the "wire" of the power supply, batteries in turn impose restrictions on the power consumption of the devices. The requirement for small energy consumption results in several technical imperatives:

↳ The power amplifiers in the transmitter have high efficiency.

↳ Signal processing must be done in an energy-saving manner.

↳ Maximum transmission power should be used only when required.

↳ For cellular phones and even more so for sensor networks, an energy efficient "standby" or "sleep" mode has to be defined.

(iv) Noise limited system

Wireless systems are required to provide a certain minimum transmission quality. This transmission quality in turn requires a minimum signal-to-noise ratio (SNR) at receiver RX. Consider now a situation where only a single BS transmits and a mobile station (MS) receives; thus the performance of the system is determined only by the useful signal and the noise. As the MS moves further away from the BS, the received signal power decreases, and at a certain distance, the SNR does not achieve the required threshold for reliable communications. Therefore, the range of the system is noise limited; equivalently, we can call it signal power limited. Depending on the interpretation, it is ~~one~~ too much noise or too little signal power that leads to bad link quality.

(v) Interference limited systems

Consider now case that the interference is so strong that it completely dominates the performance, so that the noise can be neglected. Let BS cover an area/cell that is approximately described by a circle with radius R and center at location of BS. Furthermore there is an interfering Tx at distance D from the "desired" BS, which operates at the same frequency and with the same transmit power. How large does D have to be in order to guarantee satisfactory transmission quality 90% of time, assuming that the MS is at the cell boundary (worst case)? The computations follow the link budget computations of the previous section. As a first approximation, we treat the inference as Gaussian. This allow us to treat the interference as equivalent noise and minimum SIR, takes on the same values as SNRmin the noise-limited case.

(vi) Wireless channel

The wireless channel exibibits many different forms of channel limitations such as delay, interference, distortion and noises. The interference and distortion that are multiplicative in nature along with fading causes attenuation and multilatron. The additive noise is another problem due to wireless channel. It reduces the signal intensity. The "out-of-band" noise can be suppressed by filtering but the "in-band" noise will still penetrate along the signal.

(vii) User mobility

Mobility is an inherent feature of most wireless systems has its important consequences for system design. With the limited radio spectrum, wireless systems are designed based on the cellular concept for frequency reuse to provide large number of user with communication services. A second effect is particularly to mobile users in cellular systems: the system has to know at any time which cell a user is in:

↳ If there is an incoming call for a certain MS (user), the network has to know in which cell the user is located.

↳ If an MS moves across a cell boundary, a different BS becomes the serving BS, in order words

The MS is handed over from one BS to another. Such a handover has to be performed without interrupting the call; as a matter of fact, it should not be noticeable at all to the user.

The communication between transmitter and receiver can be viewed as taking place in 3 layes

(a) Physical layer:- actual signal transmission and reception occurs here over the propagation channel.

(b) Link layer:- output from base station that is associated with power control, rate allocation and error control.

(c) Network layer:- This layer includes handoff managent, location management, traffic management and traffic control.

* Wireless communication standard

Wireless communications system that have been in deployment for sometime are those of the first and second generation. The third generation systems are also currently under deployment but continue to evolve.

	1G	2G	3G	4G
Era	1980	1990	2000	2010
Technology	Analog	Digital	Digital	Digital
Modulation	FM	BPSK, QPSK, GMSK	QPSK, QAM	OFDMA
Multiple Access	FDMA	FDMA, TDMA, CDMA	CDMA, TDMA	OFDMA
Duplexing	FDD	FDD/TDD	FDD/TDD	FDD/TDD
Switching	Circuit switching(CS)	Circuit switching	CS / PS (packet switching)	PS
Services	voice	voice + data	voice + video + data	voice, high speed data.
Roaming	Not provided	provided but only SMS	provided	provided.
Example	AMPS, ETACS, NTT	GSM, IS-54, IS-95	UMTS, CDMA 2000	LTE

* First Generation cellular system

The first generation cellular systems use analog FM for speech transmission. The individual calls use different frequencies and share the available spectrum through FDMA.

The radio interface technology of the first generation wireless cellular systems (AMPS in america, ETACS in Europe and NTT in Japan) is tabulated in table below,

Region	America	Europe	Japan
Parameter	AMPS	ETACS	NTT
Multiple Access	FDMA	FDMA	FDMA
Duplexing	FDD	FDD	FDD
Forward channel	869 - 894 MHz	935 - 960 MHz	870 - 885 MHz
Reverse channel	824 - 849 MHz	890 - 915 MHz	925 - 940 MHz
channel spacing	30 KHz	25 KHz	25 KHz
Data rate	10 Kbps (kilobits per second)	8 Kbps	0.3 Kbps
spectral efficiency	0.33 bps/Hz	0.33 bps/Hz	0.012 bps/Hz
capacity	832 channels	1000 channels	600 channels

* Second Generation cellular system

The second generation cellular systems are completely digital, employing either TDMA or CDMA as the multiple access technology. The digital technology allows greater sharing of the radio hardware in the base station among multiple users and provides a larger capacity to support more users per base station per MHertz of spectrum than analog systems. Digital systems offer a number of advantages over analog systems

(i) Natural integration with the evolving digital wireline network.

(ii) Flexibility for supporting multimedia services.

(iii) Flexibility for capacity expansion.

(iv) Reduction in RF (radio frequency) transmit power.

(v) Encryption for communication privacy

(vi) Reduction in system complexity.

The radio interface technology in this system is shown in table,

Region	US	Europe	Japan	US
Parameter	IS-54	GSM	PDC	IS-95
Multiple Access	TDMA	TDMA	TDMA	CDMA
Duplexing	FDD	FDD	FDD	FDD
Modulation	$\frac{\pi}{4}$ DQPSK	GMSK	$\frac{\pi}{4}$ DQPSK	QPSK / OQPSK
Forward channel	869-894 MHz	935-960 MHz	810-826 MHz	869-894 MHz
Reverse channel	824-849 MHz	890-915 MHz	940-956 MHz	824-849 MHz
Channel spacing	30 KHz	200 KHz	25 KHz	1250 KHz
Data/chip rate	48.6 Kbps	270.833 Kbps	42 kbps	1.2288 Mcps
Speech codec rate	7.95 Kbps	13.4 Kbps	6.7 Kbps	1.2/2.4/4.8/9.6 Kbps

* Third Generation cellular system

Third generation standardization activities were initiated in Europe and in North America under respective names IMT-2000 and CDMA-2000. IMT-2000 is wideband direct sequence code division multiple access (DS-CDMA) while CDMA-2000 is multi carrier code division multiple access (MC-CDMA). These both uses Frequency division duplex (FDD) to support two-way transmissions with frequency isolation.

It is likely that the third generation cellular systems will be equipped with infrastructure to support Personal Communication systems (PCS). The network infrastucture support will include

(a) Public land mobile networks (PLMNs)
(b) Mobile Internet Protocol (Mobile IP)
(c) wireless asynchronous transfer mode (WATM) networks
(d) Low earth orbit (LEO) satellite networks.

- Impact of high transmission rate

↳ 3G ~~are~~ have higher transmission rate and support of multimedia service.

↳ High transmission rate means the bandwidth of signal will be large compared with coherence bandwidth of propagation channel.

↳ When signal bandwidth is large compared with coherence bandwidth of channel, different frequency components of signal will experience different fading characteristics.

↳ A basic approach to combat frequency selective fading is to partition the signal into contiguous frequency bands, each of which is narrow compared with coherence bandwidth of channel.

↳ Each of signal components is then modulated onto different subcarrier and the signal components are sent over the channel in parallel so each of signal components will experience non-frequency-selective-fading

- System capacity and impact of user mobility

↳ To support higher transmission rate in 3G, the limited bandwidth of the system needs to be reused more often.

↳ Decreasing cell size allows for a higher degree of frequency reuse to increase system capacity.

↳ once handoff is complete, the mobile needs to identify its current location within the cellular array so that messages can be delivered to it in its new location.

↳ As a result, a reduction in cell size translates to large overhead for mobility management in the network.

unit-2
Wireless channel characterization

* Multipath propagation environment

Multipath is the propagation phenomenon that results in radio signals reacting the receiving antenna by two or more paths. causes of multipath include atmospheric ducting, ionospheric reflection and refraction and reflection from water bodies and terrestial objects such as mountains and buildings.

The wireless propagation channel contains objects (particles) which randomly scatter the energy of the transmitted signal. The scattered signals arrive at the destination receiver out of step. These objects/particles are referred to as scatters. Scattering by randomly located scatteres gives rise to different paths with different path lengths, resulting in multipath delay spread. A multipath situation arises when a transmitted point source is received as a multipoint source, with each of the individually received points experiencing a different transmission delay. The following figure is a view of scattering phenomenon.

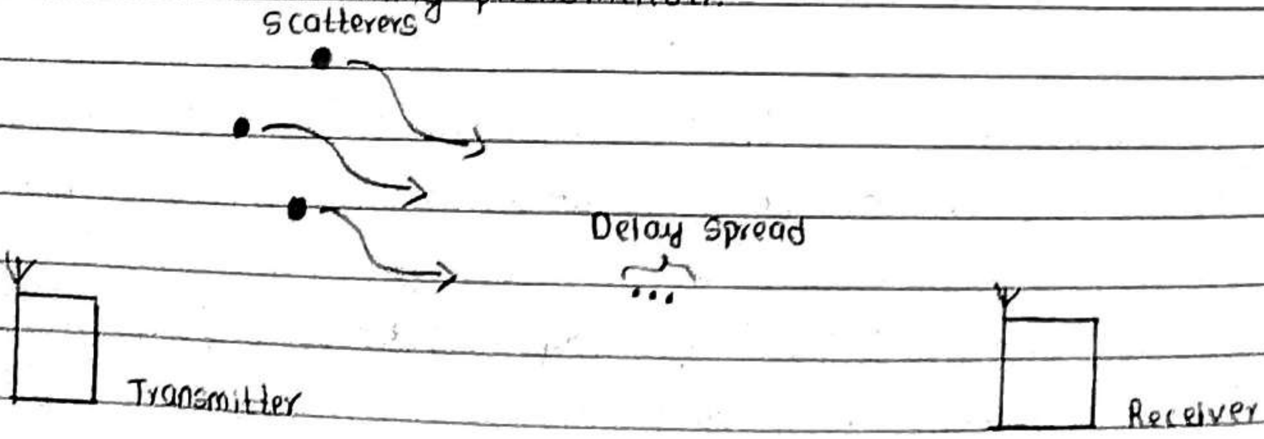


Fig:- Multipath spread due to channel scattering.

Multipath causes multipath interference including constructive & distructive interference and phase shifting of signal. Destructive interference causes fading where magnitudes of signals arriving by various paths have a distribution known as Rayleigh distribution / Rayleigh fading. The effect of multipath propagation can be characterized by time dispersion and fading.

* Time dispersion

When a signal transmitted, this signal can suffer a distortion caused by reflections and scattered propagation paths in the radio channel and these phenomenons cause that an identical signal arrives at different times at its destination. These different times are due to the signal arrives via multipaths and in different incident angles. The time difference between the arrival moment of the first multipath component and the last one is called delay spread.

As multipath propagation paths have different propagation delays, the transmitted point source signal is received as a smeared / widened waveform. Nonoverlapping scatterers give rise to distinct multiple paths, which are characterized by their locations in the scattering medium. The following figure is of ellipsoidal portrayal of scatterer location where scatterers are located on ellipses with transmitter (Tx) and receiver (Rx) as the foci. One ellipse is associated with one path length. Therefore, signals reflected by scatteres located on the same ellipse will experience the same propagation delay. The signal components from these multiple paths are indistinguishable at the receiver. Signals that are reflected by scatterers located on different ellipses will arrive at the receiver with differential delays.

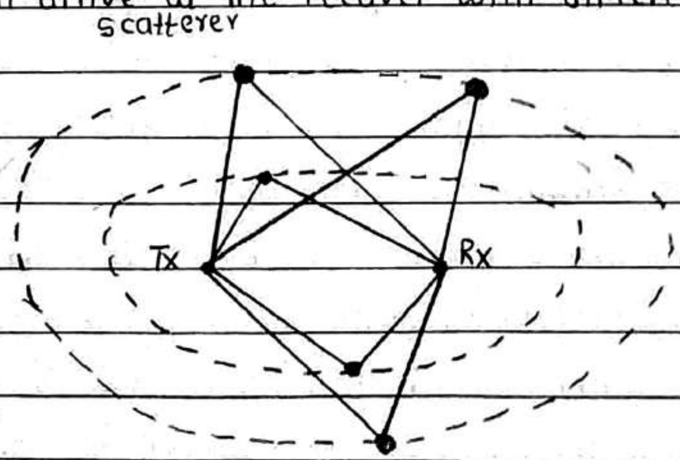


Fig. - Ellipsoidal portrayal of scatterer location.

↳ If the maximum differential delay spread is small compared with the symbol duration of the transmitted signal, the channel is said to exhibit flat fading.

↳ If the differential delay spread is large compared with the symbol interval, the channel exhibit frequency-selective fading.

↳ In the time domain, the received signals corresponding to successive transmitted symbols will overlap, giving rise to a phenomenon known as intersymbol interference (ISI)

↳ The severity of ISI increases with the width of delay spread. ISI degrades transmission performance and channel equalization techniques can be used to combat ISI.

* Fading

Fading is the deviation of the attenuation that a carrier-modulated telecommunication signal experiences over certain propagation.

when the delay differences among various distinct propagation paths are very small compared with the symbol interval in digital transmittion, the multipath components are almost indistinguishable at the receiver. These multipath components can add constructively or destructively, depending on the carrier frequency and delay differences.

As the mobile station moves, the position of each scatterer with respect to the transmitter and receiver may change. The overall effect is that the received signal level fluctuates with time, this phenomenon is called fading.

consider the transmission of a single-tone sinusoidal signal with frequency f_c over a channel with two distinct paths as in figure. Among two distinct paths, one path without the delay of los(Line-of-Sight) whose directed path is assumed to be zero and another path with the delay of non-line-of-sight(NLOS) having reflected path 'I'.

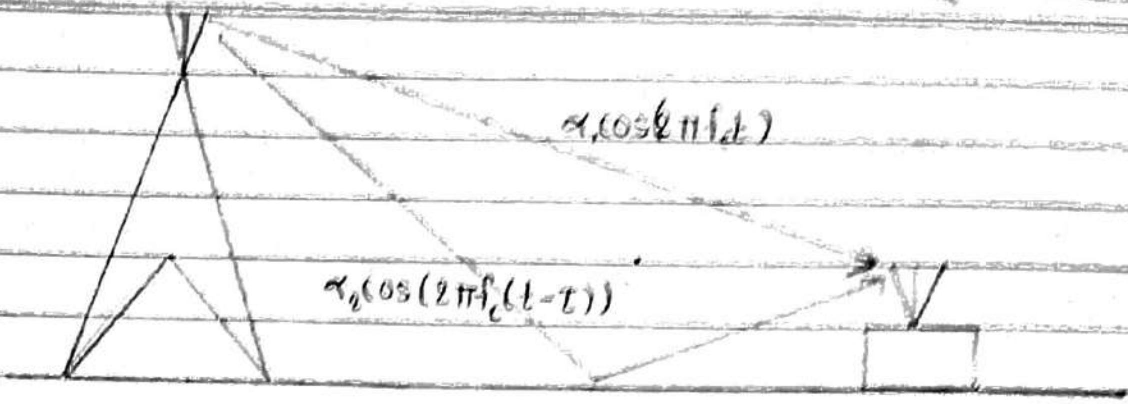


Fig:- channel with two propagation paths

The received signal, in the absence of noise can be represented as

$$r(t) = \alpha_1 \cos(2\pi f_c t) + \alpha_2 \cos(2\pi f_c (t-\tau))$$

where α_1 and α_2 are the amplitudes of the signal components from the two paths respectively. The received signal can also be represented as

$$r(t) = \alpha \cos(2\pi f_c t + \phi)$$

where,

$$\alpha = \sqrt{\alpha_1^2 + \alpha_2^2 + 2\alpha_1 \alpha_2 \cos(2\pi f_c t)}$$

and

$$\phi = -\tan^{-1}\left[\frac{\alpha_2 \sin(2\pi f_c \tau)}{\alpha_1 + \alpha_2 \cos(2\pi f_c t)}\right]$$

where α and ϕ are amplitude and phase of received signal. Both α and ϕ are functions of α_1, α_2 and τ.

↳ The two received signal components add constructively when $f_c t = 0, 1, 2, \cdots$

↳ The two received signal components add destructively when $f_c \tau = 0.5, 1.5, \cdots$

↳ The mobile station moves α_1, α_2 and τ change with time. The received signal amplitude α and phase ϕ also change with time. When the signal components from the two paths add destructively, the transmitted signal experiences deep fading with a small value of amplitude α.

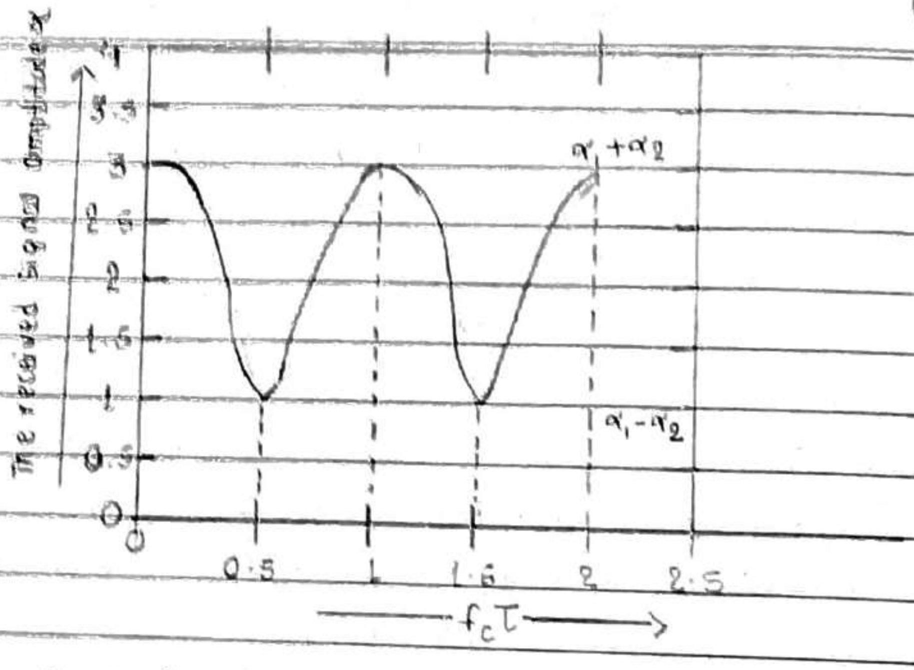


Fig :- Amplitude fluctuation of two path channel with $\alpha_1 = 2$ & $\alpha_2 = 1$

* Linear Time varient channel model

Consider a multipath propagation environment with N distinct scatters. The path associated with the n^{th} distinct scatterer is characterized by two tuples $(\alpha_n(t), \tau_n(t))$ where $\alpha_n(t)$ represents the amplitude function introduced to transmitted signal by a scatterer at time t, $\tau_n(t)$ is the associated propagation delay and $n = 1, 2, \cdots, N$.

Consider a narrowband signal $\tilde{x}(t)$ transmitted over the wireless channel at a carrier frequency f_c such that

$$\tilde{x}(t) = R\{x(t)e^{j2\pi f_c t}\}$$

where $x(t)$ is complex envelope of the signal and R denotes the real valued component. In the absence of background noise, the received signal at channel output is

$$\tilde{y}(t) = R\left\{\sum_{n=1}^{N} \alpha_n(t)\, x(t-\tau_n(t))\, e^{j2\pi f_c(t-\tau_n(t))}\right\}$$

$$= R\{y(t)e^{j2\pi f_c t}\}$$

where $y(t)$ is complex envelope of the received signal and it can be represented as

$$r(t) = \sum_{n=1}^{N} \alpha_n(t) e^{-j2\pi f_c \tau_n(t)} x(t - \tau_n(t))$$

The complex envelopes, $x(t)$ and $r(t)$ are respectively the equivalent representations of the transmitted and received narrowband signals at baseband. As the mobile user moves $\alpha_n(t)$ and $\tau_n(t)$ are a function of t, the channel is linear time-variant. The channel impulse response depends on the instant that the impulse is applied to the channel.

* channel Impulse Response

It is defined as the output of a channel or any other I/O system when the input of it is impulse. Impulse input is the very short time triggering input whose amplitude is very high let $h(t)$ denote the impulse response i.e. the channel output when the channel input is an impulse applied at $t=0$, $\delta(t)$. Here $\delta(t)$ is Dirac delta function and is defined by

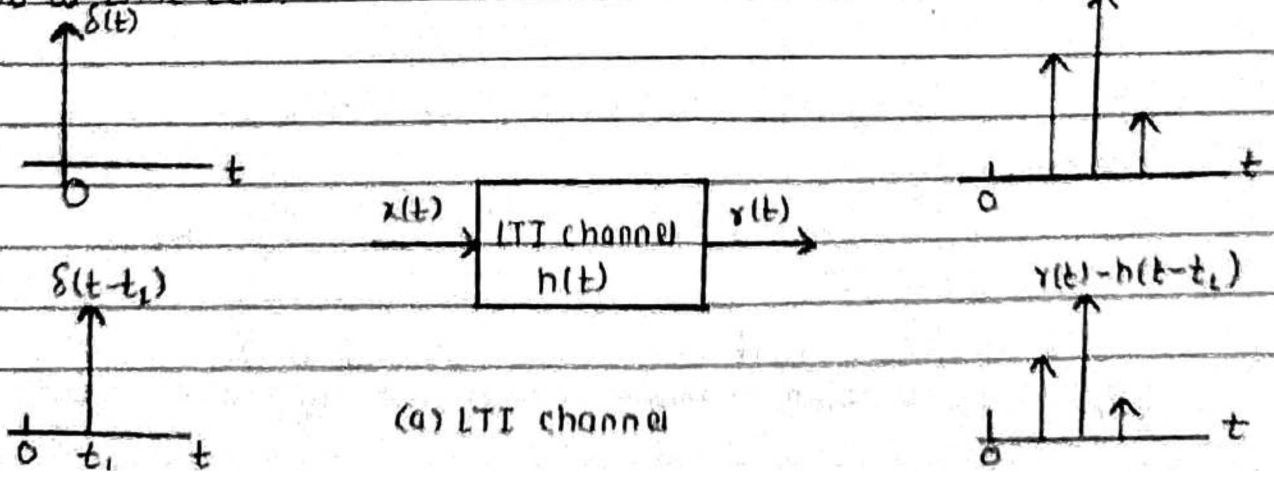
$$\delta(t) = \begin{cases} 0 & t \neq 0 \\ \infty & t = 0 \end{cases}$$

and,

$$\int_{-\infty}^{\infty} \delta(t) = 1$$

- Linear time-invariant (LTI) channel

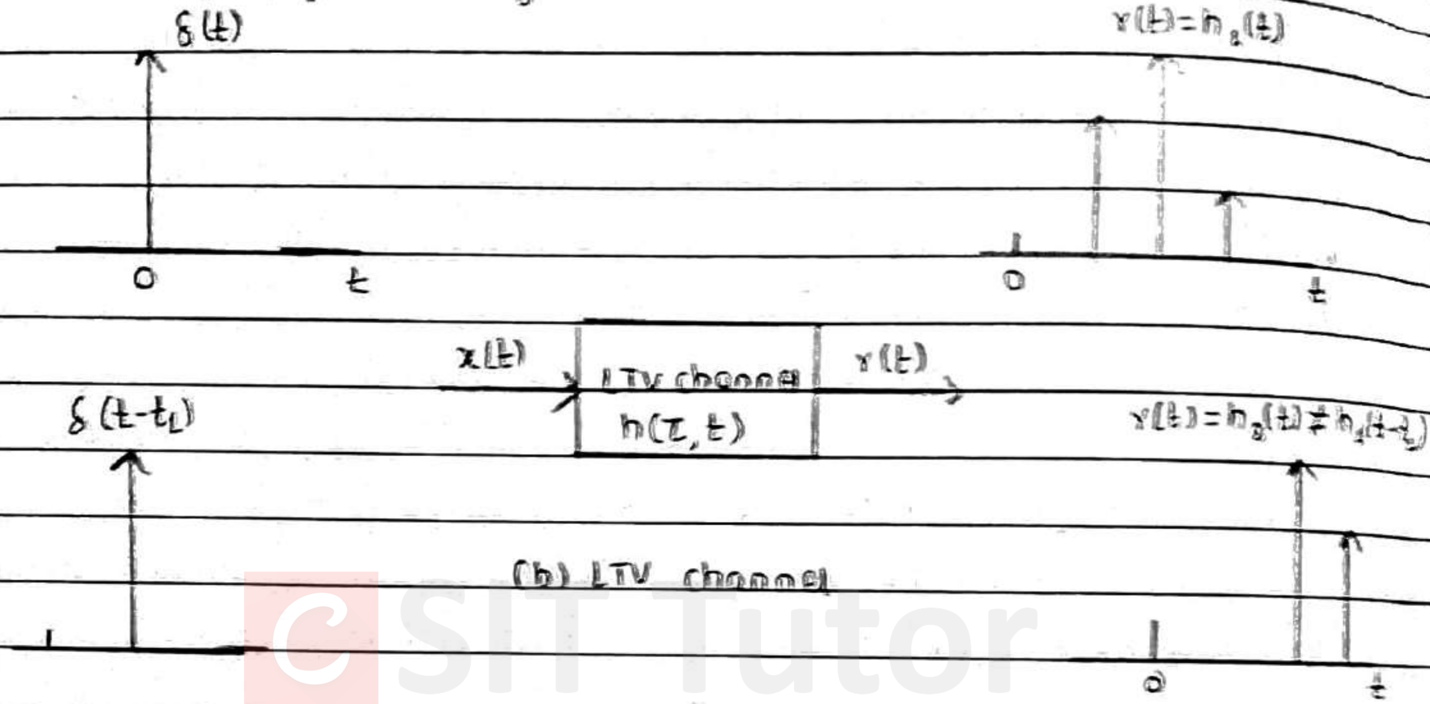
We can use $h(t)$ to describe the channel, where t is a variable describing propagation delay, assuming that impulse is always applied to the channel at time zero.



(a) LTI channel

- linear time varient (LTV) channel

The channel impulse response is a function of two variables: one describing when the impulse is applied to the channel, the other describing the moment of observing the channel output of the associated propagation delay.



(b) LTV channel

The impulse response of an LTV channel $h(\tau, t)$ is the channel output at t in response to an impulse applied to channel at $t-\tau$

The received signal is

$$r(t) = \int_{-\infty}^{\infty} h(\tau, t)\, x(t-\tau)\, d\tau$$

The channel impulse response for the channel with N distinct scatters is $h(\tau, t) = \sum_{n=1}^{N} \alpha_n(t)\, e^{-j\theta_n(t)}\, \delta(\tau - \tau_n(t))$

where $\theta_n(t) = 2\pi f_c \tau_n(t)$ represent the carrier phase distortion introduced by n^{th} scatterer

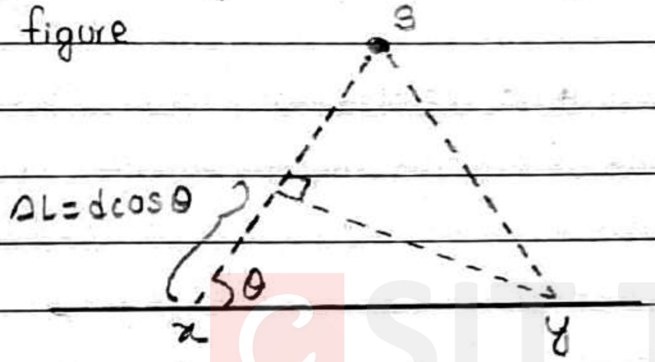
NOTE: $\tau_n(t)$ changes by $1/f_c \rightarrow \theta_n(t)$ changes by 2π

A small change in propagation delay -> a small changes in $\alpha_n(t)$ and $\tau_n(t)$ but a significant change in $\theta_n(t)$. That is the carrier phase distortion $\theta_n(t)$ is much more sensitive to user mobility than the amplitude distortion $\alpha_n(t)$.

* Doppler shift

It is also called doppler effect and it is the change in frequency or wavelength of a wave in relation to observer who is moving relative to the wave source. An example of doppler shift is the change of pitch heard when a vechicle surrounding a horn approaches and receds from an observer.

Consider a base station (BS) transmitting a singletone pilot signal at frequency 'f_c' and the mobile station (MS) is moving along a x-axis with a constant velocity 'v' and path segment between two points x and y have length 'd' and the signals are received from a remote source 'S' as in figure



The difference in path length travelled by the wave from the Source 'S' to the mobile at points x and y is

$$\Delta L = d\cos\theta$$
$$= v\Delta t\cos\theta$$

where, Δt is the time required for the mobile to travel from x to y. θ is assumed to be same at points x and y since the source is assumed to be very ~~big~~ far away.

Now,

From figure the path difference for signal upto point x and y is assumed to be equal to $\Delta L = d\cos\theta$ so, phase difference is

$$\Delta\phi = \frac{2\pi(\Delta L)}{\lambda} = \frac{2\pi d\cos\theta}{\lambda} = \frac{2\pi v\Delta t\cos\theta}{\lambda}$$

Now,

The apparent change in frequency (or doppler shift) is given by

$$f_d = \frac{1}{2\pi}\frac{\Delta\phi}{\Delta t}$$

$$= \frac{1}{2\pi}\frac{2\pi v \Delta t \cos\theta}{\lambda \Delta t}$$

$$= \frac{v}{\lambda}\cos\theta \quad \text{where, } \lambda = \frac{c}{f} = \frac{\text{speed of light}}{\text{carrier frequency}}$$

Total shift in frequency is thus

$$f = f_c \pm f_d.$$

The above equation relates the doppler shift to the mobile velocity. so we can say that if the mobile is moving towards the direction of source, the shift is positive and vice versa.

* channel correlation function

Assumptions

(a) The channel impulse response $h(\tau, t)$ is a wide-sense stationary (WSS) process.

(b) The channel impulse responses at τ_1 and τ_2, $h(\tau_1, t)$ and $h(\tau_2, t)$ are uncorrelated if $\tau_1 \neq \tau_2$ for any t.

A channel under assumption (a) and (b) is said to be a wide-sense stationary uncorrelated scattering (WSSUS) channel.

• Delay power spectral density (psd)

under assumption (a) the autocorrelation function of $h(\tau, t)$ is $\phi_h(\tau_1, t_2, \Delta t) \triangleq \frac{1}{2} E[h^*(\tau_1, t)\, h(\tau_2, t+\Delta t)]$ ------ (1)

under assumption (b), function can be represented as

$$\phi_h(\tau_1, \tau_2, \Delta t) = \phi_h(\tau_1, \Delta t)\,\delta(\tau_1 - \tau_2) \quad \text{--------- (2)}$$

τ_1, τ_2 & Δt doesnot depend on t and subscript (*) denotes complex conjugation.

Equivalently,

$$\phi_h(\tau, \tau+\Delta\tau, \Delta t) = \phi_h(\tau, \Delta t)\, \delta(\Delta\tau) \quad \text{-----}(3)$$

where,

$$\phi_h(\tau, \Delta t) = \int \phi_h(\tau, \tau+\Delta\tau, \Delta t)\, d\Delta\tau$$

At $\Delta t = 0$,

$$\phi_h(\tau) \triangleq \phi_h(\tau, 0) \quad \text{-----------}(4)$$

From eqns (3) & (4)

$$\phi_h(\tau) = F_{\Delta\tau}\left[\phi_h(\tau, \tau+\Delta\tau, \Delta t)\right]\Big|_{\Delta t=0}$$

$$= F_{\Delta\tau}\left\{\frac{1}{2} E\left[h^*(\tau, t)\, h(\tau+\Delta\tau, t)\right]\right\}$$

$\phi_h(\tau)$ measures the average psd at the channel output as a function of the propagation delay, τ and is therefore called the delay psd of the channel, also known as the multipath intensity profile. The nominal width of the delay psd pulse is called the multipath delay spread denoted by T_m.

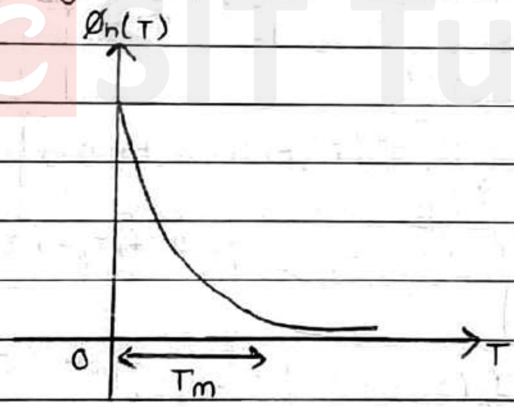


Fig:- Delay power spectral density

The nth moment of the delays

$$\overline{\tau^n} = \frac{\int \tau^n \phi_h(\tau)\, d\tau}{\int \phi_h(\tau)\, d\tau}$$

The mean propagation delay or first moment denoted by $\bar{\tau}$ is

$$\bar{\tau} = \frac{\int \tau\, \phi_h(\tau)\, d\tau}{\int \phi_h(\tau)\, d\tau}$$

and the rms (root mean square) delay spread, denoted by σ_τ is

$$\sigma_\tau = \left[\frac{\int (\tau-\bar{\tau})^2 \phi_h(\tau)\, d\tau}{\int \phi_h(\tau)\, d\tau}\right]^{1/2}$$

In calculating a value for the multipath delay spread, it is usually assumed that

$$T_m \approx \sigma_\tau$$

- Frequency and time correlation functions

The autocorrelation function of $H(f,t)$ is

$$\phi_H(f_1, f_2, t, \Delta t) \triangleq \frac{1}{2} E\left[H^*(f_1,t)\, H(f_2, t+\Delta t)\right]$$

$$\overset{WSS}{\Rightarrow} \quad \phi_H(f_1, f_2, \Delta t) = \frac{1}{2} E\left|H^*(f_1,t)\, H(f_2, t+\Delta t)\right|$$

$$\overset{US}{\Rightarrow} \quad \phi_H(f_1, f_2, \Delta t) = \frac{1}{2} E\left\{\left[\int h(\tau_1, t)\, e^{-j2\pi f_1 \tau_1} d\tau_1\right]^* \left[\int h(\tau_2, t+\Delta t)\, e^{-j2\pi f_2 \tau_2} d\tau_2\right]\right\}$$

$$= \iint \frac{1}{2} E\left[h^*(\tau_1, t)\, h(\tau_2, t+\Delta t)\right] e^{-j2\pi(f_2\tau_2 - f_1\tau_1)} d\tau_1 d\tau_2$$

$$\overset{WSS}{=} \iint \phi_h(\tau_1, \tau_2, \Delta t)\, e^{-j2\pi(f_2\tau_2 - f_1\tau_1)} d\tau_1 d\tau_2$$

$$\overset{US}{=} \iint \phi_h(\tau_1, \Delta t)\, \delta(\tau_1 - \tau_2)\, e^{-j2\pi(f_2\tau_2 - f_1\tau_1)} d\tau_1 d\tau_2$$

$$= \int \phi_h(\tau_1, \Delta t)\, e^{-j2\pi(f_2 - f_1)\tau_1} d\tau_1$$

$$= \int \phi_h(\tau, \Delta t)\, e^{-j2\pi(\Delta f)\tau} d\tau \quad (\Delta f = f_2 - f_1)$$

$\triangleq \phi_H(\Delta f, \Delta t)$ – time-frequency correlation function

Frequency correlation function

Let $\Delta t = 0$ we have

$$\phi_H(\Delta f) \triangleq \frac{1}{2} E\left[H^*(f,t) H(f+\Delta f, t)\right]$$

$$= \int \phi_h(\tau) e^{-j2\pi\Delta f \tau} d\tau$$

↳ $\phi_h(\tau)$ and $\phi_H(\Delta f)$ is a pair of Fourier transform

↳ uncorrelated scattering $\Rightarrow$ WSS $H(f,t)$ w.r.t f

↳ $\phi_H(\Delta f)$ characterizes the correlation of channel gains at f and $f+\Delta f$ for any t.

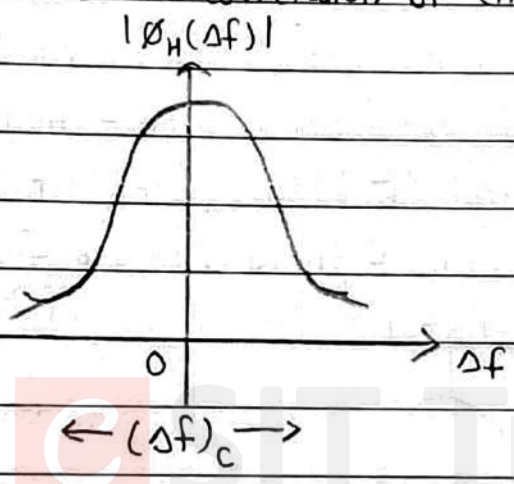


Fig:- frequency-correlation function and channel coherence bandwidth

↳ Coherence bandwidth

The maximum frequency difference for which the signals are still strongly correlated is called the coherence bandwidth of channel, denoted by $(\Delta f)_c$

Two sinusoids with frequency separation larger than $(\Delta f)_c$ are affected differently by the channel at any t.

Let W_s denote the bandwidth of transmitted signal.

↳ If $(\Delta f)_c < W_s$, the channel is said to exhibit frequency selective fading which introduces severe ISI to the received signal.

→ If $(\Delta f)_c >> W_s$, the channel is said to exhibit non selective fading or the fading which introduces negligible ISI.

$$\phi_h(t) \leftrightarrow \phi_H(\Delta f) \Rightarrow (\Delta f)_c \approx \frac{1}{T_m}$$

Time correlation function $\phi_H(\Delta t)$

Letting $\Delta f = 0$ in the time-frequency correlation function $\phi_H(\Delta f, \Delta t)$ we have,

$$\phi_H(\Delta t) \triangleq \phi_H(0, \Delta t) = \frac{1}{2} E\left[H^*(f,t)\, H(f, t+\Delta t)\right]$$

↳ $\phi_H(\Delta t)$ characterizes on average how fast the channel transfer function changes with time at each frequency.

↳ The nominal width of $\phi_H(\Delta t)$, $(\Delta t)_c$ is called coherence time of fading channel.

↳ If the channel coherence time is much larger than symbol interval of the transmitted signal, the channel exhibit slow fading.

↳ $\phi_H(\Delta t)$ is independent of f due to the US assumption ⇒ US in the time domain is equivalent to WSS in the frequency domain.

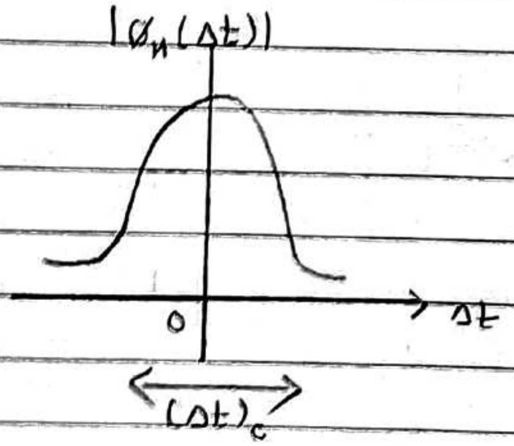


Fig:- Time correlation function and channel coherence time

* Large scale path loss and shadowing

Most cellular radio system operate in the highly populated area where there is no direct LOS path between transmitter and receiver and where the presence of large buildings due to which there is loss in signal.

↳ The interaction between reflection wave components due to obstacle cause multipath fading and the strength of wave decreases as the distance between T_x and R_x increases.

↳ In order to take into account for the losses and to predict the mean signal strength for certain $Tx - Rx$ separation, the propagation models are used and the corresponding models are called large scale propagation models.

Consider a flat channel with channel impulse response

$$h(\tau, t) \approx h(\bar{\tau}, t) \triangleq g(t)\,\delta(\tau - \bar{\tau})$$

The received signal is

$$r(t) = \int_{-\infty}^{\infty} h(\tau, t)\, x(t-\tau)\, d\tau \approx g(t)\, x(t-\bar{\tau})$$

$$x(t) \longrightarrow \boxed{\begin{array}{c}\text{Flat fading channel} \\ h(\tau, t) = g(t)\,\delta(\tau-\bar{\tau})\end{array}} \longrightarrow r(t) = g(t)\, x(t-\bar{\tau})$$

Fig :- Flat fading channel

* Free space propagation model

When the distance between the transmitting antenna and receiving antenna is much larger than the wavelength of the transmitted wave and the largest physical linear dimension of the antennas, the power P_r at the output of receiving antenna is given by

$$P_r = P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2$$

where,

P_t = Total power radiated by an isotropic source

G_t = transmitting antenna gain

G_r = receiving antenna gain

d = distance between transmitting and receiving antennas

λ = wavelength of carrier signal = $\frac{c}{f_c}$

c = speed of light (3×10^8 m/s)

f_c = carrier frequency

$P_t G_t \stackrel{\Delta}{=}$ effective isotropically radiated power (EIRP)

The term $(4\pi d/\lambda)^2$ is known as free-space path loss denoted by $L_p(d)$ which is

$$L_p(d) = \frac{\text{EIRP} \times \text{Receiving antenna gain}}{\text{Received power}}$$

$$= -10\log_{10}\left[\frac{\lambda^2}{(4\pi d)^2}\right] \text{ (dB)}$$

$$= -20\log_{10}\left(\frac{c/f_c}{4\pi d}\right) \text{ (dB)}$$

The path loss is

$$L_p(d) = 20\log_{10} f_c + 20\log_{10} d - 147.56 \text{ (dB)}$$

Note that the free space path loss increases by 6 dB for every doubling of the distance and also for every doubling of the radio frequency.

* Propagation over smooth plane

Free space propagation model does not accurately apply in a mobile radio environment and the propagation loss does not only depend upon the distance and wavelength but also on antenna heights.

A simple two path model (two ray model) is used to take into account of the effect of antenna heights.

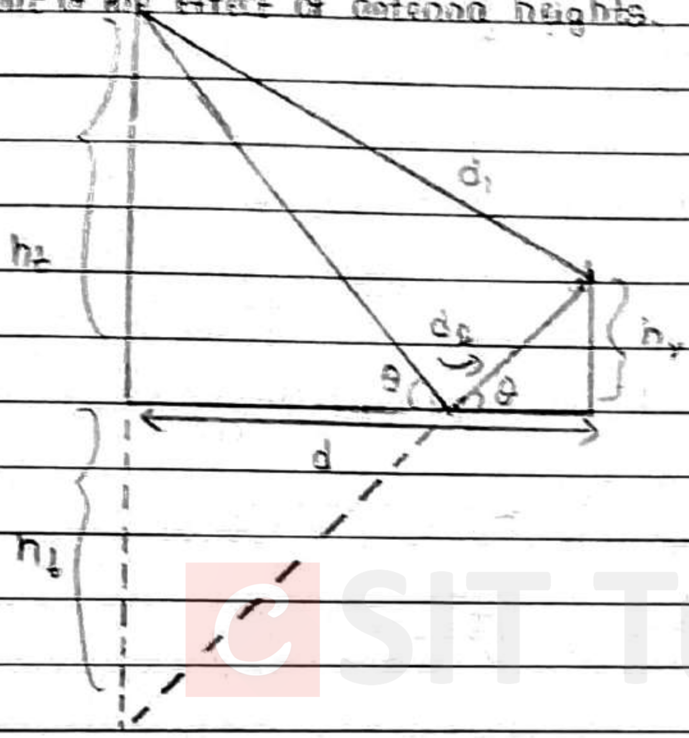


Fig:- Two path propagation over a Smooth plane

d_1 = direct LOS path

d_2 = propagation distance of reflected path

Consider a signal transmission over smooth, reflecting plane as in the following figure

From figure,

$$d_1 = \sqrt{d^2 + (h_t - h_r)^2}$$

where, h_t = height of transmitting antenna

h_r = height of receiving antenna

$$d_2 = \sqrt{d^2 + (h_t + h_r)^2}$$

In large scale propagation model as d is very large

$d >> h_t$ & h_r and $d_1 \approx d$, $d_2 \approx d$

Due to different delays for these two rays, there occurs slight change in carrier phases from the two paths and received signal component from two paths may enhance or cancel each other.

The carrier phase difference between the two propagation distances d_1 & d_2 is

$$\Delta\phi = (\phi_2 - \phi_1) = \frac{2\pi}{\lambda}(d_2 - d_1)$$

Taking this phase differences the received signal power is

$$P_r(d) = P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \left|1 + \alpha_f e^{-j\beta_f} e^{j(\phi_2 - \phi_1)}\right|^2$$

where,

α_f = amplitude attenuation due to reflection

β_f = carrier phase shift coefficient

It is considered that $\alpha_f \approx 1$ as reflection loss is negligible and β_f is assumed to be π for very large d.

So,

$$P_r(d) = P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \left|1 + e^{-j\pi} e^{j\frac{2\pi}{\lambda}\Delta d}\right|^2$$

$$= P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \left|1 - e^{j\frac{2\pi}{\lambda}\Delta d}\right|^2$$

$$= P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \left|1 - \cos\frac{2\pi}{\lambda}\Delta d - j\sin\frac{2\pi}{\lambda}\Delta d\right|^2 \quad [\because e^{j\theta} = \cos\theta + j\sin\theta]$$

$$= P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \left[\sqrt{\left(1 - \cos\frac{2\pi}{\lambda}\Delta d\right)^2 + \left(\sin\frac{2\pi}{\lambda}\Delta d\right)^2}\right]^2 \quad [\because |x - jy| = \sqrt{x^2+y^2}]$$

$$= P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \left[1 - 2\cos\frac{2\pi}{\lambda}\Delta d + \cos^2\frac{2\pi}{\lambda}\Delta d + \sin^2\frac{2\pi}{\lambda}\Delta d\right]$$

$$= P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \left[2 - 2\cos\frac{2\pi}{\lambda}\Delta d\right]$$

$$= P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \left[4\sin^2\frac{\pi\Delta d}{2\lambda}\right] \quad [1 - \cos 2\theta = 2\sin^2\theta]$$

Also,

$$\Delta d = d_2 - d_1 = \sqrt{d^2 + (h_t + h_r)^2} - \sqrt{d^2 + (h_t - h_r)^2}$$

$$= d \times \left[1 + \left(\frac{h_t + h_r}{d}\right)^2\right]^{1/2} - d \times \left[1 + \left(\frac{h_t - h_r}{d}\right)^2\right]^{1/2}$$

using Taylor series approximation upto two terms

$$\Delta d = d \left\{ \left[1 + \frac{1}{2}\left(\frac{h_t + h_r}{d}\right)^2\right] - \left[1 + \frac{1}{2}\left(\frac{h_t - h_r}{d}\right)^2\right] \right.$$

$$= d \left\{ 1 + \frac{1}{2}\left(\frac{h_t + h_r}{d}\right)^2 - 1 - \frac{1}{2}\left(\frac{h_t - h_r}{d}\right)^2 \right\}$$

$$= d \times \frac{1}{2d^2} \left\{ (h_t + h_r)^2 - (h_t - h_r)^2 \right\}$$

$$= \frac{1}{2d} \left[h_t^2 + 2h_t h_r + h_r^2 - h_t^2 + 2h_t h_r - h_r^2 \right]$$

$$= \frac{4 h_t h_r}{2d}$$

$$\therefore \Delta d = \frac{4 h_t h_r}{2d}$$

Substituting value of Δd in $P_r(d)$

$$P_r(d) = 4 P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \sin^2\left(\frac{2\pi h_t h_r}{d\lambda}\right)$$

$$\therefore L(d) = -4 \left(\frac{\lambda}{4\pi d}\right)^2 \sin^2\left(\frac{2\pi h_t h_r}{\lambda d}\right)$$

* okumura model
- frequency range (150-1920) MHz
- distance (1Km-100Km)
- BS antenna heights (30m-1000m)
- MS antenna height (1m - 3m)

* Hata model
- frequency range (150-1500) MHz
- BS antenna height (30m-200m)
- MS antenna height (1m-10m)

* okumura model

The okumura model is a radio propagation model that was built using the data collected in the city of Tokyo, Japan. The model is ideal for using in cities with many urban structures but not many tall block structures. The model served as a base for the Hata model.

okumura model was built into three modes urban, suburban and open areas.

The path loss in dB is

$$L(dB) = L_f(f,d) + A_{mu}(f,d) - G(h_{te}) - G(h_{re}) - G_{area}$$

where,

$L_f(f,d)$ = free space path loss in dB at distance 'd' and frequency 'f'

~~where~~, A_{mu} - median attenuation relative to free space

$G(h_{te})$ = Gain at transmitter

$G(h_{re})$ = Gain at receiver

G_{area} = Gain provided by type of area

$$G(h_{te}) = 20\log_{10}\frac{h_{te}}{200m}, \quad \text{for } 30m < h_{te} < 1000m$$

$$G(h_{re}) = \begin{cases} 10\log_{10}\dfrac{h_{re}}{3m}, & 0m < h_{re} < 3m \\ 20\log_{10}\dfrac{h_{re}}{3m}, & 3m < h_{re} < 10m \end{cases}$$

* Hata model

The Hata model is a radio propagation model for predicting the path loss of cellular transmissions in exterior environments. It is an empirical formulation based on the data from the okumura model and is also commonly called as "okumura-Hata model" the model incorporate the graphical information from okumura model and develops it further to realize the effects of diffraction, reflection and scattering caused by city structures. The Hata model applies corrections for applications in suburban and rural environments.

The path loss in dB is given by

$$L_p(d) = \begin{cases} A + B\log_{10}(d) & \text{for urban area} \\ A + B\log_{10}(d) - C & \text{for suburban area} \\ A + B\log_{10}(d) - D & \text{for open area} \end{cases}$$

where,

$$A = 69.55 + 26.16\log_{10}(f_c) - 13.82\log_{10}(h_{te}) - a(h_{re})$$

$$B = 44.9 - 6.55\log(h_{te})$$

$$C = 5.4 + 2\left(\log_{10}(fc/28)\right)^2$$

$$D = 40.94 + 4.78\left(\log_{10}(f_c)\right)^2 - 18.33\log_{10} f_c$$

Now,

$a(h_{re})$ is correction factor for mobile antenna and is given by

$$a(h_{re}) - (1.1\log_{10} f_c - 0.7)h_{re} - (1.56\log_{10} f_c - 0.8)$$

for small to medium city

$$a(h_{re}) = \begin{cases} 8.29\left(\log_{10}(1.54h_{re})\right)^2 - 1.1 & \text{for } f_c \leq 200 \text{ MHz} \\ 3.2\left(\log_{10}(11.75h_{re})\right)^2 - 4.97 & \text{for } f_c \geq 300 \text{ MHz} \end{cases}$$

* Small scale fading
- Based on multipath time delay spread
  Flat fading , Frequency selective fading

- Based on Doppler spread
  Fast fading , slow fading

For the measurement of the signal power and hence fluctuation for the movement of small distances we use small scale fading model.

This type of the small scale fading depends upon the different channel parameters like coherence time, doppler spread, etc

Fading due to delay spread

(i) Flat fading

If the mobile radio channel has a constant gain over a bandwidth which is greater than the bandwidth of transmitted signal than received signal will undergo flat fading.

BW of signal < BW of channel

Delay spread (σ_{τ}) < symbol period (T_s)

(ii) Frequency selective fading

If the channel posses a constant gain over a bandwidth that is smaller than the bandwidth of transmitted signal then the received signal undergo frequency selective fading

BW of signal > BW of channel

Delay spread (σ_{τ}) > signal period (T_s)

Fading due to doppler spread

(i) Fast fading

If the coherence time of the channel is smaller than the symbol period of the transmitted signal and leads to signal distortion then such a fading is fast fading.

↳ High Doppler spread

↳ coherence time (T_c) < symbol period (T_s)

(ii) slow fading

If the coherence time of the channel is greater than the symbol period of transmitted signal with less signal distortion then such fading is slow fading.

↳ less Doppler spread

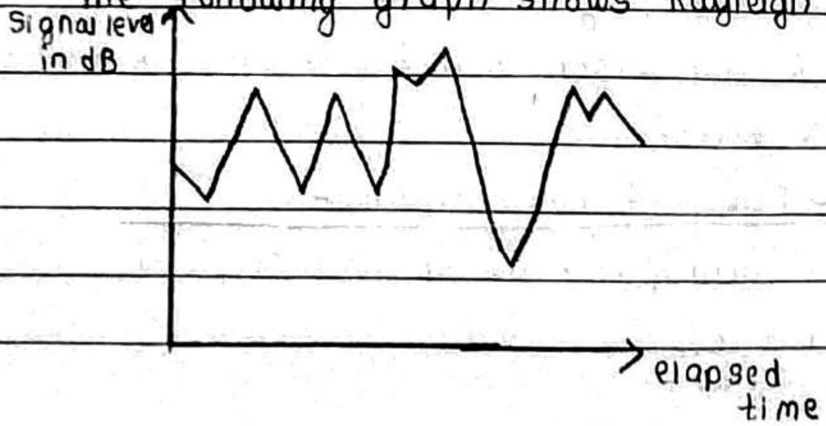
↳ coherence time (T_c) > symbol period (T_s)

* Rayleigh fading distribution

In mobile radio channels, the Rayleigh fading distribution is commonly used to describe the statistical time varying nature of the received envelop of a flat fading signal.

This Rayleigh fading model is used in that case when the every frequency components experience the extreme fading and there is no dominant high power components. so, it is used in case of NLOS communication scenario.

The following graph shows Rayleigh distributed signal envelope.



The Rayleigh fading distribution has pdf given as

$$P(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(\frac{-r^2}{2\sigma^2}\right) & ; 0 \le r \le \infty \\ 0 & ; r < 0 \end{cases}$$

where, σ is amplitude value of received signal.
σ^2 is time average power of received signal.

* Ricean fading distribution

When there is a dominant stationary (non-fading) signal component present, the small scale fading envelope follows Ricean distribution.

In this situation, the random multipath components arriving at different angles may superimpose on the stationary dominant signal.

As the dominant signal gets weaker due to interference of other low power components, the Ricean distribution approximates Rayleigh distribution.

The Ricean distribution is given by

$$P(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2 + A^2}{2\sigma^2}\right) I_0\left(\frac{Ar}{\sigma^2}\right) & ; \text{for } A \ge 0, r \ge 0 \\ 0 & ; r < 0 \end{cases}$$

where, A = peak amplitude of dominant signal
$I_0(\,)$ = modified Bessel function,
where, $I_0(x) = \frac{1}{2\pi}\int_0^{2\pi} e^{x\cos\theta}\, d\theta$

The Ricean distribution fading channel, has an important parameter called 'K' factor.

$$K \simeq \frac{\text{Power of LOS component}}{\text{Total power of all other components.}}$$

$$\text{or, } K = \frac{A^2}{2\sigma^2}$$

As 'K' approaches 0, the ricean distribution approaches the Rayleigh distribution on the other hand as 'K' approaches ∞, only the dominant component matters and it can be said that there is no fading.

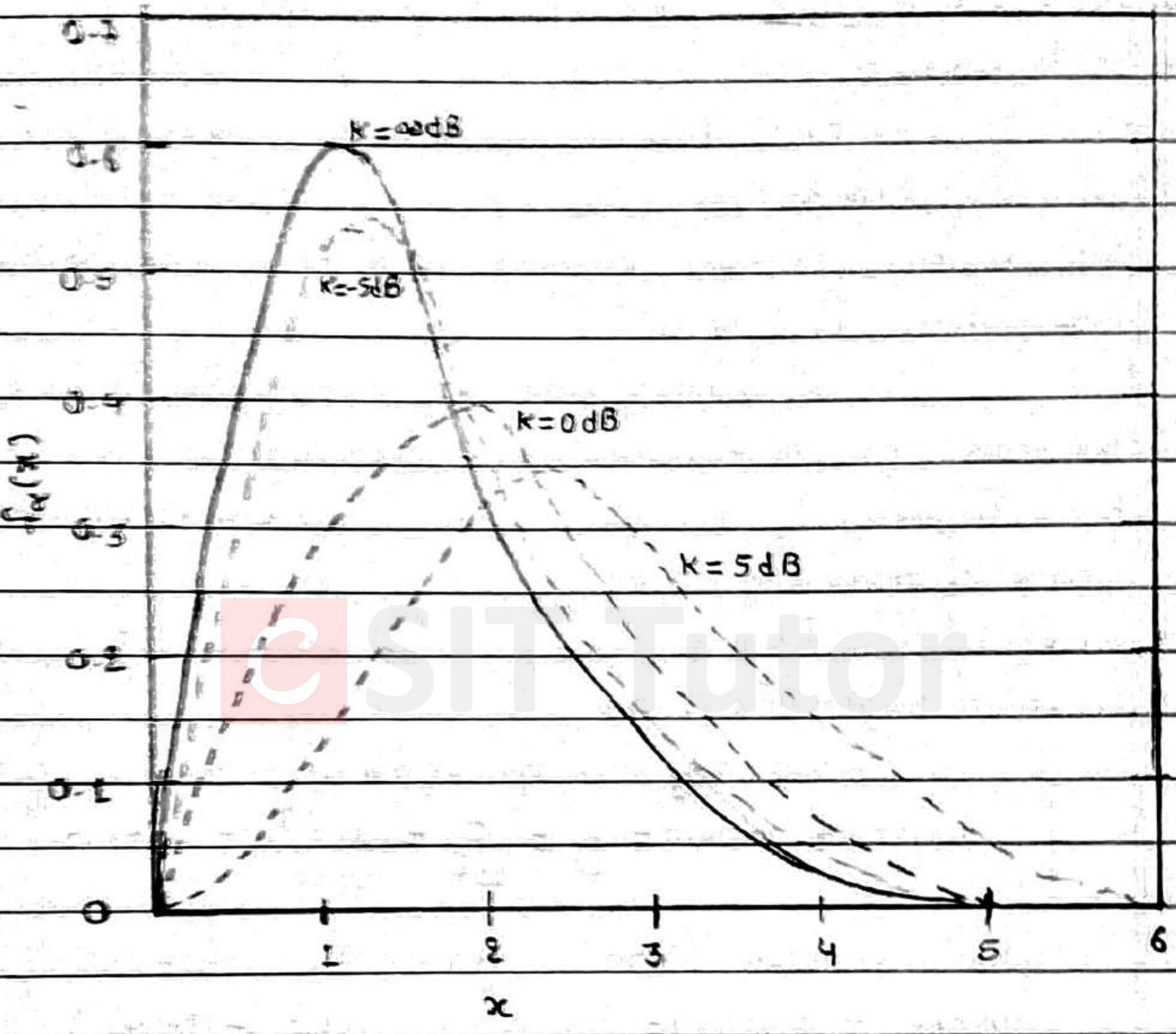


Fig:- Rayleigh and Rician fading distribution with $\sigma_z = 1$.

unit - 3

Bandpass Transmission Techniques

* Introduction
* Modulation

Modulation is the process of varying one or more properties of a periodic waveform called the carrier signal with a modulating signal that typically contains information to be transmitted. A modulator is a device that performs modulation.

It involves translating a baseband message signal to a bandpass signal at frequencies that are very high when compared to baseband frequency. The bandpass signal is modulated signal and the baseband signal is modulating signal.

- Necessity of modulation

(i) To reduce height of antenna

when only message signal is transmitted $m(t) = 5\,KHz$

$$\text{Height of antenna} = \frac{\lambda}{10} = \frac{c}{10f} = \frac{3\times10^8}{10\times5\times10^3} = 6000\,m$$

when modulated signal is transmitted $m(t) = 10\,MHz$

$$\text{Height of antenna} = \frac{\lambda}{10} = \frac{c}{10f} = \frac{3\times10^8}{10\times10^6} = 30\,m$$

During modulation high frequency carrier signal is used which results in drastic reaction of antenna height as seen in example.

(ii) Helps in reducing inferences and crosstalk

(iii) Long distance transmission

(iv) Multiplexing

* Demodulation

Demodulation is extracting the original information bearing signal from a carrier wave.

Here, the baseband message signal is extracted from the modulated bandpass signal so that it may be processed and interpreted by the receiver. The main goal of any modulation technique is to transport message signal through a radio channel with best possible quality while occupying least amount of bandwidth.

* Elements of a digital communication system

The following block diagram shows the elements of digital communication system.

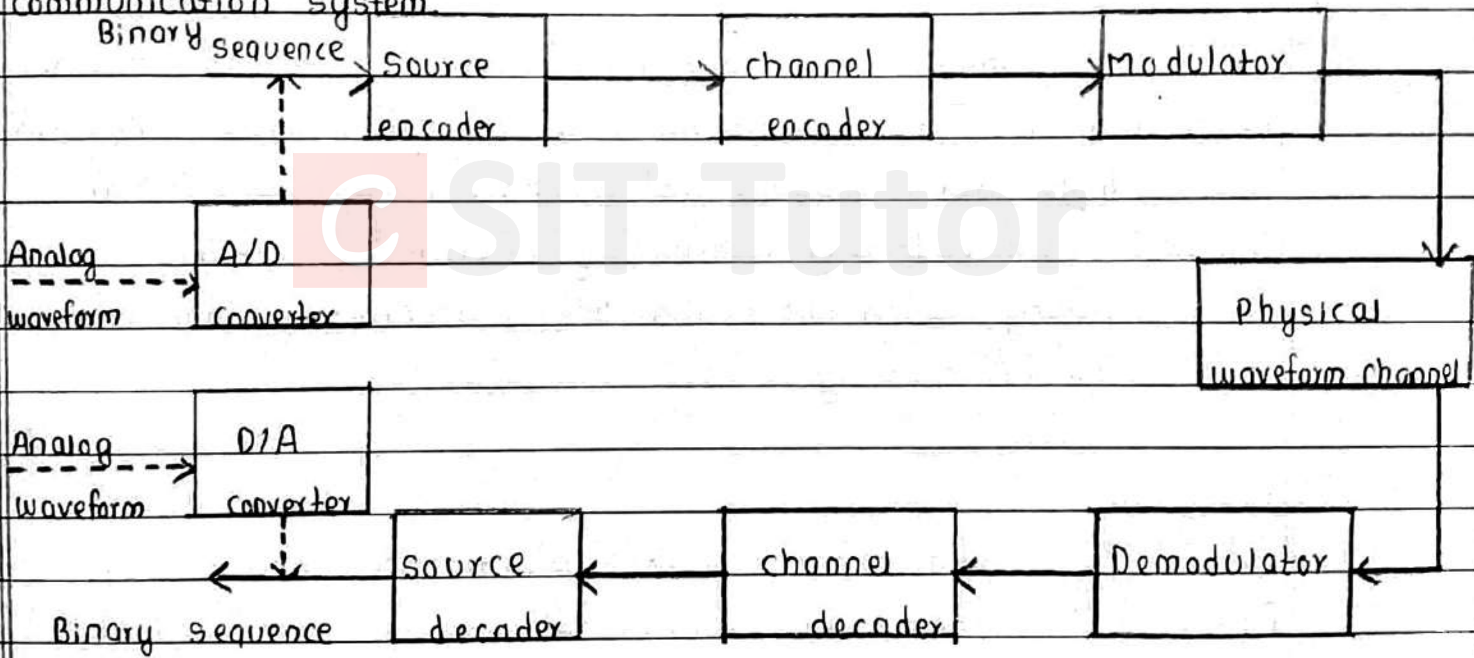


Fig:- Functional block diagram of a binary digital communications system.

In a digital communications system, the source to be transmitted is discrete, both in time and in magnitude. Unless specifically generated, information sources e.g voice, video, etc are analog in nature. An analog-to-digital (A/D) convertex is needed to conver the analog source to digital representation. At the receiving end, a digital-to-analog (D/A) converter is used to recover the analog waveform before presentation to the end user. In the most basic form, digital sequences are represented in a binary format with symbol drawn from the set [0,1].

(i) Source encoder

The information sources are analog in nature. In source encoder, an analog-to-digital converter is present to convert the analog waveform to digital representation.

It removes the unknown redundant information known as data compression which helps in fast data rate exchange.

(ii) channel encoder

The channel encoder inserts a controlled amount of known redundancy prior to modulation and transmission. The reason for having both source encoder and a channel encoder is that the source encoded sequence lacking redundancy is very susceptible to channel noise channel coding, which introduces known redundancy that can be used at the receiver for decoding.

(iii) Modulator

Modulator encoder the information along with the carrier frequency to transmit the information upto the destination.

(iv) channel

The effective communication channel seen by the channel encoder/decoder pair is often refered to as coding channel. In channel, there may be the noise which helps to change the original information. The residual disturbance in the coding channel should be of the variety that channel should be of the variety that channel should be of the variety that channel coding is capable of mitigating.

(v) Demodulator

It helps to extract the information decoding or removing the carrier frequency from the signal.

(vi) channel decoder

It decreases the controlled redundancy bits from the original message.

(vii) Source decoder

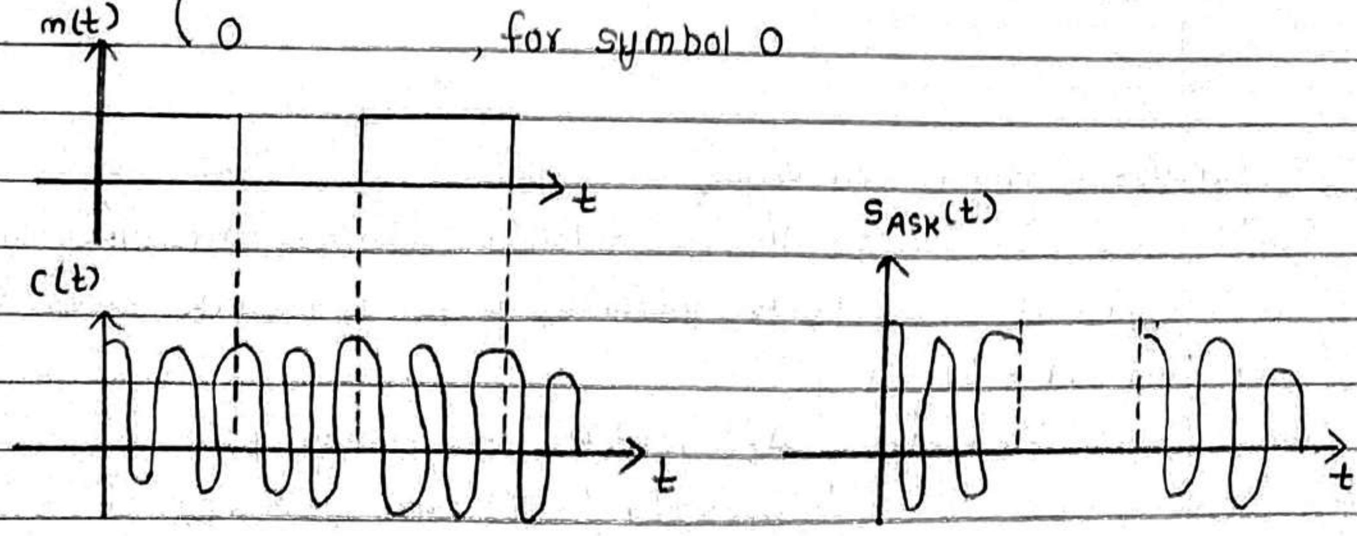
It is used to recover the analog waveform before presentation to the end user using digital to analog converter.

* Basic Modulation techniques

(i) Binary Amplitude shift Keying

In this modulation technique, the carrier's amplitude is switched between '0' and 'A_c' carrier maximum amplitude. For '0' no carrier is sent and for '1' carrier of amplitude A_c and frequency f_c is sent. The modulated signal is

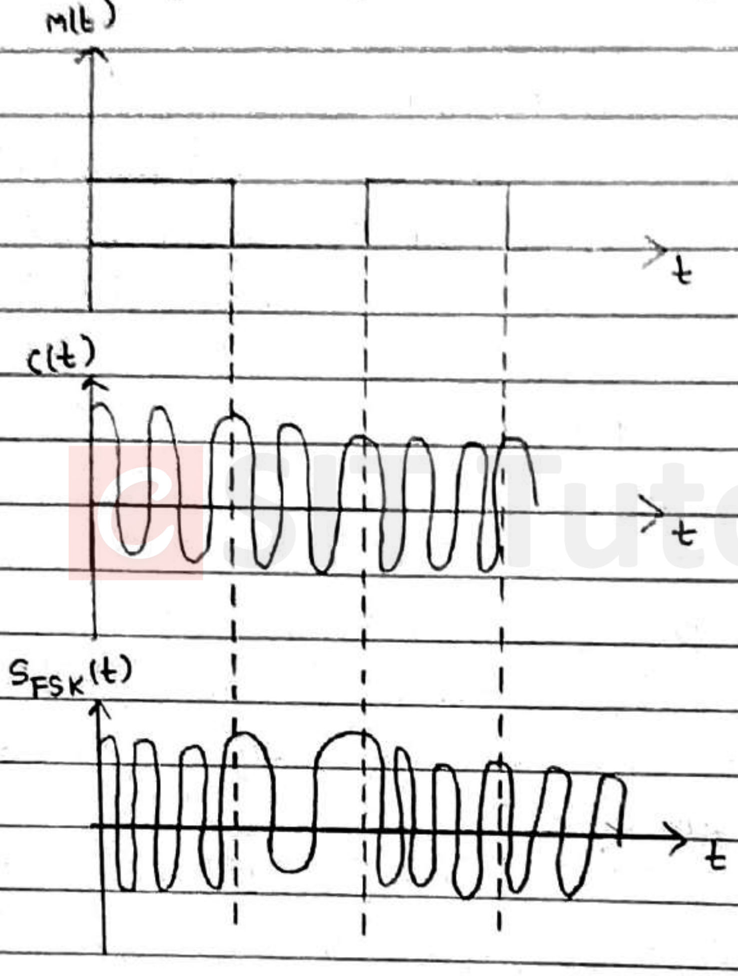
$$x(t) = \begin{cases} A_c \cos 2\pi f_c t, & \text{for symbol } 1 \\ 0, & \text{for symbol } 0 \end{cases}$$



(ii) Binary frequency shift keying

In this notation, the modulation is non-linear and $a(t)=1$. The carrier's frequency of transmitted signal is switched between two minimum and maximum frequency. The modulation signal is

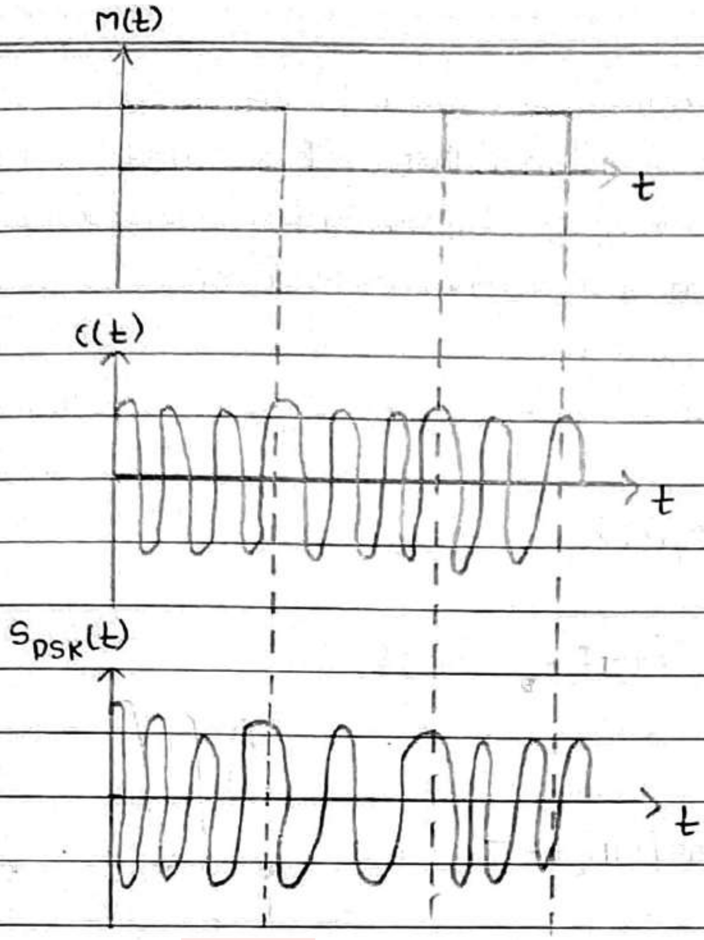
$$x(t) = \begin{cases} A_c \cos 2\pi (f_c + \Delta f) t, & \text{for symbol } 1 \\ A_c \cos 2\pi (f_c - \Delta f) t, & \text{for symbol } 0 \end{cases}$$



(iii) Binary phase shift keying

In this modulation technique, the phase of the transmitted signal is varied by 180° according to the symbol to be transmitted. The modulated signal is

$$x(t) = \begin{cases} A_c \cos 2\pi f_c t, & \text{for symbol } 1 \\ A_c \cos(2\pi f_c t + \pi), & \text{for symbol } 0. \end{cases}$$



* Signal space and Decision Region

Digital modulation involves choosing a particular signal waveform $s_i(t)$ from a finite set of possible signal waveforms based on the information bits applied to the modulator.

If there are 'm' possible signals, the modulation signal set 's' can be represented as,

$S = \{s_1(t), s_2(t), \ldots, s_m(t)\}$ which is called signal space.

For binary modulation schemes, a binary information bit is mapped directly to a signal and 's' (signal space) will contain two signals. But for M-ary modulation schemes there will be M no. of signals in signal space.

Generally, the elements of 's' are analyzed in a vector space using one or more no. of basic function which provides valuable insight of the performance of particular modulation scheme

So to represent signal in vector space, it is necessary to determine the basic of signal. Basic of a signal can be defined as a common part of two or more signal space where each signal can be represented as the linear combination of that part.

For example,

For binary phase shift keying signals, the two signals in signal space 'S' are:

$$S = \{S_1(t), S_2(t)\}$$

where,

$$S_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_c t, \text{ for } 1$$

$$S_2(t) = \sqrt{\frac{2E_b}{T_b}} \cos(2\pi f_c t + \pi), \text{for } 0$$

$$= -\sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_c t$$

where, E_b = energy per bit

T_b = symbol period

The basis of these signal would be $\phi_1(t)$ where

$$\phi_1(t) = \sqrt{\frac{2}{T_b}} \cos 2\pi f_c t$$

Using this basis signal, the BPSK signal set can be represented as $S_{BPSK} = \{\phi_1(t)\sqrt{E_b}, -\phi_1(t)\sqrt{E_b}\}$

This signal set can be represented geometrically by giving the value of envelop of each signal called as constellation diagram (or signal space diagram). So signal space diagram is the graphical representation of complex envelope of each signal space. The x-axis of the constellation diagram represent the inphase components and y-axis represents the quadrature components of the signal.

The distance between signals on a constellation diagram gives how different the modulated waveforms are and how well the signals can be differentiated by a receiver. It also represents that if a constellation diagram is densely packed the corresponding modulation scheme is more bandwidth efficient.

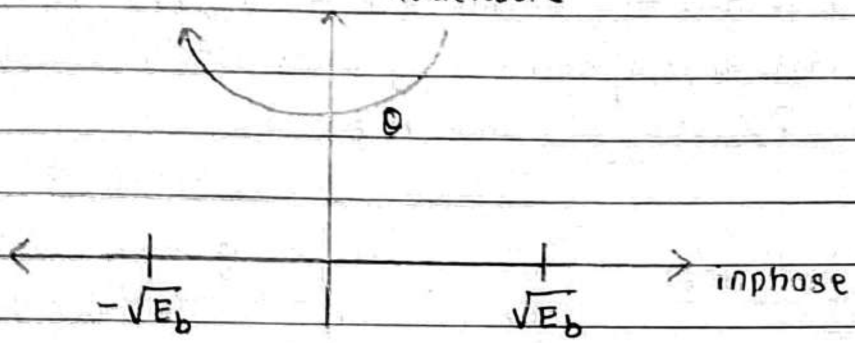


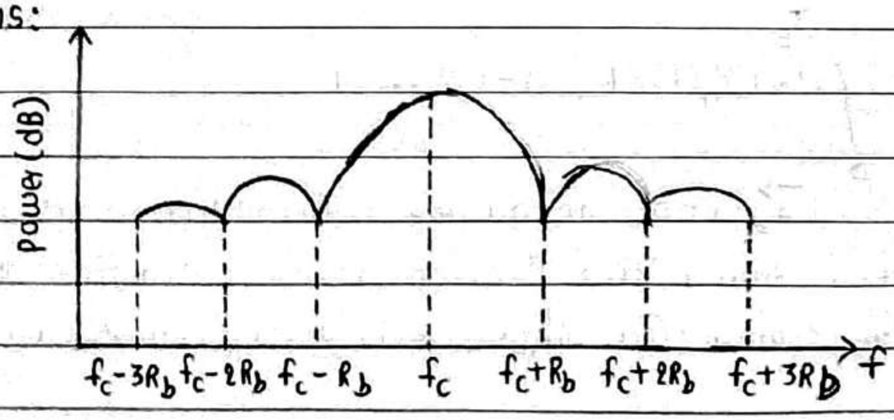
Fig:- signal space

* Spectrum and bandwidth of BPSK

The power spectral density (power spectrum) of BPSK signal would be

$$P_{BPSK}(f) = \frac{E_b}{2}\left[\left(\frac{\sin(f-f_c)}{\pi(f-f_c)T_b}\right)^2 + \left(\frac{\sin(-f-f_c)}{\pi(-f-f_c)T_b}\right)^2\right]$$

The PSD of the BPSK signal when plotted would be obtained as follows:



The first null to null B.W is found to be equal to twice the bit rate. i.e.

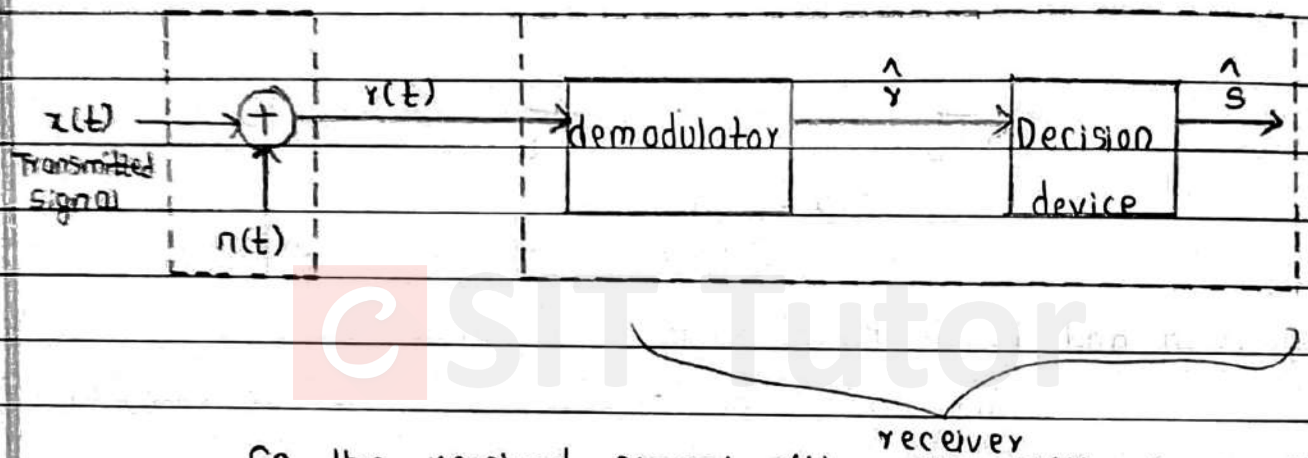
$BW = 2R_b$ where $R_b = \frac{1}{T_b}$

* Signal Detection and optimal Receiver

Detection Problem

Consider channel with white noise where the received signal is the transmitted signal plus a noise component. The following figure represent the connection block diagram of channel and receiver where M-ary transmitted signal $x(t) \in \{x_m(t)\}_{m=1}^{M}$ can be represented in a N-dimensional signal specified by basis

$$\{\phi_n(t)\}_{n=1}^{N} \quad \text{where } M \geq N$$



So, the received signal $r(t) = x(t) + n(t)$ where $n(t)$ is the noise introduced by the channel. The received signal can be represented by a vector

$$\vec{r} = \{r_1, r_2, \ldots, r_N\}$$

where,

$$r_n = \int_0^{T_s} r(t)\phi_n(t)\,dt, \quad n = 1, 2, \ldots, N$$

The $\vec{r}_N$ contains all the information of $r(t)$ which is basically a demodulated signal. The decision device estimates the information carried by the transmitted signal $x(t)$ based on the received vector $\vec{r}$.

Matched filter

As given by the equation

$$Y_n = \int_0^{T_s} r(t)\, \phi_n(t)\, dt$$

The demodulated may contain N-parallel correlators as in the figure (a). But it requires the multipliers and integrators which makes the demodulator more expensive to implement in practice. So each correlator as a demodulator can be replaced by using a matched filter and sampling the filter output at the end of symbol interval as in figure (b). The impulse response of a matched filter at n^{th} branch should be as

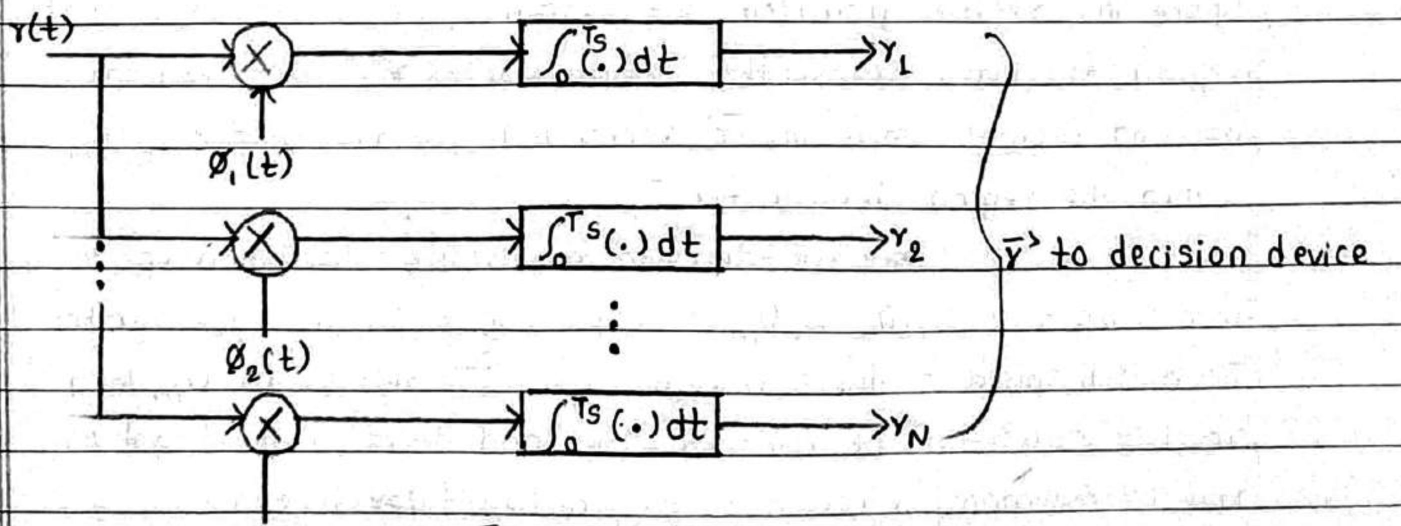
$$h_n(t) = \phi_n(T_s - t)$$

So the output of the matched filter would be

$$y_n(T_s) = \int_{-\infty}^{\infty} r(\tau)\, h_n(T_s - \tau)\, d\tau$$

$$= \int_{-\infty}^{\infty} r(\tau)\, \phi_n(\tau)\, d\tau,$$ which is equal to received vector $\vec{Y}_n$

This device at the receiver is called matched filter as it's impulse response exactly matches to the basis of the corresponding transmitted signal.



$\phi_3(t)$ Fig (a) Modulator

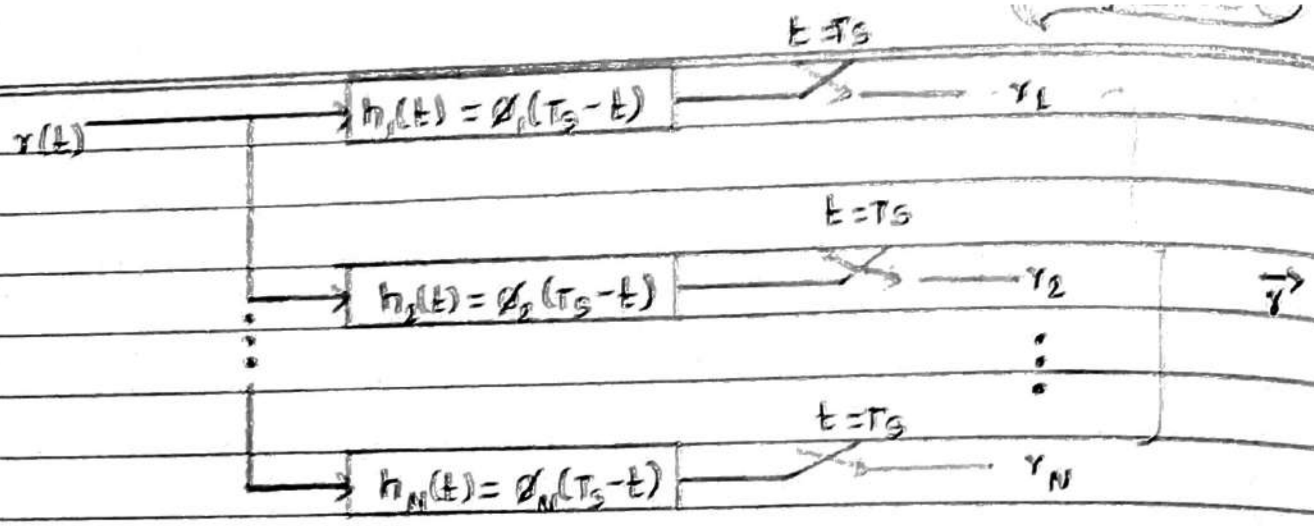


Fig:-(b) Demodulator using match filter

* Decision Region

As in the figure for signal detection the decision device has to take a value of signal transmitted by mapping $\vec{r}$ to an estimation of $\hat{s}$, in a way that it will minimized the probability of error in the decision making process.

Maximum likelihood (ML) decision rule

Set, $\hat{s} = s_m$ if $|\vec{r}-\vec{x}_m| < |\vec{r}-\vec{x}_k|$ for all $k = 1, 2, \cdots, M$ & $k \neq m$

It can also be stated as the ML decision rule to choose the message point closest to the received signal point for n-dimensional signal space the receiver partitions the signal space in regions called decision region for every transmitted signal vector $\vec{x}_m$. These regions are called decision regions where the $\vec{x}_m$ vector will be recognized by s_m, if $\vec{r}$ lies within the region defined for $\vec{x}_m$.

So for the decision region the observation space is partitioned into subspaces $D_1, D_2, \cdots, D_m$ corresponding to the message symbol $s_1, s_2, \cdots, s_m$. Each sub space is decision region so if $\vec{r}$ belongs to D_m then $\hat{s} = s_m$. The receiver consisting of matched filter and decision device that uses ML decision rule is commonly referred to as matched filter receiver.

* Minimum shift keying

It is the modulation technique in which both phase and frequency is change continuously keeping the separation between frequency minimum. Here, the modulation index is kept between 0 and 1. μ is usually 0.5.

MSK is a continuous phase frequency shift key. It is a type of keying with modulation index 0.5. A modulation index of 0.5 corresponds to minimum frequency spacing that allows two FSK signal to be coherently orthogonal and the name MSK implies minimum frequency separation.

MSK is spectrally efficient modulation scheme and is particularly attractive for use in mobile radio communication systems. It possesses properties like almost constant envelope spectrally efficient and good BER (Bit error rate) performance.

Let f_1 and f_2 denote the frequencies of two FSK signals for symbol 0 and 1 respectively. The frequency separation between two frequencies is, $\Delta = f_1 - f_2$

The modulation index denoted by 'h' is defined as the product of frequency separation and symbol period

i.e. $h = (f_1 - f_2) * T_b$

$= \frac{1}{2}$

So for MSK the two signals will be

$$x_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos[2\pi(f_c + f)t] \text{ ------- (a)}$$

$$x_2(t) = \sqrt{\frac{2E_b}{T_b}} \cos[2\pi(f_c - f)t] \text{ ------- (b)}$$

The frequency separation between these two signals

$f_1 - f_2 = f - (-f) = 2f$

So for orthogonality

$$\frac{1}{2} = 2f \times T_b$$

$$\Rightarrow f = \frac{1}{4T_b}$$

Hence, the eq^n (a) & (b) will be

$$x_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos\left[2\pi\left(f_c + \frac{1}{4T_b}\right)t\right] \text{, for 1}$$

$$x_2(t) = \sqrt{\frac{2E_b}{T_b}} \cos\left[2\pi\left(f_c - \frac{1}{4T_b}\right)t\right] \text{, for 0}$$

Now, from above two signals, the phase change of the carrier for $(t = T_b)$ is

$$2\pi \times \frac{1}{4T_b} \times T_b = \frac{\pi}{2} \text{ for 1}$$

$$2\pi \times \frac{-1}{4T_b} \times T_b = \frac{-\pi}{2} \text{ for 0}$$

Now, if the carrier signal has original phase $\phi(0)$, the above signals can be expressed as

$$x_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos\left[2\pi f_c t + \frac{\pi t}{2T_b} + \phi(0)\right] \text{ and}$$

$$x_2(t) = \sqrt{\frac{2E_b}{T_b}} \cos\left[2\pi f_c t - \frac{\pi t}{2T_b} + \phi(0)\right]$$

Consider that, $\phi(t) = \phi(0) \pm \frac{\pi t}{2T_b}$

Then, the MSK modulated signal is represented as

$$x(t) = \sqrt{\frac{2E_b}{T_b}} \cos\left[2\pi f_c t \pm \phi(t)\right] \quad \text{------}(1)$$

Now,

Assuming the initial phase to be only 0 or π i.e. $\phi(0) = 0$ or π, the above eqn(1) is expressed as

$$x(t) = \sqrt{\frac{2E_b}{T_b}} \cos\left(\frac{\pi t}{2T_b}\right)\cos[\phi(0)] - \sqrt{\frac{2E_b}{T_b}} \sin\left(\frac{\pi t}{2T_b}\right) \sin 2\pi f_c t \; \sin(\phi(T_b))$$

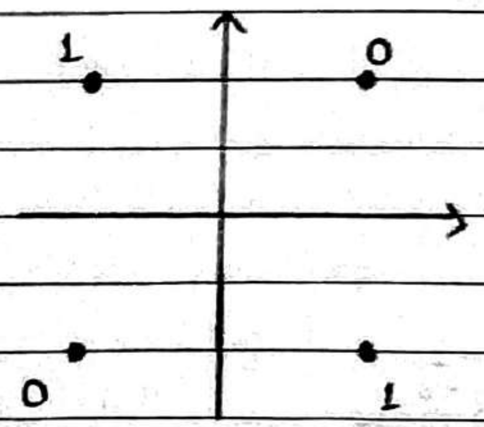
where the normalized in-phase envelope and quadrature envelope are expressed as

$$a_I = \cos[\phi(0)]$$
$$a_Q = \sin[\phi(T_b)]$$

Now to find out the signal points in constellation diagram, use following tables

a_I	a_Q	$\phi(0)$	$\phi(T_b)$	Phase change $\phi(T_b) - \phi(0)$	Symbol
1	1	0	$-\pi/2$	$-\pi/2$	0
-1	1	π	$-\pi/2$	$-3\pi/2$	1
-1	-1	π	$-\pi/2$	$-\pi/2$	0
1	-1	0	$-\pi/2$	$\pi/2$	1

so, constellation diagram will be



* MSK transmitter and receiver

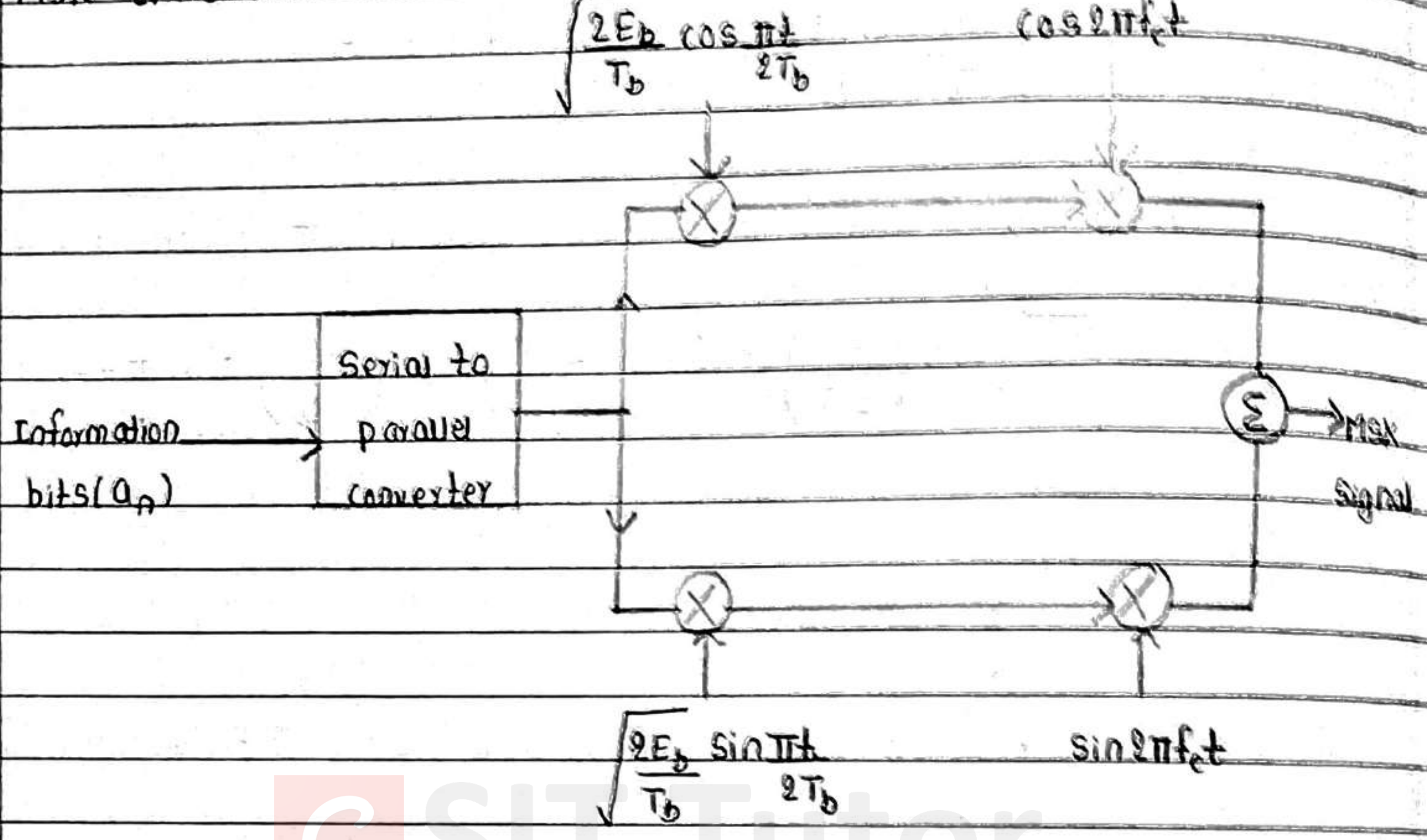


Fig:- MSK transmitter (Modulator)

Using serial to parallel converter information sequence $\{a_n\}$ with bit interval T_b is converted to a waveform $a_{1n}(t)$ and $a_{2n}(t)$ representing even part of the signal and add part of the signal.

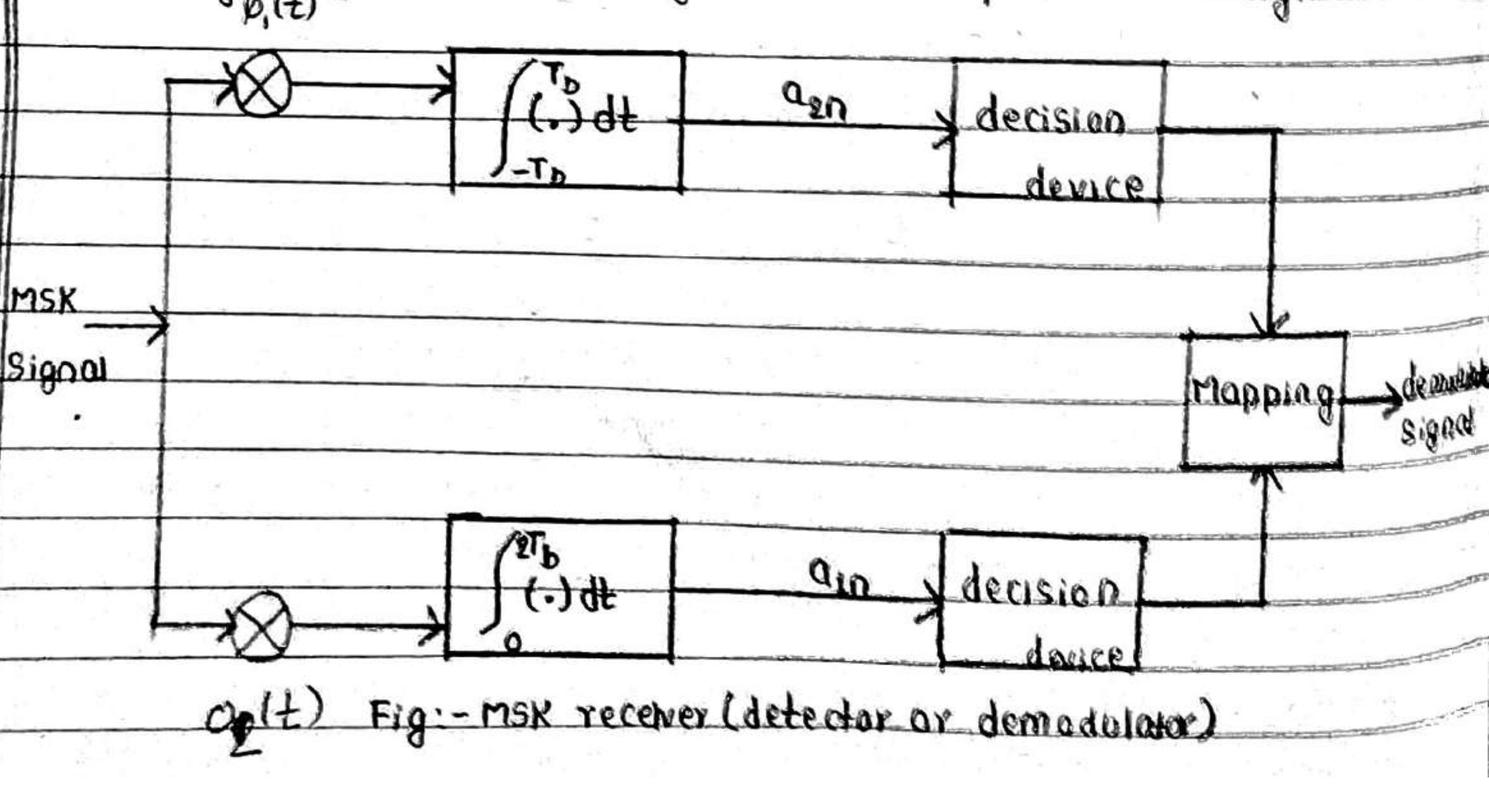


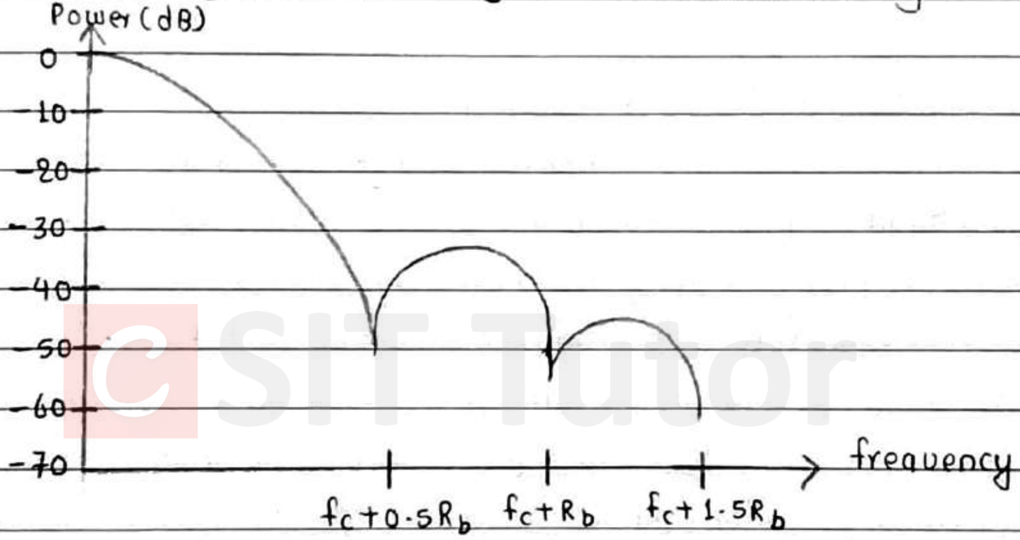
Fig:- MSK receiver (detector or demodulator)

Above figure is MSK receiver which consist of a signal integrator to approximate each bits and a decision device which devices whether a received bits is 0 or 1 on each branch. Both outputs of decision device are then mapped to generate the original transmitted sequence.

MSK power spectrum

The power spectral density for MSK signal can be expressed as

$$P_{MSK}(f) = \frac{16}{\pi^2}\left[\frac{\cos 2\pi(f+f_c)}{1-16f^2T^2}\right]^2 + \frac{16}{\pi^2}\left[\frac{\cos 2\pi(f-f_c)}{1-16f^2T^2}\right]^2$$



* Gaussian Minimum Shift Keying

For the mobile radio transmission it is desired that the modulation scheme have compact power spectral density, small or no envelope, fluctuation and high transmission accuracy. One efficient way to obtain such feature of the transmitted signal is to filter properly the baseband signal ~~is~~ before modulation. If the source signal is prefiltered using gaussian pulse shaping filter, the resultant modulated signal is received. This is called Gaussian minimum shift keying.

If H(f) is the frequency response of such Gaussian filter it should be in the following form.

$$H(f) = e^{(-f^2/B^2)(\ln 2/2)} \quad \text{------(1)}$$

where, B = Bandwidth of the filter

f = frequency of baseband signal.

The impulse response of the filter in time domain is obtained by using inverse Fourier transform of H(f).

$$h(t) = F^{-1}[H(f)] = \sqrt{\frac{2\pi}{\ln 2}}\, B \left[e^{-\frac{2\pi^2 B^2 t^2}{\ln 2}}\right] \quad \text{------(2)}$$

GMSK Transmitter and Receiver

The simple way to generate GMSK signal is to pass the baseband signal to a pre-modulation filter (Gaussian low pass filter or Gaussian pulse shaping filter) having impulse response as given by h(t). In above equation and passing the output of filter to MSK transmission. The modulation scheme is as shown in diagram.

This system is very popular in digital communication system as digital packet data and global system for mobile communication.

Fig:- Block diagram of a GMSK transmitter

The receiver of GMSK signal is as follows:

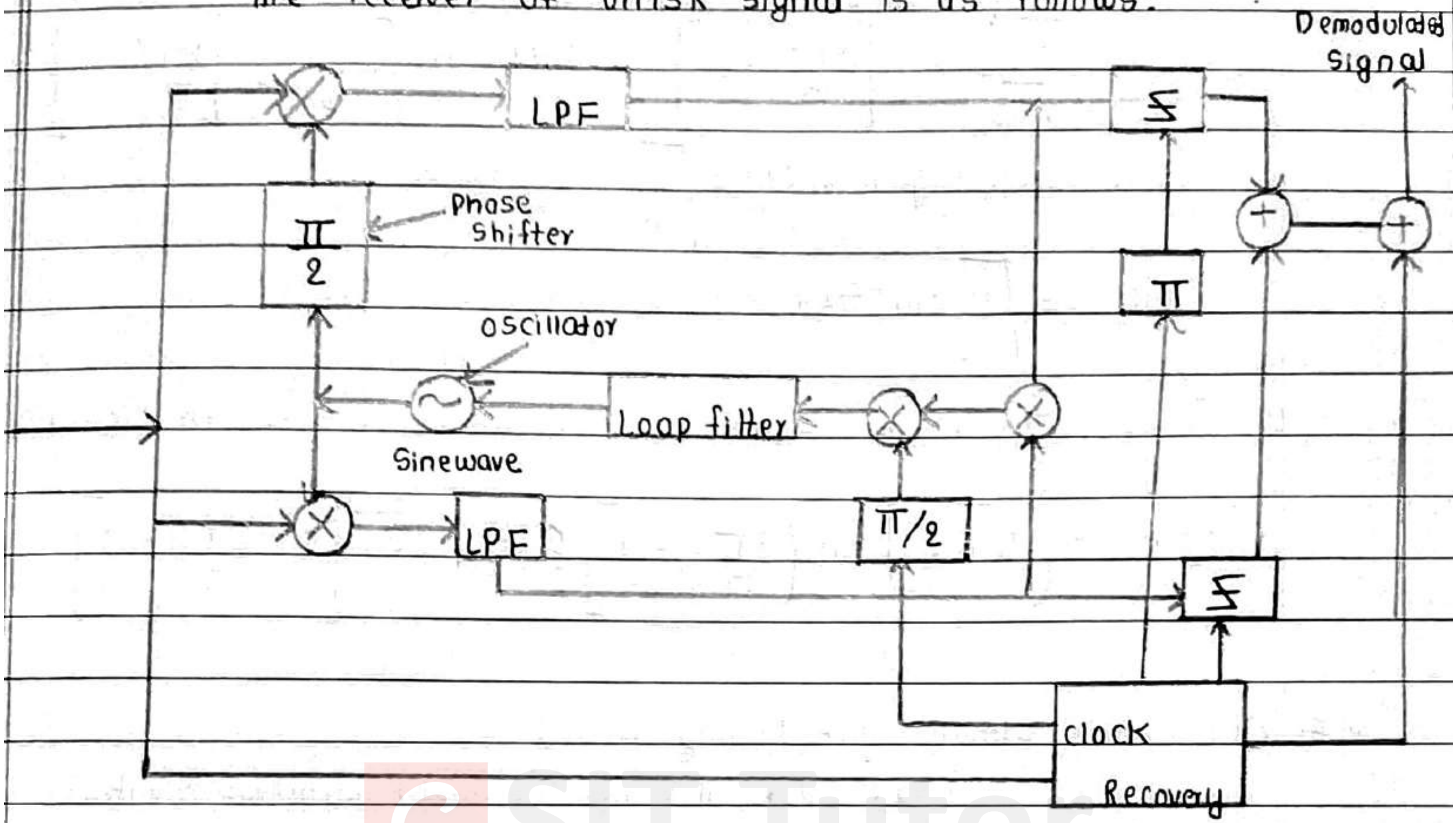


Fig:- GMSK demodulator

Advantages of GMSK

Improve spectral efficiency

GMSK signals are highly noise immune in comparison to FSK & MSK.

M-ary phase shift keying (MPSK)

In M-ary PSK, the carrier phase takes one of 'M' possible values, namely $\theta_i = 2(i-1)\frac{\pi}{M}$ where $i = 1, 2, \cdots, M$. The M-ary modulated signal can be expressed as

$$S_i(t) = \sqrt{\frac{2E_s}{T}} \cos\left[2\pi f_c t + \frac{2\pi}{M}(i-1)\right] \quad \text{for } i = 1, 2, \cdots, M$$

where,

E_s is the energy per symbol and 'T_s' is the symbol period.

The above equation can be written as

$$s_i(t) = \sqrt{\frac{2E_s}{T_s}} \cos\left[\frac{2\pi(i-1)}{M}\right] \cos 2\pi f_c t - \sqrt{\frac{2E_s}{T_s}} \sin\left[\frac{2\pi(i-1)}{M}\right] \sin 2\pi f_c t$$

choosing basis signal $\phi_1(t) = \sqrt{\frac{2}{T_s}} \cos 2\pi f_c t$ &

$$\phi_2(t) = \sqrt{\frac{2}{T_s}} \sin 2\pi f_c t,$$

The M-ary signal can also be represented in signal space as

$$s_{MPSK}(t) = \left\{ \sqrt{E_s}\left[\cos\left(\frac{2\pi}{M}(i-1)\right)\right] \phi_1(t),\ \sqrt{E_s}\left[\sin \frac{2\pi}{M}(i-1)\right] \phi_2(t) \right\}$$

Quadrature phase shift keying (QPSK)

In QPSK, the four different symbols are to be transmitted which are to be represented as equivalent of two bits. so this scheme can also be called as 4-ary modulation scheme. so for QPSK the modulated signal equation is expressed as

$$s_{QPSK}(t) = \sqrt{\frac{2E_s}{T_s}} \cos\left[2\pi f_c t + \frac{2\pi}{4}(i-1)\right] \quad \text{where } i = 1,2,3,4$$

So, the possible 4 signals for QPSK are

$$s_1(t) = \sqrt{\frac{2E_s}{T_s}} \cos 2\pi f_c t$$

$$s_2(t) = -\sqrt{\frac{2E_s}{T_s}} \sin 2\pi f_c t$$

$$s_3(t) = -\sqrt{\frac{2E_s}{T_s}} \cos 2\pi f_c t$$

$$s_4(t) = \sqrt{\frac{2E_s}{T_s}} \sin 2\pi f_c t$$

Now for these 4 signals lets choose two basis as

$$\phi_1(t) = \sqrt{\frac{2}{T_S}} \cos 2\pi f_c t \ \& \ \phi_2(t) = \sqrt{\frac{2}{T_S}} \sin 2\pi f_c t$$

So, in signal space the 4-signals can be represented as

$$S_{QPSK}(t) = \{ \sqrt{E_S}\,\phi_1(t), -\sqrt{E_S}\,\phi_1(t), \sqrt{E_S}\,\phi_2(t), -\sqrt{E_S}\,\phi_2(t) \}$$

Now, the constellation diagram for QPSK signal will contain 4-points as shown in following figure

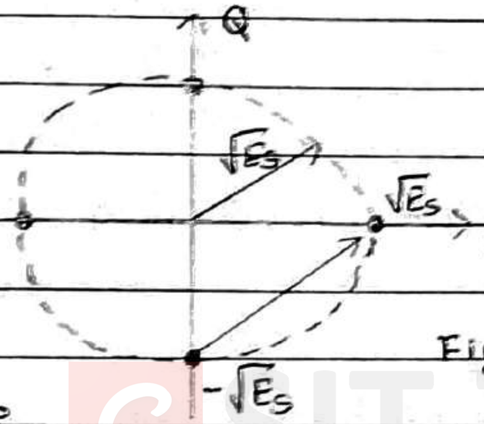


Fig:- constellation diagram for QPSK signal.

Euclidean distance

$$(d) = \sqrt{(\sqrt{E_S})^2 + (\sqrt{E_S})^2}$$
$$= \sqrt{2E_S}$$

QPSK transmitter and receiver

At the input of QPSK transmitter the message bit stream m(t) is splitted into two bit streams $m_I(t)$ & $m_Q(t)$. The bit stream $m_I(t)$ is called even bit stream and $m_Q(t)$ is called odd bit stream. The binary sequences are then separated modulated by two carrier $\phi_1(t)$ & $\phi_2(t)$. The two modulated signals each of which can be considered to be BPSK signal are assumed to produce QPSK signal. The bandpass filter at the output of the transmitter is used to pass the main signal containing band and to reject the out of band spurious components.

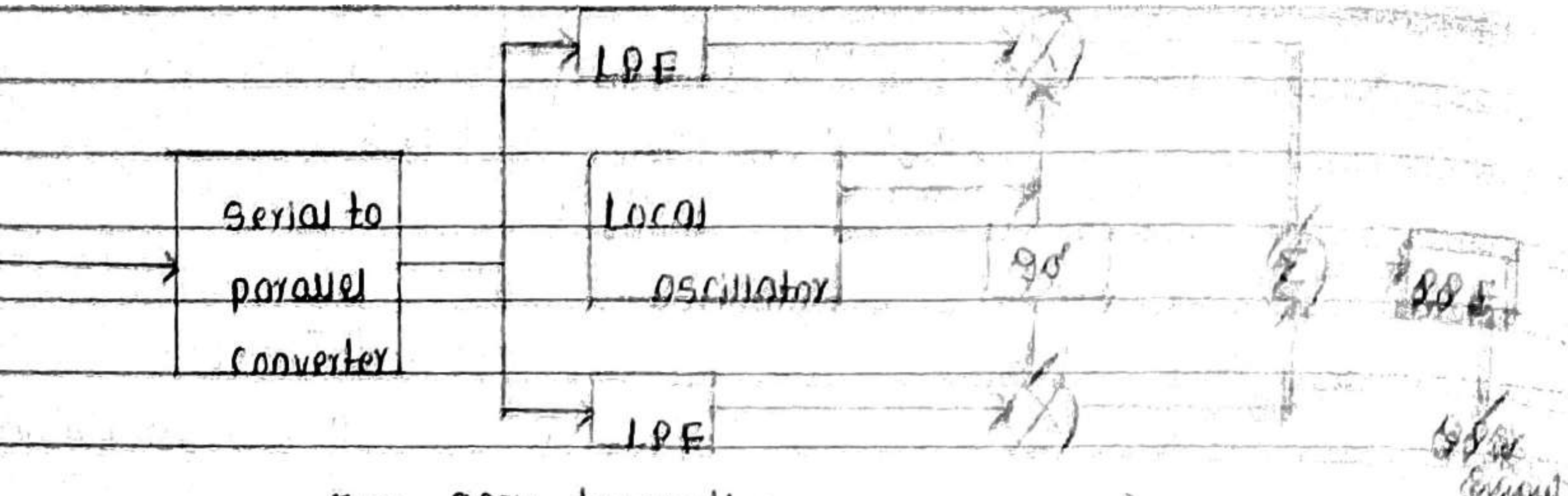


Fig:- QPSK transmitter

LPF = Low Pass filter BPF = Band Pass filter

At the receiver side the QPSK is detected coherently. The BPF at the starting and rejects out of band noise. The output of the BPF is then splited in two parts where each parts are separately demodulated. The carrier recovery circuit generates the signal having frequency equal to that of carrier frequency. The outputs of demodulators are passed through decision device which generates the inphase and quadrature binary strings. The two components are then multiplexed to reproduce the original binary sequence.

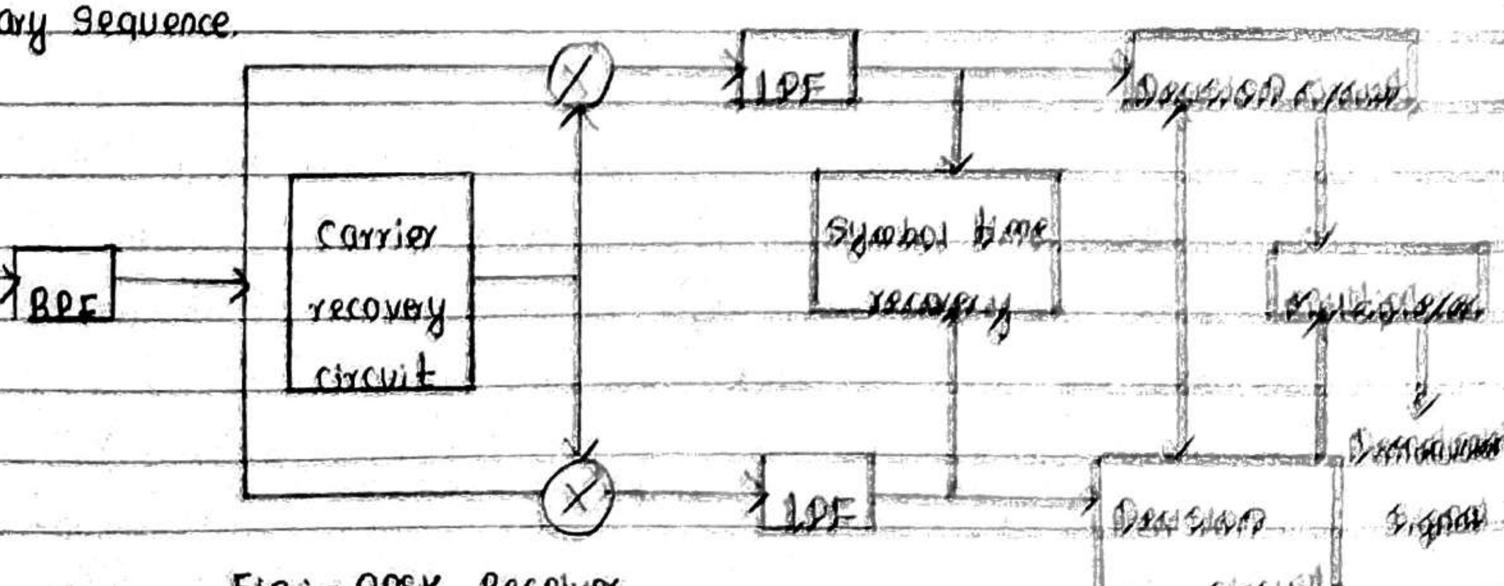


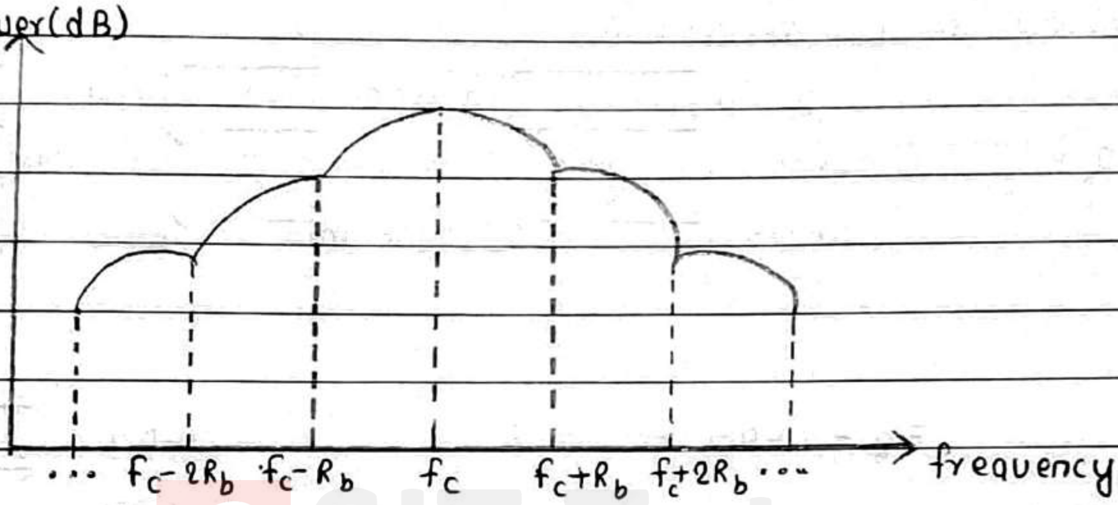
Fig:- QPSK Receiver

Spectrum of QPSK signal

The power spectral density of a QPSK signal is obtained in similar manner to that of in BPSK with bit period 'T_b' replaced by 'T_s'

$$P_{QPSK}(f) = \frac{E_s}{2}\left[\left(\frac{\sin\pi(f-f_c)T_s}{\pi(f-f_c)T_s}\right)^2 + \left(\frac{\sin\pi(-f-f_c)T_s}{\pi(-f-f_c)T_s}\right)^2\right]$$

on plotting this equation



Orthogonal frequency division multiplexing (OFDM)

OFDM is a frequency division multiplexing scheme used as a digital multicarrier modulation method. A large no. of closely spaced orthogonal sub carrier signals are used to carry data on several parallel data strings or channel. Each sub carrier is modulated with convential modulated scheme like QPSK, BPSK, etc.

The incoming digital signals are first converted into parallel data bit streams which are digital data in frequency domain. Each data in frequency domain are converted into time domain using inverse Fourier transform. Every data are then modulated by using different different sub carrier that are required to maintain orthogonality.

To mitigate the problem of time dispersion, extra bits called cyclic prefix are added. The signal is then transmitted where the data under transmission have different frequency and occupy the channel.

Towards receiver side the signal are processed through a block that deletes cyclic prefix and are demodulated and also finally converted into digital counter part. This parallel data are then converted to frequency domain by Fourier transform. The data are then serialized and passed to the decision device.

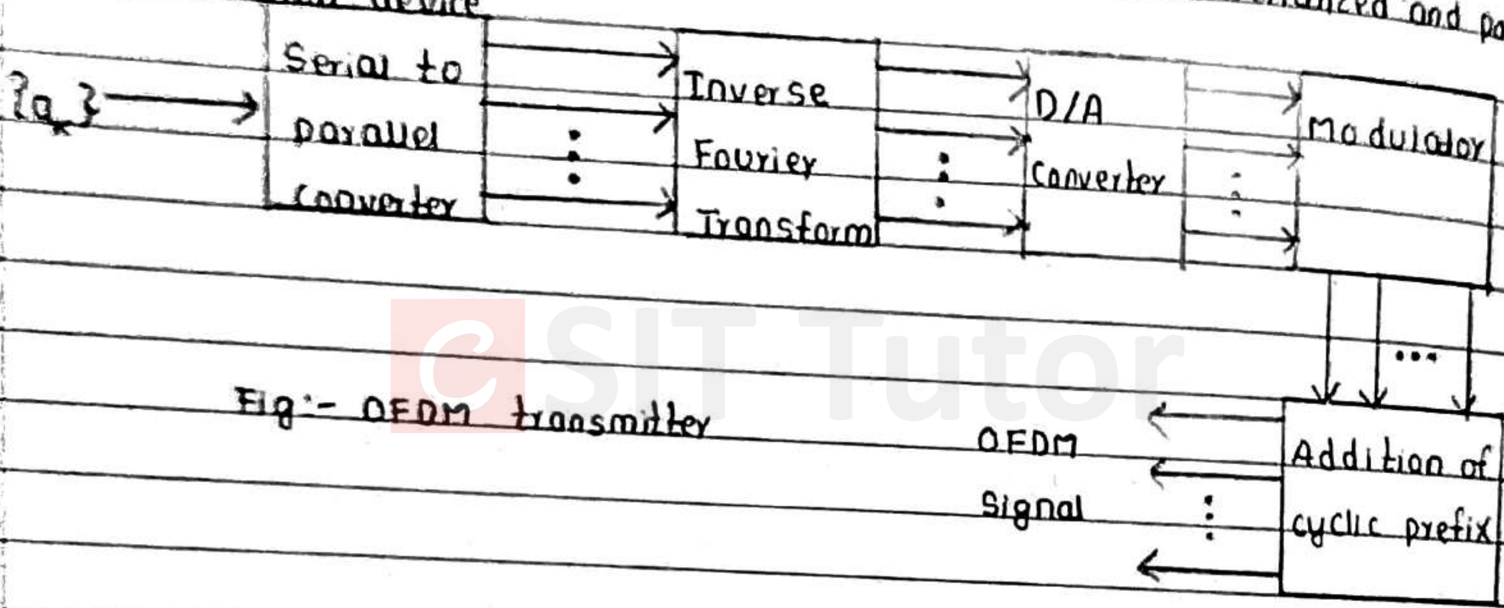


Fig:- OFDM transmitter

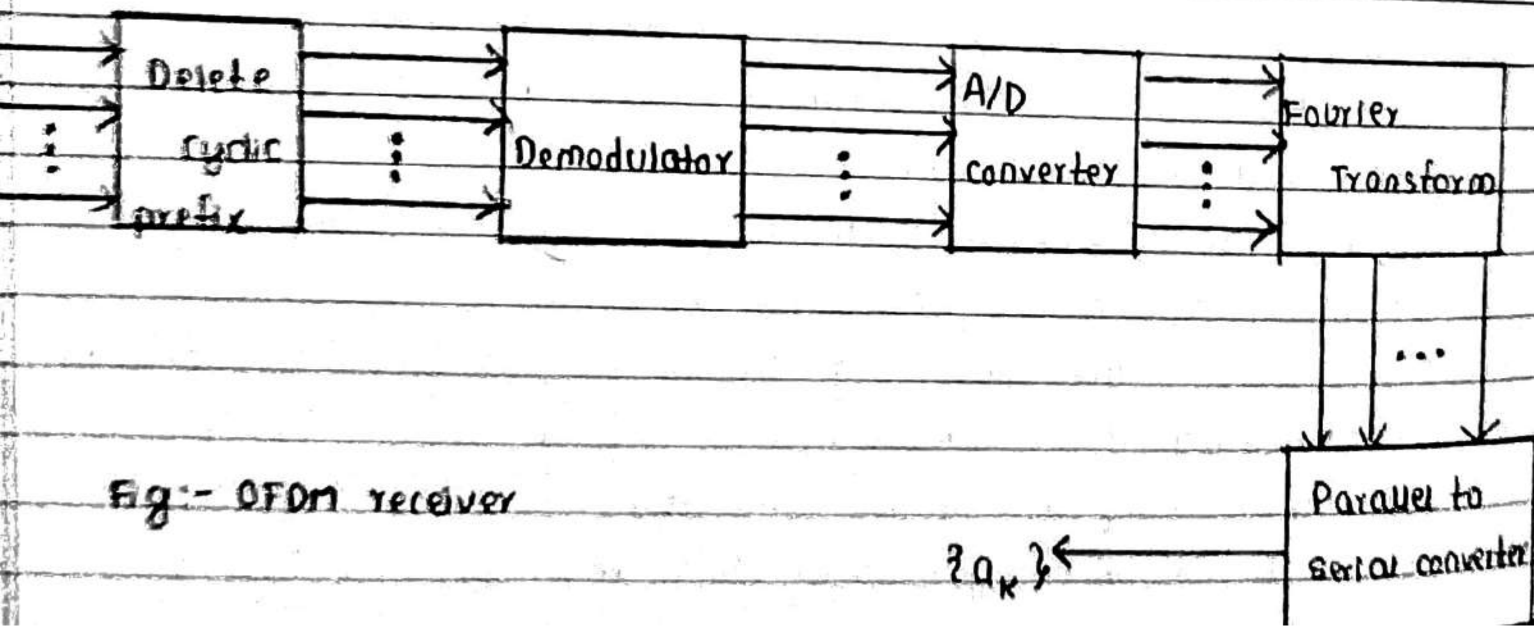


Fig:- OFDM receiver

Probability of Transmission error

Consider for a BPSK signal, the detection of a transmitted symbol over the time interval $[0, T_b]$. The transmitted signal is

$$x(t) = \begin{cases} \sqrt{E_b}\,\phi_1(t), & \text{for symbol } 1 \\ -\sqrt{E_b}\,\phi_1(t), & \text{for symbol } 0 \end{cases}$$

Now, the received signal $r(t)$ along with the noise $n(t)$ is considered a process with zero mean with PSD $\frac{N_0}{2}$ so, $\vec{r}$ is considered to be a random variable with variance $\sigma_N^2 = \frac{N_0}{2}$ and condition mean $\mu_1 = \sqrt{E_b}$ given that 1 was sent and $\mu_0 = -\sqrt{E_b}$ given that 0 was sent. Then the conditional probability density function of any received vector r, having gaussian distribution is expressed as

$$f_{r_1}(x) = \frac{1}{\sqrt{2\pi}\,\sigma_N} e^{\frac{-(x-\mu_i)^2}{2\sigma_N^2}}, \quad i = 0, 1$$

For the two decision regions, as

$$\text{if } \begin{cases} r_1 \geq 0, & 1 \text{ was sent} \\ r_1 < 0, & 0 \text{ was sent} \end{cases}$$

Now,

For equiprobable scenario, the probability of transmission error is

$$P_b = P(r_1 \geq 0 \mid \text{"0"})\,P(0) + P(r_1 < 0 \mid \text{"1"})\,P(1)$$

$$= \frac{1}{2}\left[\int_0^{\infty} \frac{1}{\sqrt{2\pi}\,\sigma_N} e^{\frac{-(x-\mu_0)^2}{2\sigma_N^2}}\,dx + \int_{-\infty}^{0} \frac{1}{\sqrt{2\pi}\,\sigma_N} e^{\frac{-(x-\mu_1)^2}{2\sigma_N^2}}\,dx\right]$$

$$= \frac{1}{2}\left[\underbrace{\int_0^{\infty} \frac{1}{\sqrt{2\pi}\,\sigma_N} e^{\frac{-(x+\sqrt{E_b})^2}{2\sigma_N^2}}\,dx}_{\text{First integral}} + \underbrace{\int_{-\infty}^{0} \frac{1}{\sqrt{2\pi}\,\sigma_N} e^{\frac{-(x-\sqrt{E_b})^2}{2\sigma_N^2}}\,dx}_{\text{Second integral}}\right]$$

First integral

put, $\frac{x+\sqrt{E_b}}{\sigma_N} = z$

So, as $x \to 0$, $z \to \frac{\sqrt{E_b}}{\sigma_N}$

as $x \to \infty$, $z \to \infty$

Also,

$$x = z\sigma_N - \sqrt{E_b}$$

So, $dx = \sigma_N dz$

Second integral

put, $\frac{x-\sqrt{E_b}}{\sigma_N} = z$

as, $x \to 0$, $z \to -\frac{\sqrt{E_b}}{\sigma_N}$

as $x \to -\infty$, $z \to -\infty$

Also,

$$x = z\sigma_N + \sqrt{E_b}$$

So, $dx = \sigma_N dz$

Then,

$$P_b = \frac{1}{2\sqrt{2\pi}\,\sigma_N}\left[\int_{\frac{\sqrt{E_b}}{\sigma_N}}^{\infty} e^{-\frac{z^2}{2}}(\sigma_N)\,dz + \int_{-\infty}^{-\frac{\sqrt{E_b}}{\sigma_N}} e^{-\frac{z^2}{2}}(\sigma_N)\,dz\right]$$

$$= \frac{1}{2\sqrt{2\pi}}\left[\int_{\frac{\sqrt{E_b}}{\sigma_N}}^{\infty} e^{-\frac{z^2}{2}}\,dz + \int_{\frac{\sqrt{E_b}}{\sigma_N}}^{\infty} e^{-\frac{z^2}{2}}\,dz\right]$$

$$= \frac{1}{\sqrt{2\pi}}\left[\int_{\frac{\sqrt{E_b}}{\sigma_N}}^{\infty} e^{-\frac{z^2}{2}}\,dz\right]$$

since,

$$\frac{1}{\sqrt{2\pi}}\int_{\frac{\sqrt{E_b}}{\sigma_N}}^{\infty} e^{-z^2/2}\,dt = Q(t)$$

So,

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

where, $Q(t)$ is called Q function and gives the area under the Gaussian curve.

So, probability of bit error,

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

$$= Q\left(\frac{\sqrt{E_b}}{\sqrt{\frac{N_0}{2}}}\right)$$

$$= Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

Also, the probability of ~~detec~~ correct detection (P_c) is expressed as

$$P_c = 1 - P_b$$

$$P_c = 1 - Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

Now,

For QPSK the probability of detection error for odd number bits and even numbered bits are same and also it is noted that each symbol is composed ~~two~~ of two bits.

So, $P_b = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$ & $P_c = 1 - P_b$

For two symbols

$$P_c = (1-P_b) * (1-P_b) = \left[1 - Q\left(\sqrt{\frac{2E_b}{N_0}}\right)\right]^2$$

Hence, probability of symbol error

$$P_s = 1 - P_c = 1 - \left[1 - Q\left(\sqrt{\frac{2E_b}{N_0}}\right)\right]^2$$

So, for M-ary, modulation scheme, probability of symbol error

$$P_s = 1 - \left[1 - Q\left(\sqrt{\frac{2E_b}{N_0}}\right)\right]^{\log_2 M}$$

unit -4

Receiver Techniques for fading Dispersive channels.

Overview of channel impairment mitigation techniques.

In a wireless communication system due to user mobility and multipath propagation, transmitted signals often experience channel fading and time dispersion. For the mitigation of these channel impairments different techniques can be used such as

Power control

Diversity

channel equalization

channel coding and interleaving

Power control

Power control is used to maintain the minimum necessary transmitted power. In cellular systems, the service area is divided into radio cells. Mobile users in each cell are served by single base station. Different cells can use the same frequency spectrum subject to the interference constraint. Power control is essential in cellular systems for high system capacity and satisfactory transmission quality by

limiting interference among users in different cells using the same frequency channel (i.e. cochannel interference

alleviating the near-far effect which reduces interference among users in the same or different cells using adjacent frequency channels.

Near-far effect is a condition in which a nearby transmitter captures the receiver of the mobile or base station so that the latter is unable to detect the signal of a ~~sig~~ second transmitter located farther away. The near-far effect can be equalized through power control.

There are two types of power control.

(a) open-loop power control

In this technique, the transmitter itself estimates the nature of channel and adjusts its transmission power accordingly. The channel estimates is based on nearby links. As no feedback hardware is required this technique avoids the resource overhead for signalling and is relatively faster. However it suffers from power estimation accuracy.

(b) close-loop power control

In this technique, the receiver estimates the received power level, the received signal to interference ratio and the transmission error rate and compares the measurement with target desired value and sends the information to transmitter for power control adjustment. It is more accurate than open loop power control.

(ii) Diversity

Diversity is an effective way to combat channel fading. Diversity improves transmission performance by making use of more than one independently faded version of transmitted signal. If several replicas of the signal, carrying the same information, are received over multiple channels that exhibit independent fading with comparable strengths, the chances that all the ~~ide~~ independently faded signal components experience deep fading simultaneously are greatly reduced. Diversity is commonly used technique in wireless systems to combat channel fading due to following facts,

(a) The degradation of transmission quality due to channel fading cannot be simply overcome by increasing the transmitted signal power.

(b) The power available on reverse link is severely limited by the battery capacity in hand-held subscriber units. with diversity, the required transmitted power can be greatly reduced;

(c) cellular communications systems are mostly interference limited and once again, mitigation of channel fading by diversity reception can translate into improved interference tolerance which, in turn means greater ability to support additional users and therefore higher system capacity.

Different types of diversity techniques are

(a) space diversity

In this technique, the desired message is transmitted by using multiple transmitting antennas and/or receiving antennas. The number of different receiving antennas are used that are separated by certain optimal distance so this technique can also be caved as antenna diversity. The space separation between adjacent antennas should be large enough to ensure that the signals from different antennas are independently faded. Usually a separation of at least 10 carrier wavelengths is required between two adjacent antennas.

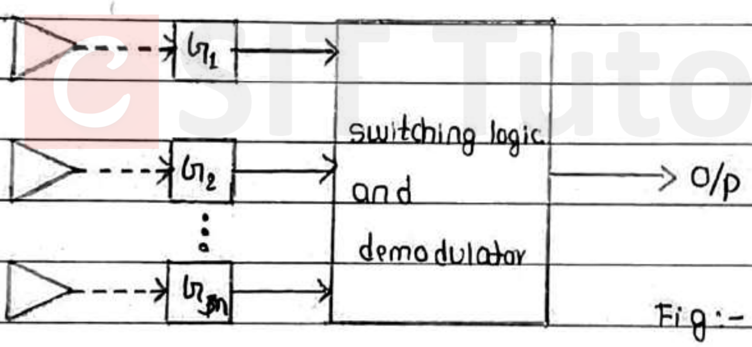


Fig:- Block diagram of space diversity.

(b) Space diversity reception method can be classified into following types

(i) Selection diversity

The receiver monitors the SNR value of each diversity channel and selects that one having maximum SNR. It is very simplest diversity technique.

(2) Feedback or scanning diversity

The number of signals received by corresponding antenna are scanned in a fixed sequence until one is found to have signal power above or equal to predetermined threshold, then the selected signal is received as the main information carrying signal. once the received signal falls below the threshold value thus scanning process is again initiated.

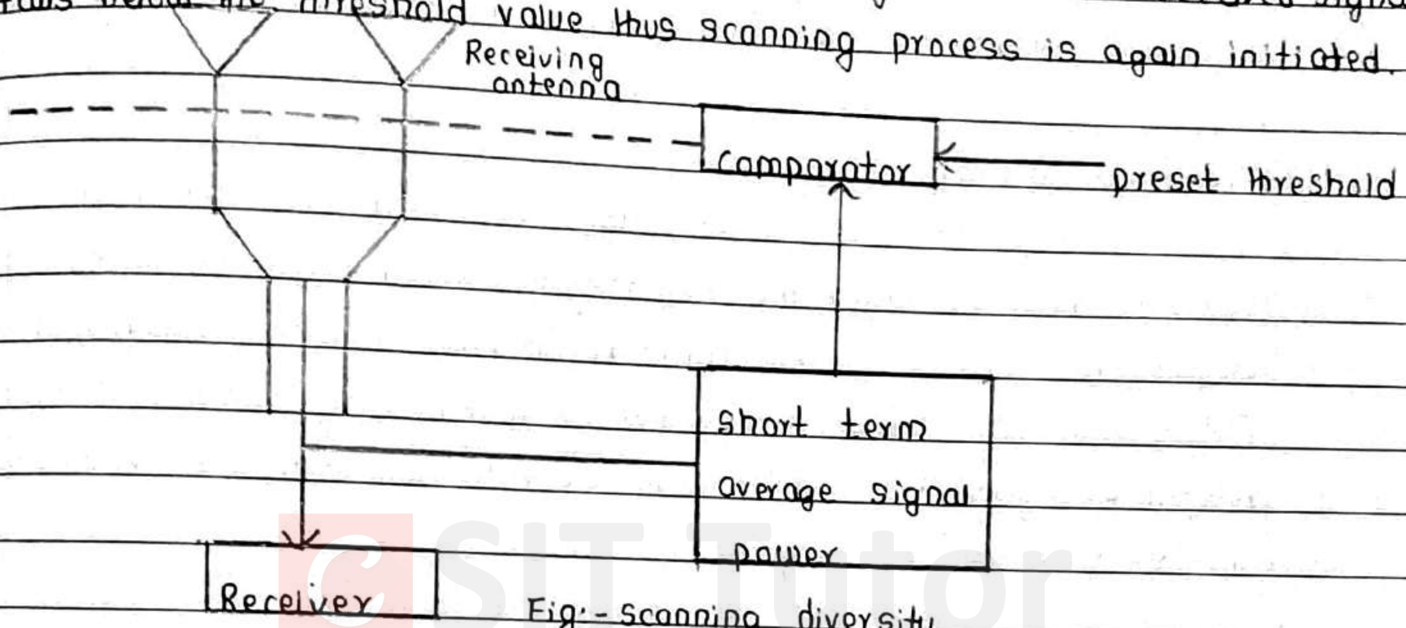


Fig:- scanning diversity

(3) Maximal ratio combining

The signal received from each branch are weighted in certain proportion to obtain the maximum SNR possible. The individual signals are cophased and summed, it has the advantage of producing an output with acceptable SNR even when none of the received signals are acceptable

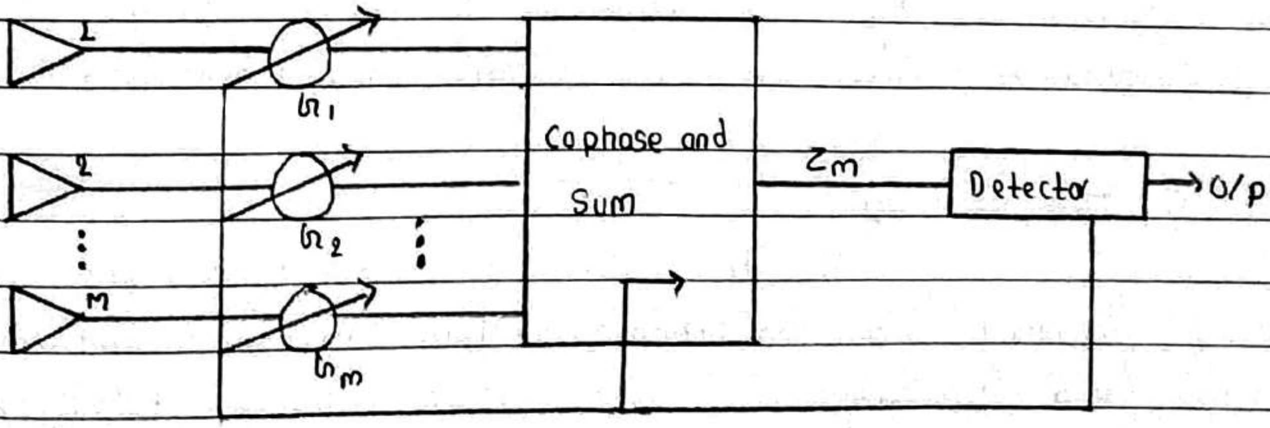


Fig:- Maximal ratio combining.

(4) Equal gain combining

It is similar to maximal ratio combining but the signals of all branches are summed without weighting, this reduces the system complexity. The possibility of producing an acceptable signal from a number of unexpectable inputs is still retained and performance is only marginally inferior to maximal ratio combining but superior to selection diversity.

(b) Frequency diversity

The desired message is transmitted simultaneously over several frequency slots. The separation between adjacent frequency slots should be larger than the channel coherence bandwidth such that channel fading over each slot is independent of that in any other slot. By using redundant signal transmission, this diversity improves link transmission quality at cost of extra frequency bandwidth.

(c) Polarization diversity

At the base station space diversity is considerably less practical as large spacing between the antenna is to be maintained in comparison to that required in mobile station. In polarization diversity technique usually two ~~pol~~ polarization diversity techniques are used which are vertical or horizontal polarization or both. The two receiving antennas separated by certain angle, one to receive vertically. Polarized signal and next to receive horizontally polarized signal are used.

(d) Time diversity

The transmitter sends information repeatedly with certain time interval the separation between the time intervals should always be greater than that of coherence time of the channel, so that the totally uncorrelated info are received by receiver. Among these different information sent at different time the information is selected by the receiver which has experience least fading and interference.

(iii) channel equalization

IST caused by multipath in band limited time dispersive channel distorts the transmitted signal causing bit error at the receiver. Equalization is a technique used to combat IST. In a radio channel the variety of adaptive equalizers can be used to cancel interference. Since, the mobile fading is random and time varying, equalizers must track the time varying characteristics of mobile channel and are thus caued adaptive equalizer.

The general operating modes of an adaptive equalizer includes two parts training and tracking. First a known fixed length training sequence is sent by the transmitter so that the receiver equalizer may adapt to a proper setting for minimum bit error rate detection. Immediately following this training sequence the user data is sent and adaptive equalizer at the receiver utilizes a recursive algorithm to evaluate the channel and estimate filter coefficient to compensate the distortion. The training sequence is designed to permit equalizer to accquire proper filter coefficient in worst possible channel conditions like fastest velocity, longest time delay spread extreme fades, etc when an equalizer is properly trained it is said to be converged.

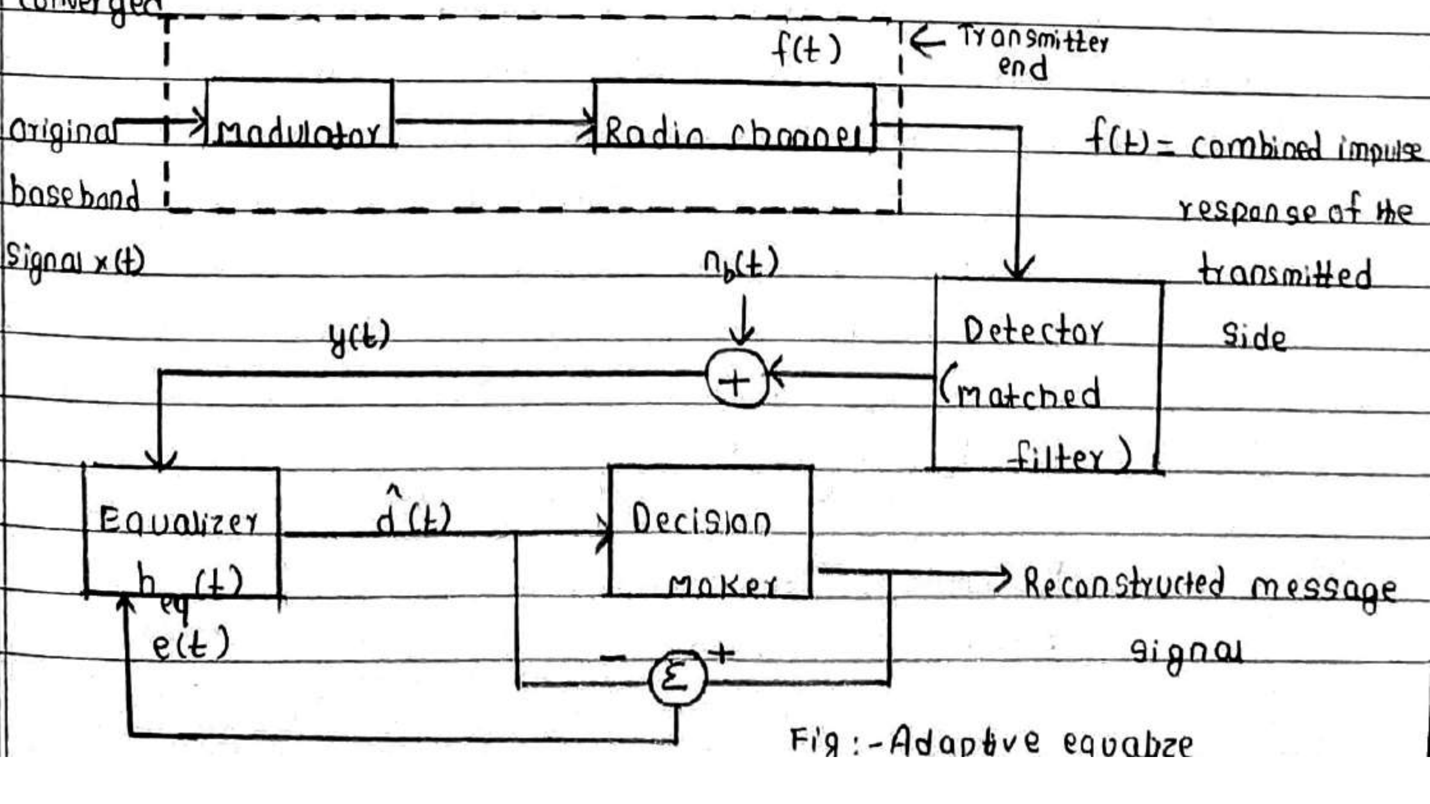


Fig :- Adaptive equalize

As in the figure communication system with adaptive equalizer, consider $x(t)$ be the original message signal and $f(t)$ be the combined impulse response of the transmitter end. The received signal $y(t)$ can be expressed as

$$y(t) = x(t) * f(t) + n_b(t)$$

where, '$*$' denote the convolution operator. If the impulse response of equalizer is $h_{eq}(t)$ then the output of equalizer is

$$\hat{d}(t) = x(t) * f(t) * h_{eq}(t) + n_b(t) * h_{eq}(t)$$
$$= x(t) * g(t) + n_b(t) * h_{eq}(t)$$

considering the baseband noise to be zero and for the best reception of original signal that is $\hat{d}(t) = x(t)$

$$g(t) = f(t) * h_{eq}(t)$$

so that

$$g(t) = \delta(t)$$

$$\hat{d}(t) = x(t) * \delta(t)$$
$$\hat{d}(t) = x(t)$$

- Linear equalizer

This type of equalizer is the simplest type of equalizer. In this type the present and past values of received signals are linearly waited by the filter coefficients and summed to produced the output as in the following figure. If the delays and filter gains are analog, the continuous output of the equalizer is sampled at the symbol rate and samples are provided to the decision device. The implementation of linear equalization is implemented in digital domain.

The output of filter before the decision is made can be expressed as

$$\hat{d}_k = \sum_{n=-N_1}^{N_2} (c_n^*) \, y_{k-n}$$

where,

c_n^* = complex filter coefficient

$\hat{d}_k$ = output at instant of k

$y_1 (= y_{K+N_1})$ = input received signal

$N = N_1 + N_2 + 1$ is total no. of taps

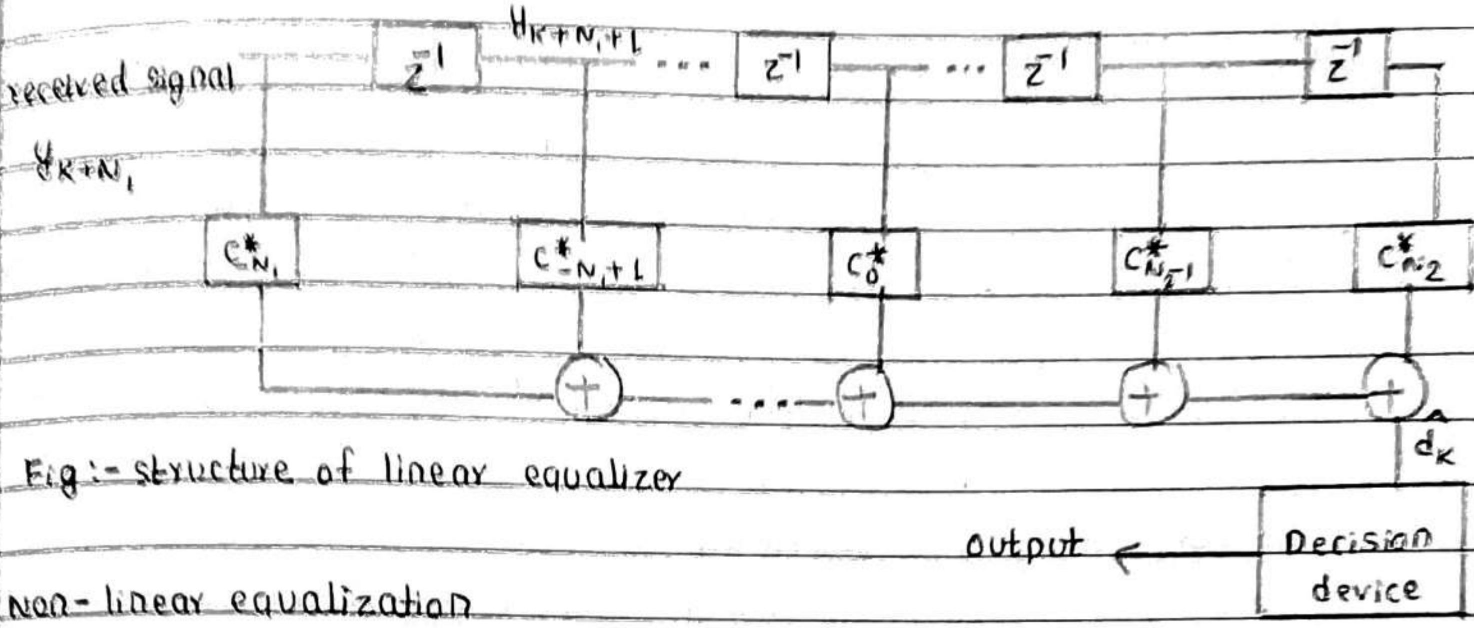


Fig :- structure of linear equalizer

- Non-linear equalization

Non-linear equalizer are used in application where the channel distortion is too severe for a linear equalizers to handle and are used frequently in wireless system. one of the most important non-linear equalizer is maximum likelihood sequence estimation equalizer

Maximum likelihood sequence estimation equalizer

This equalizer test all the possible data sequence rather than decoding each received symbol itself and chooses the data sequence with maximum probability as the output. The block diagram of MLSE receiver is shown in following figure. The MLSE is optimal in the sense that it minimizes the probability of sequence error. The MLSE requires the knowledge of channel characteristics in order to make the decisions. It also requires the knowledge of statistical distribution of noise corrupting the signal.

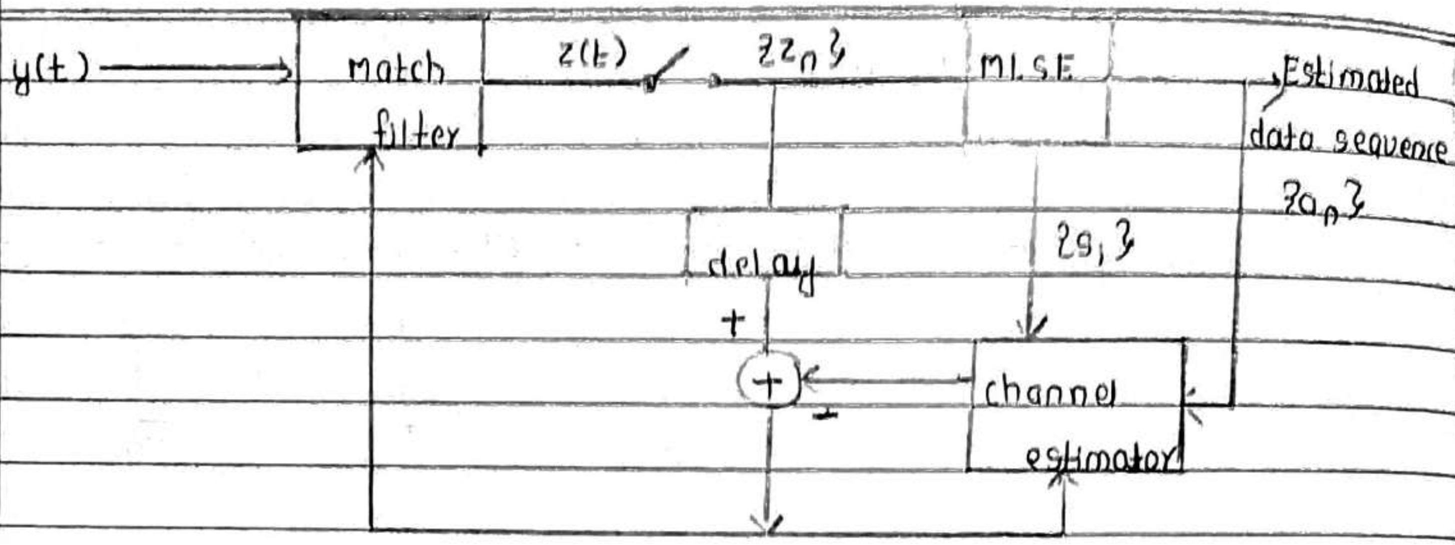


Fig:- structure of MLSE with adaptive filter.

(iv) Channel coding and Interleaving

Channel coding is an elaborate cross-checking technique to overcome transmission errors over a noisy channel. It introduces redundancy in a deterministic manner, at the transmitter, to the information sequence and exploits the controlled redundancy at the receiver demodulator output for error correction. The redundancy may seem to be waste of system resources such as frequency bandwidth. However, if an error-correction code is designed properly, using the same resources we should be able to transmit coded sequence at faster rate such that, when the transmission rate for the information bits remain the same as in uncoded system, the transmission accuracy for coded system is higher.

Channel coding is effective in combating independent random symbol errors. Interleaving is a popular to address deep fading. The coded symbols are first interleaved before being mapped to modulated waveforms in transmitter. At receiver, demodulator output symbols are then de-interleaved being applied to decoder. If the interleaver length is sufficient, the negative effect of channel fading correlation on coding gain can be eliminated.

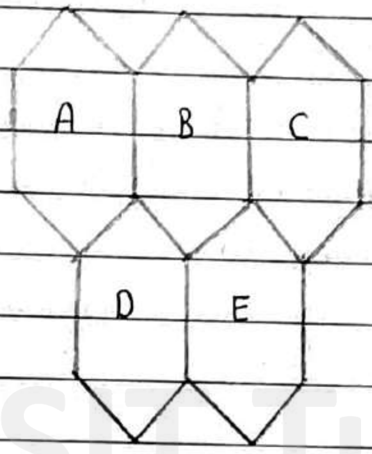
unit - 5

Fundamental of cellular communications.

* Introduction

Cell

The geographical area in cellular communication is divided into a number of uniform geometrical shapes. such uniform geometric area is called cell.



The above figure is group of cells of certain geographical area which is called cluster.

The area is divided into number of hexagonal cells. The other possible shapes are circular, square and equilateral triangle. The circular cell is not used as it can't cover the whole area i.e. there occur a gap in between the circular cells. So among other shapes, hexagonal shape is used as the hexagon covers layers area for a given distance of extreme point in comparison to that of triangle and square.

So, the objective of this type of system design is mainly to achieve high capacities with limited spectrum and higher coverage with a single high power transmitter with an antenna mounted on a tall tower in each base station.

Frequency reuse

If a given set of frequencies or radio channels, can be reused without increasing the interference, then the large geographical area covered by a small high power transmitter can be divided into a number of small areas, each allocated a subset of frequencies. With a smaller geographical coverage, lower power transmitters with lower antennas can be used, provided that the physical separation of two cells is sufficiently wide the same subset of frequencies can be used in both cells. This is concept of frequency reuse.

Frequency reuse or frequency planning is a technique of reusing frequencies and channels within a communication system to improve capacity and spectral efficiency. OR,

The use of same set of channel in the next set of cells, called cluster is frequency reuse.

Capacity of a cellular system

Consider a cellular system which has a total of 'S' duplex channel. Each cell is ~~same~~ allocated a group of 'K' channels. The 'S' channels are divided among 'N' cells so, total number of available radio channel is

$$S = KN$$

The 'N' cells which use the complete set of channels is called cluster. Here the cluster can be repeated 'M' times within the system. So, the total number of channels 'C' is used as a measure of capacity as

$$C = MKN$$

$$\therefore C = MS$$

For this equation

- The capacity is directly proportional to the number of replication M
- The cluster size N is typically equal to 4, 7, 12, 21.
- Smaller the 'N', maximum the capacity.
- The frequency reuse factor is given by $\frac{1}{N}$.

* Mobility management

(1) Handoff management

When a mobile moves from one cell to another, to keep the communication between the user pair, the user channel has to shifted from one BS to other without interupting the call. This is called handoff.

When a MS moves into another cell while the conversation is still in progress, the MSC automatically transfer the call to a new FDD channel without disturbing the conversation.

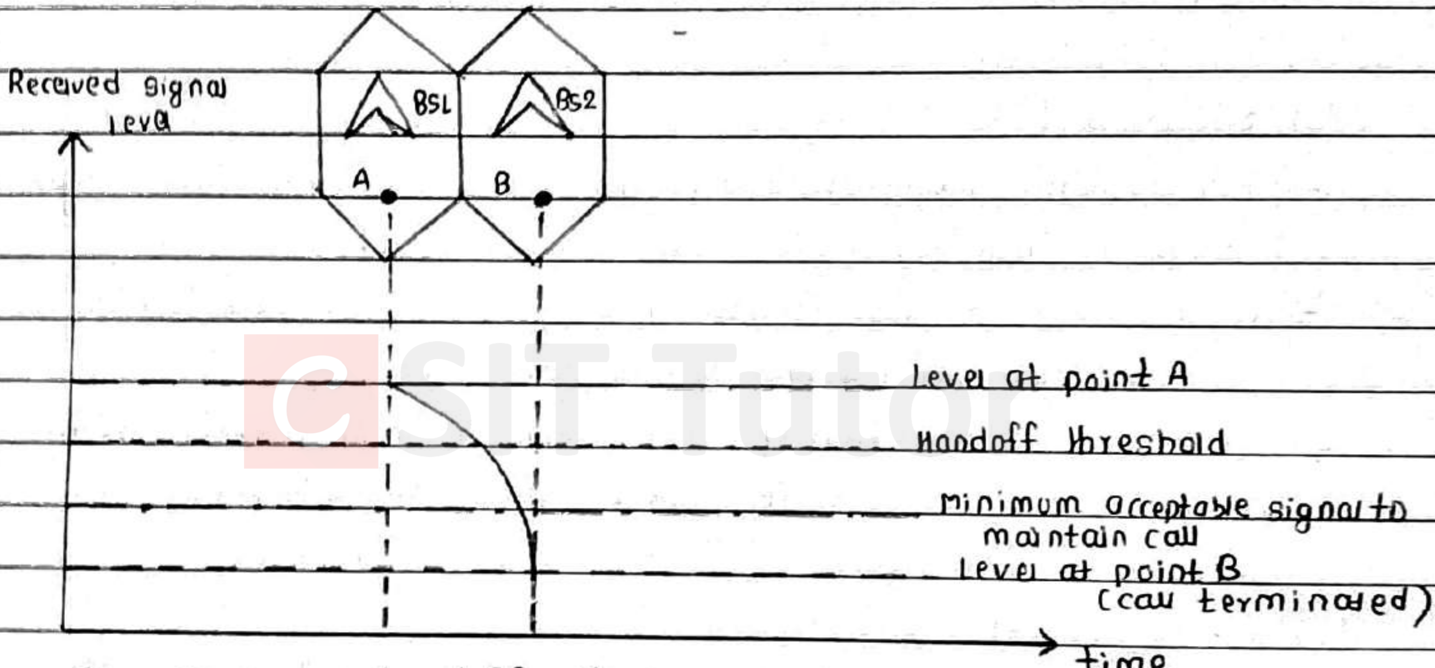


Fig:- Improper handoff situation

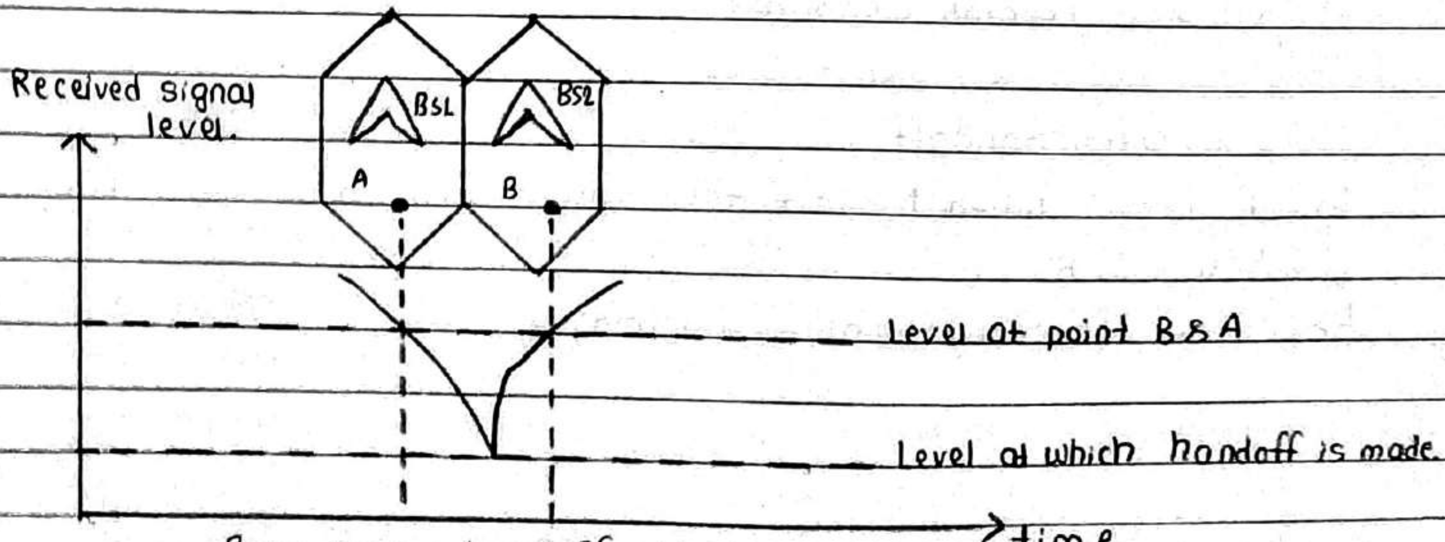


Fig:- Proper handoff situation

- Types of handoff

(i) Hard handoff

↳ It occurs when carrier frequency changes during handover from one cell to another.

↳ It is the one in which channel in source cell is released and only the channel in the target cell is engaged.

↳ Thus the connection to the source is broken before the connection to the target is made and hence called break-before-make.

↳ It is intended to be instanteneous in order to minimize the disruption to the call

(ii) soft handoff

↳ It occurs when serving cell and neighbor cell has same carrier frequency and no change in frequency takes place during handoff.

↳ It is the one in which channel in source cell is retained and used for a while until the next new channel is allocated in MS.

↳ In this type, connection to new channel is made before breaking of connected with current BS. Hence called make-before-break.

(iii) softer handoff

↳ It occurs when handoff takes place between two sectors of same B.S.

↳ carrier Frequency remains unchanged.

(iv) soft-softer handoff

↳ It occurs when handoff takes place between different sectors of two different BS.

↳ carrier frequency remains unchanged.

classmate
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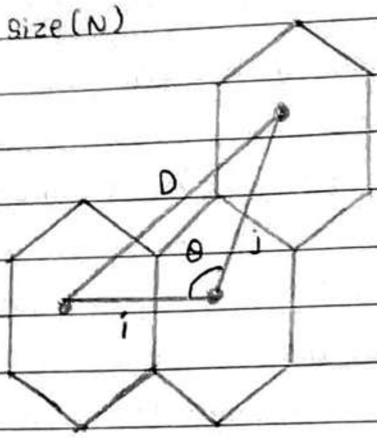
Location management
A MS is associated with a home network and it's home address resides with its home agent.
when MS moves away from its home network it enters into foreign network
The MS has to register with its home agent through the foreign agent to let the agent known its current location.
when home agent has messages to MS, it forwards them to MS via the foreign agent.
During registration process, the home agent ensures through the foreign agent that the MS is correct and can be registered in foreign network and this process is called authentication process.

Cell cluster concept
A group of cells form a cluster when the entire available spectrum is divided equally among the cells. cells in this group have a disjoint set of frequencies.
No. of cells in cluster must be determined so that the cluster can be repeated continuously within the covering area of service provider.
smaller the no. of cells per cluster, bigger will be the number of channel per cell. (capacity is increased)
However only certain cluster sizes and cell layout are possible.

$$N = i^2 + j^2 + ij$$

where i, j are non-negative integers
No. of cells per cluster N can only have values which satisfy above equation

* Derivation of cluster size (N)



Let us consider three regular hexagonal cells out of which cell A and C is cochannel cells.

D is distance between the centers of cochannel cells.

i, j is non-negative integers

Applying cosine law of triangle,

$$D^2 = i^2 - 2ij\cos\theta + j^2$$

As per cluster design rule, θ is 120°

so, $D^2 = i^2 + ij + j^2$ ---------- (i)

Here, i is a direction in which there can be a number of cells.

j is a direction 60° anticlockwise from i and along j there can be a number of cell.

If R_p is a perpendicular distance from center of a cell to its one of the side from cell geometry.

$$R_p = \frac{\sqrt{3}}{2}R$$

Then, center to center distance of two cells is,

$$2R_p = \sqrt{3}R$$

If we generalize equation (i) can be written as

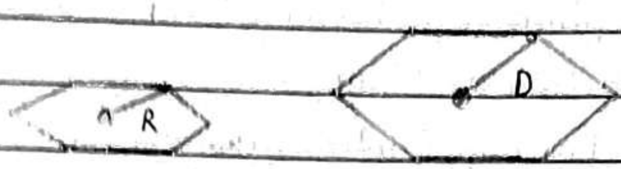
$$D^2 = (i \times 2R_p)^2 + (i \times 2R_p)(j \times 2R_p) + (j \times 2R_p)^2$$

or, $D^2 = (2R_p)^2 (i^2 + ij + j^2)$

or, $D^2 = \left(2 \times \frac{\sqrt{3}}{2}R\right)^2 (i^2 + ij + j^2)$

or, $D^2 = 3R^2(i^2 + ij + j^2)$

considering area of hexagon.



Area of hexagon with radius R

$$A_R = \frac{3\sqrt{3}}{2} R^2$$

Area of hexagon with radius D

$$A_D = \frac{3\sqrt{3}}{2} D^2$$

Taking the ratio of two expression of area we get

$$\frac{A_D}{A_R} = \frac{D^2}{R^2}$$

Also, $\frac{A_D}{A_R}$ = no. of cells enclosed inside large hexagon

From geometry of cells of large and small hexagon above

$$N + 6\left(\frac{1}{3} N\right) = \frac{D^2}{R^2}$$

or, $3N = \frac{D^2}{R^2}$

or, $3N = 3(i^2 + ij + j^2)$

$\therefore N = i^2 + ij + j^2$

* Interference and system capacity

Interference problems associated with mobile communication equipment are because of the problem of time congestion within the electromagnetic spectrum.

↳ It is the limiting factor in the performance of cellular systems.

↳ Interference can occur from clash with another mobile in the same cell or because of the call in the adjacent cells.

↳ There can be interference between the base station operating at the same frequency band or any other non cellular systems energy leaking accidently into the frequency band of cellular system.

↳ If there is an interference between the base station operating at the same frequency band or any other non cellular systems energy leaking accidently into the frequency band of cellular system.

↳ If there is an interference in the voice channels cross talks are heared and will appear as noise between the users.

↳ The interference in the control channels leads to missed and error calls because of digital signalling.

↳ More severe in urban areas because of greater RF noise and greater density of MS and BS.

- Types

(1) cochannel interference (CCI)

The cells that use the same set of frequencies are called cochannel cells and interference between the signals from that cell is called cochannel interference. It is the cross talk between two different radio transmitters using the same radio frequency

↳ The reason of CCI can be because of adverse weather conditions or poor frequency reuse.

↳ To reduce cochannel interference co-channel cells must be physically separated to provide sufficient isolation for signal propagation.

↳ If the cell size and power transmitted at BS are same then cochannel interference will become independent of the transmitted power and will depend on the radius of cell and distance between the interfering cochannel cells.

So if $\frac{D}{R}$ ratio is increased effective distance between cochannel cells will increase and interference will decrease.

For hexagonal geometry

$$Q = \frac{D}{R} = \sqrt{3N}$$

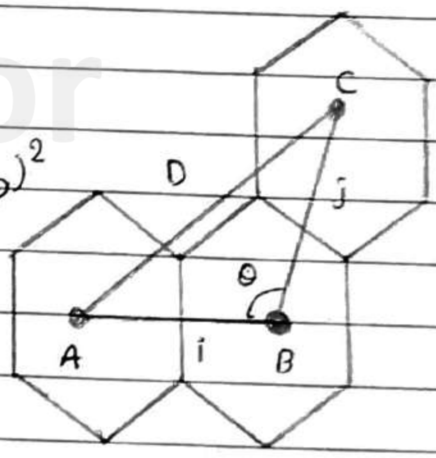
The parameter Q is frequence reuse ratio

Small Q -> ~~Larg~~ small N, increase capacity

Large Q -> Large N, decrease capacity but increase transmission quality.

- Derivation.

From geometry of regular hexagon.



$$D^2 = i^2 - 2ij\cos\theta + j^2$$

$$D^2 = (i \times 2R_p)^2 + 2(i \times 2R_p)(j \times 2R_p)\cos 120^\circ + (j \times 2R_p)^2$$

where, R_p = center to center distance of hexagonal cells

$$D^2 = 4R_p^2(i^2 + ij + j^2)$$

For regular hexagonal cells

$$R_p = \frac{\sqrt{3}}{2}R$$

Thus,

$$D^2 = 4 \times \frac{3}{4}R^2(i^2 + ij + j^2)$$

We know,

$$N = i^2 + ij + j^2$$

$$\therefore \frac{D}{R} = \sqrt{3(i^2 + ij + j^2)}$$

$$= \sqrt{3N}$$

$$\therefore Q = \frac{D}{R} = \sqrt{3N}$$

Signal-to-Interference ratio (S/I)

Consider the size of each cell is approximately same, then co channel interference is independent of transmitted power and it is the function of cell radius 'R' and the reuse distance or distance between cochannel cells 'D'. Then cochannel reuse ratio $Q = \frac{D}{R}$

For hexagonal cell,

$$Q = \frac{D}{R} = \sqrt{3N}$$

The signal of interference ratio for a mobile receiver which monitors the forward channel can be calculated as

$$\frac{S}{I} = \frac{S}{\sum_{i=1}^{i_0} I_i}$$

where,

S = desired signal power from BS

i_0 = no. of co channel interfering cells

I_i = interference power caused by ith interfering cochannel cell BS.

In order to solve this equation from power calculations we need to look into the signal power characteristics. The average power in the mobile radio channel decays as a power law of the distance of separation between Tx & Rx. The expression for the received power P_r at a distance 'd' can be approximately calculated as

$$P_r = P_0 \left(\frac{d}{d_0}\right)^{-n}$$

where,

P_0 = power received at a reference point

d_0 = small distance from transmitting antenna

n = path loss exponent (gives extent of which the signal is lost along channels)

If D_i is the distance of the ith interferer from the mobile, the received power at a given mobile due to ith interfering cell is proportional to $(D_i)^{-n}$

Let us take that the path loss exponent is same throughout the coverage area and the transmitted power be same, the S/I can be approximated as

$$\frac{S}{I} = \frac{S}{\sum_{i=1}^{i_0} I_n} = \frac{R^{-n}}{\sum_{i=1}^{i_0} D_i^{-n}}$$

$$= \frac{R^{-n}}{i_0 D_i^{-n}}$$

or, $$\frac{S}{I} = \frac{1}{i_o\left(\frac{D_i^{-n}}{R^{-n}}\right)} = \frac{(D/R)^n}{i_o} = \frac{(\sqrt{3N})^n}{i_o}$$

In hexagonal cell, there are 6 equidistant cell

So, $i_o = 6$

$$\frac{S}{I} = \frac{(\sqrt{3N})^n}{6}$$

$$N = \frac{1}{3}\left(6\frac{S}{I}\right)^{2/n}$$

The effect of CCI can be minimized by optimizing the frequency assignments of BS and their transmit powers. Tilting the BS antenna to limit the spread of the signals in the system can also be done.

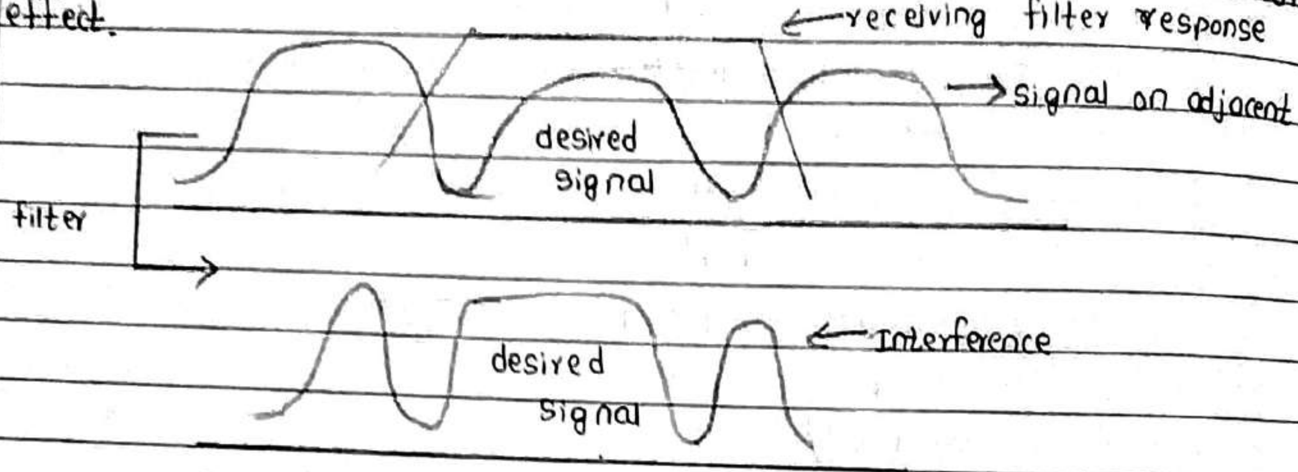
(2) Adjacent channel Interference (ACI)

Interference which is caused by adjacent channels is termed as adjacent channel interference

↳ It is signal impairment which occurs to one frequency due to presence of another signal on a nearby frequency.

↳ This occurs when imperfect receiver filters allow nearby frequencies to leak into bandpass.

↳ The problem is enhanced if the adjacent channel user is transmitting in a close range compared to the subscriber's receiver while the receiver attempts to receive a base station on the channel. This is called near-far-effect.



To reduce ACI, we should

(a) use modulation schemes which have low out-of-band radiation
(b) carefully design the bandpass filter at the receiver front end
(c) use proper channel interleaving by assigning adjacent channels to different cells
(d) avoid using adjacent channels in adjacent cells to further reduce ACI if the cell cluster size is large enough.
(e) Separate the uplink and downlink properly by TDD or FDD.
(f) Keep maximum frequency separations between each channel in a given cell.

* channel assignment strategies
(1) Fixed channel assignment strategy (FCA)
(2) Dynamic channel assignment strategy (DCA)

(1) Fixed channel Assignment strategy

↳ Each cell is allocated a predetermined set of voice channels. Any call attempt within the cell can only be served by the unused channels in that particular cell.

↳ If all the channels are occupied the cell is blocked and subscriber has to wait.

↳ This provides simple circuitry but worst channel utilization.

↳ To improve utilization, a borrowing option may be considered. With the borrowing option, a cell is allowed to borrow channels from a neighboring cell if all of its own channels are already occupied and the neighboring cells has spare channels. Borrowing is normally supervised by the MSC.

↳ MSC has full knowledge of capacity usage of cluster of cells within its jurisdiction. So, MSC is natural subsystem to oversee functions such as channel borrowing.

(2) Dynamic channel assignment strategy

↳ Voice channels are not allocated to different cells on a permanent basis. Each time a call request is made, the serving BS requests a channel from MSC.

↳ MSC determines the availability of a channel and executes its allocation procedure accordingly. The MSC only allocates a given frequency (radio channel) if that frequency is not presently in use in cell or any other which falls within minimum restricted distance of frequency reuse to avoid CCI.

↳ It reduces the likelihood of call blocking which increases the trunking capacity of system

↳ It require MSC to collect real-time data on channel occupancy, traffic distribution and radio signal quality of all channels on continuous basis.

* Call blocking and delay at cell site

In wireless communication system there are two crucial questions,

a) How successful can a new user get a connection established?

b) After connection establishment, how successfully will the connection be ~~maint~~ maintained as the user moves from one cell to another?

The first question refers to admission of new calls while the second one refers to the admission of handoff calls. The measure for both cases is given by Grade of Service (GOS).

The probability of blocking (P) for any call at any time in block call system (loss system) is given by Erlang B formula

$$P(\text{blocking}) = \frac{A^C/C!}{\sum_{i=0}^{C} A^i/i!} = \text{GOS (Grade of service)}$$

where, loss system is the system in which any attempted call is lost if any unused channels are not available, C is total no. of channel.

The second type of system is one in which a queue is provided to hold calls which were to be blocked in loss system. If a channel is unavailable immediately the call request may be delayed for specific time until when channel may be available. This system is called delay system or queuing system. The probability of non-zero delay is

$$P(\text{delay} > 0) = \frac{A^C}{A^C + C!\left(1 - \frac{A}{C}\right)\sum_{K=0}^{C-1} \frac{A^K}{K!}}$$

The propability that a call is delay for a time greater than 't' seconds is calculated as

$$P(\text{delay} > t) = P(\text{delay} > 0) \times \frac{H}{C-A}$$

* Terminologies

(1) Arrival rate (λ)

It is the average number of call requests per unit time for each user.

(2) Mean service time or holding time (H)

It is the average time of a call that occupies a channel.

(3) Traffic intensity (A)

It is a measure of channel time utilization which is average channel occupancy measured in Erlang.

$$A = \lambda H$$

It is dimensionless quality.

(4) offered traffic

It is total amount of traffic sent by user.

(5) Carried traffic

It is the total served traffic in the system. The maximum possible carried traffic is the total number of available channels in the system (C).

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* Mechanism to increase capacity of a cellular system

(1) cell splitting

↳ splitting of cell means subdividing a cell into smaller cells, each with its own BS and a corresponding reduction in height of antenna and power.

↳ with the split of cell in given area, as the size gets decreased more number of clusters are possible.

↳ This is equivalent to replication of cluster more times.

$$C = MKN$$

↳ Hence, the no. of channels available for the same geographical area gets increased i.e. capacity is increased.

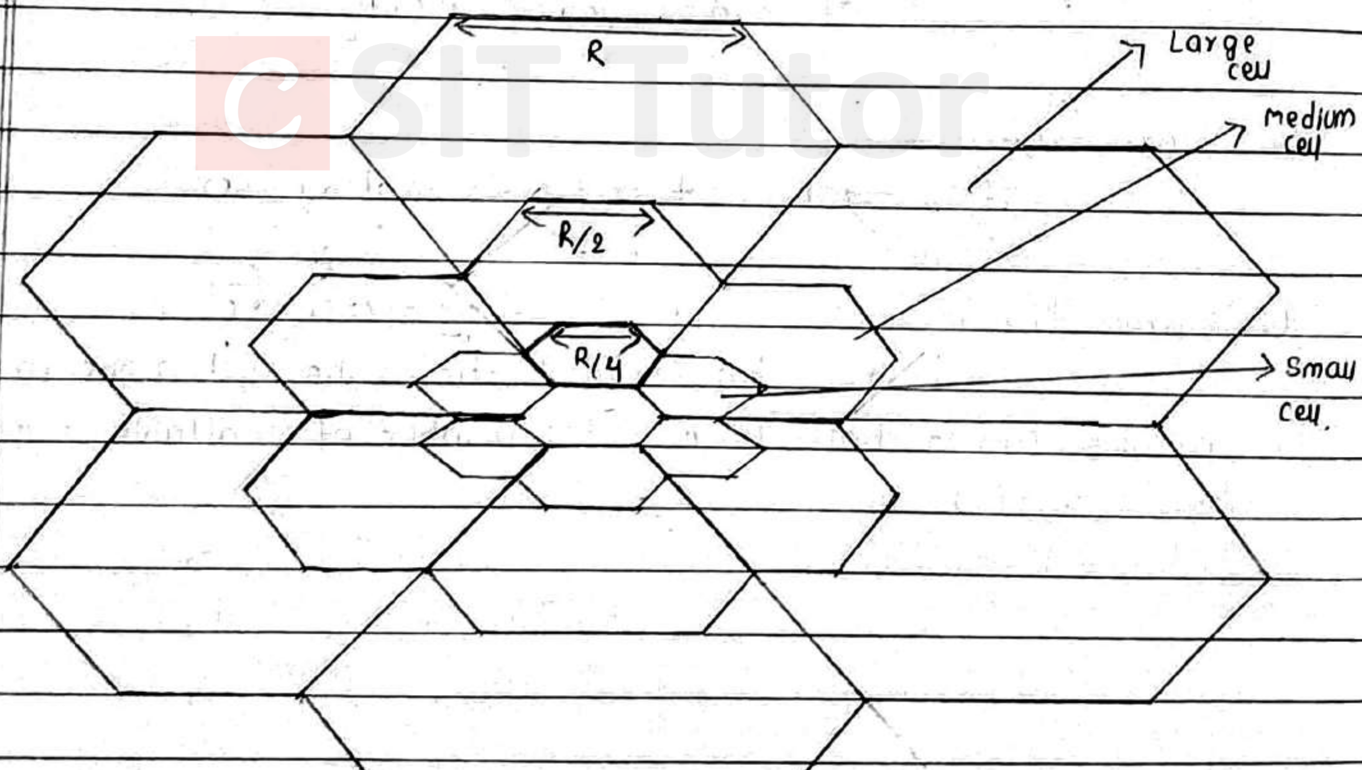


Fig:- Illustration of cell splitting from radius R to R/2 to R/4

In the figure, the original cell with radius 'R' is splitted into medium cells of radius 'R/2' and medium cells into smaller cell with radius 'R/4'. The cell splitting reduces the cell blocking probability in the area and increases the frequency of mobile handoff from cell to cell.

Let Pt_1 and Pt_2 be the power of large cell BS and small cell BS i.e cell with radius R and $R/2$ respectively. Consider received power P_r at the large cell boundary which is proportional to $Pt_1 R^{-n}$ and P_r at smaller cell boundary which is proportional to $Pt_2 (R/2)^{-n}$ on the basis of equal received power at cell boundary we have

$$Pt_1 R^{-n} = Pt_2 (R/2)^{-n}$$

$$\text{or, } \frac{Pt_1}{Pt_2} = \left(\frac{R/2}{R}\right)^{-n} = \left(\frac{1}{2}\right)^{-n}$$

$$\Rightarrow \frac{Pt_1}{Pt_2} = 2^n$$

Taking $10\log_{10}$ on both sides

$$10\log_{10}\left(\frac{Pt_1}{Pt_2}\right) = 10\log_{10} 2^n$$

$$= 10n\log_{10} 2$$

Consider the path loss exponent in worst case scenario is $n=4$

$$10\log_{10}\left(\frac{Pt_1}{Pt_2}\right) = 40\log_{10} 2 \approx 12 \text{ dB}$$

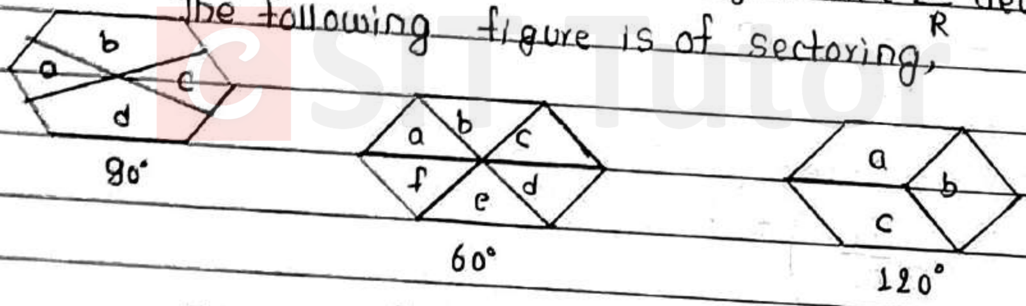
Thus splitting cell where the radius of new cell is one half of that old cell we can gain 12 dB reduction in transmit power.

(2) Cell sectoring

The cochannel interference in a cellular system may be reduced by replacing a single omni-directional antenna at the base station by several directional antennas radiating within specified sectors.

- The sectoring is done by replacing a single omni-directional antenna with 3 directional antennas (120° sectoring) or with 6 directional antennas (60° sectoring) i.e the antenna is directional.
- Each sector uses a directional antenna at BS and is assigned with a set of channels.
- The number of channels in each sector is the number of channels in a cell divided by number of sectors.
- As opposed to cell splitting where $\frac{D}{R}$ is kept constant while decreasing R, sectoring keeps R unchanged and $\frac{D}{R}$ decreases.

The following figure is of sectoring,



In case of 120° cell sectoring the no. of co-channel cells get reduced from 6 to 2 so, in using omnidirectional antenna

$$\left(\frac{S}{I}\right)_{omni} = \frac{D^n}{6}$$

In case of 120° cell sectoring

$$\left(\frac{S}{I}\right)_{directional} = \frac{D^n}{2}$$

Increase in $\frac{S}{I}$ ratio $= \frac{D^n/2}{D^n/6} = 3$

So, in case of 120° cell sectoring, the S/I ratio is increased by 3 times in comparison to that of without cell sectoring.

(3) Microcell zone concept

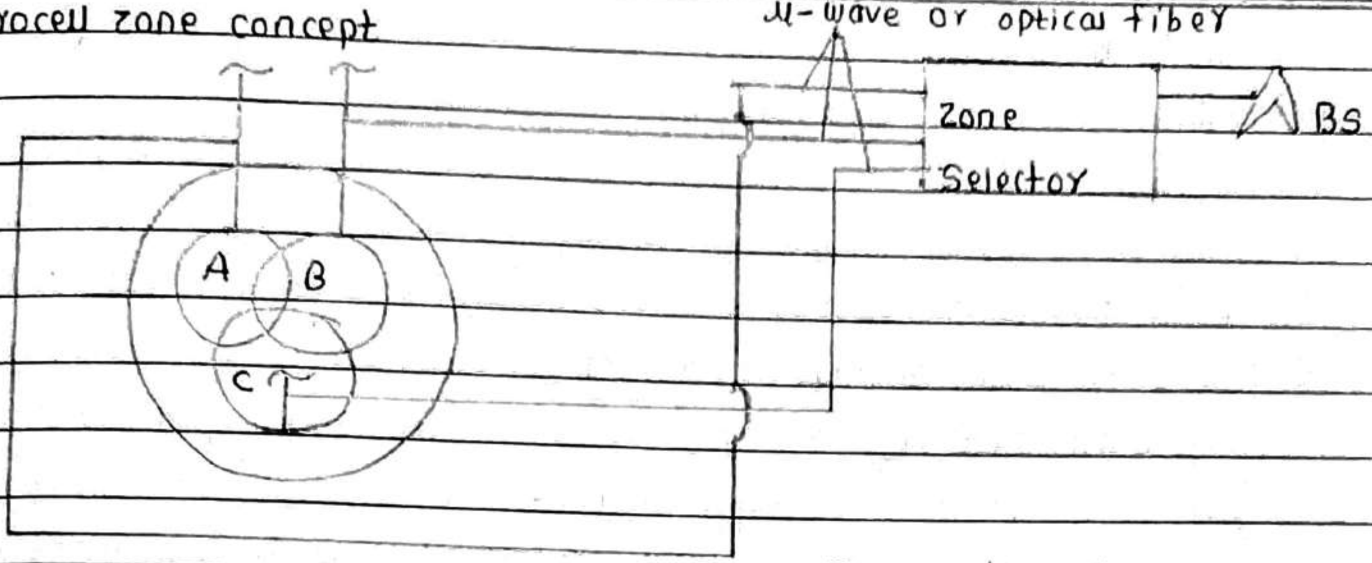


Fig:- Microcell zone

In this method the large cell is sub-divided into different zones called μ-cell. Each μ-cell has its own low power transmission which is connected to its BS through μ-wave or fiber links. A user inside a cell can continue its call through a same channel it was communicating previously on crossing from one μ-cell to another. As all channels are controlled by same BS, the same channel is retained in another μ-cell.

Every μ-cell has its own dedicated low power transmitter that relays the signal that has been assigned by BS. So, as it uses the separate transmitter the S/I ratio at any point is increased. Also as a cell can be divided into different μ-cell the no. of channels are also increased due to decrease in cluster size. As same channel is retained in the next μ-cell, there occurs no overhead due to handoff. The BS just switches the original channel to next transmitter.

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Multiple Access Technologies

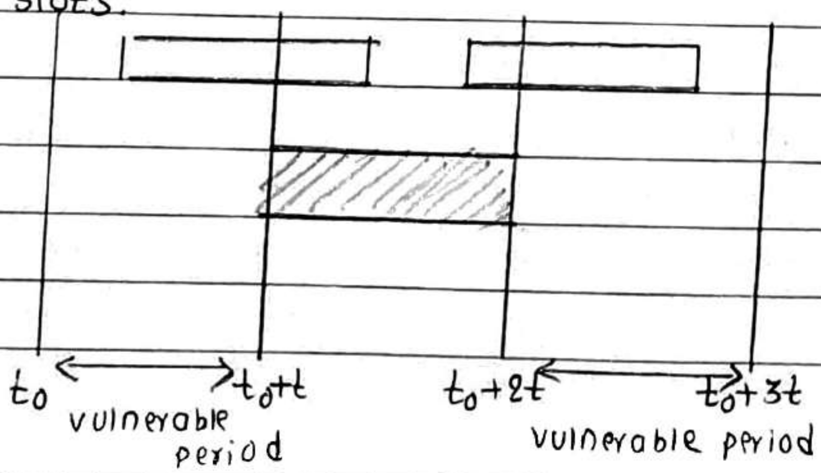
Multiple access is a single transmission techniques in which two or more user transmit the data or information to each other using the same propagation channel. In order to avoid collision between the data sent by multiple users on the same channel some methods are used which are called conflict free multiple access. However there are some methods to access the same channel by the number of users in an uncoordinated way that cause the great chance of collision of data. The access method that have probability of data collision are called random access methods. There are two types of random access technologies

(1) ALOHA

(2) Carrier Sense Multiple Access (CSMA)

(1) ALOHA

The time required to transmit one packet is called slot. In ALOHA when transmission from two or more users overlap they destroy each other, whether the time is complete or partial overlap. The maximum interval over which two packets can overlap and destroy each other is called vulnerable period. The mode of random access in which users can transmit at any time is called pure ALOHA. In pure ALOHA system where the length of packet is fixed, the vulnerable period of two times slots.



A version of ALOHA in which users are restricted to transmit only from corresponding slot ... each referred to as slotted ALOHA. So as the packets can be imposed on the channel at slot boundary there is a chance of collison of data due to complete overlap of the packet of two or more users. The vulnerable period in slotted ALOHA is thus one slot time only. Due to this maximum throughput rate of slotted ALOHA is double to that of pure ALOHA.

- Throughput of ALOHA system

Consider G be the mean number of traffic attended by the user within one time slot. Now, if we consider the traffic imposed on channel has poisson distribution then the probability of K number of traffic to be generated within slot can be expressed as

$$P_r(K) = \frac{e^{-G} G^K}{K!}$$

So for the non overlap of the packet there should be no other packet generated (ie. K=0)

$P_r(0) = e^{-G} = P$ (no other packet generation within specific slot)

let, T be vulnerable period time, then

$P_r(0) = e^{-TG} = P$ (no other packet generation within vulnerable time)

Now,

consider 'S' be the throughput, defined as successful transmitted traffic load in the given specific slot, then propability of successful transmission is expressed as

$P[\text{success}] = \frac{S}{G} = P$ (no other packets generation during the specific slot)

we can write,

$$\frac{S}{G} = e^{-TG} \Rightarrow S = G e^{-TG}$$

Now, for maximum value of throughput

$$\frac{dS_{max}}{dG} = G(-T)e^{-TG} + e^{-TG} = 0$$

$$\Rightarrow e^{-TG}(1-TG) = 0$$

$$\Rightarrow 1-TG = 0$$

$$\therefore G = \frac{1}{T}$$

So

when $G = \frac{1}{T}$; $S = S_{max}$

$$S_{max} = \frac{1}{T} e^{-T \times 1/T} = \frac{1}{Te}$$

$$\therefore S_{max} = \frac{1}{Te}$$

For pure ALOHA Vulnerable time (T) = 2 time slot and for slotted ALOHA, T = 1 time slot so,

$$(S_{max})_{pure} = \frac{1}{2e} = 0.184$$

$$(S_{max})_{slotted} = \frac{1}{e} = 0.368$$

(2) CSMA

In contradiction to that of ALOHA this random access technology involves the sensing of channel by user before transmission.

CSMA is a media access control protocol in which a node verifies the absence of other traffic before transmitting on a shared transmission medium, such as electrical bus or a band of electromagnetic spectrum.

A transmitter attempts to determine whether another transmission is in progress before initiating a transmission using a carrier sense mechanism i.e. it tries to detect the presence of a carrier signal from another node before attempting to transmit.

Types

(i) Non-persistant CSMA

In this CSMA, the transmitter first senses the channel, when it is ready to transmit, the transmitter sense the packet if the channel is idle otherwise waits for a random amount of time and again sends the packet.

(ii) 1-peristant CSMA

In this CSMA, the transmitter first senses the channel when it is ready to send and if the channel is idle the packet is sent but if the channel is busy it again senses the channel after a certain period of time and sends its packet only when channel is idle.

* conflic-free multiple access

* FDMA (Frequency Division Multiple Access)

Here, the total bandwidth is divided into non-overlapping frequency subbands. Each user is allocated a unique frequency subband for the duration of the connection, whether the connection is in active or idle state. In the frequency domain the signal from each user is transmitted through different bandpass filter. In order to avoid inteference between the uplink and downlink channels, the frequency allocated is to be separated by a sufficient amount and both channel should operat at different frequency band. This division of transmission channel into downlink and uplink channels called frequency division duplexing

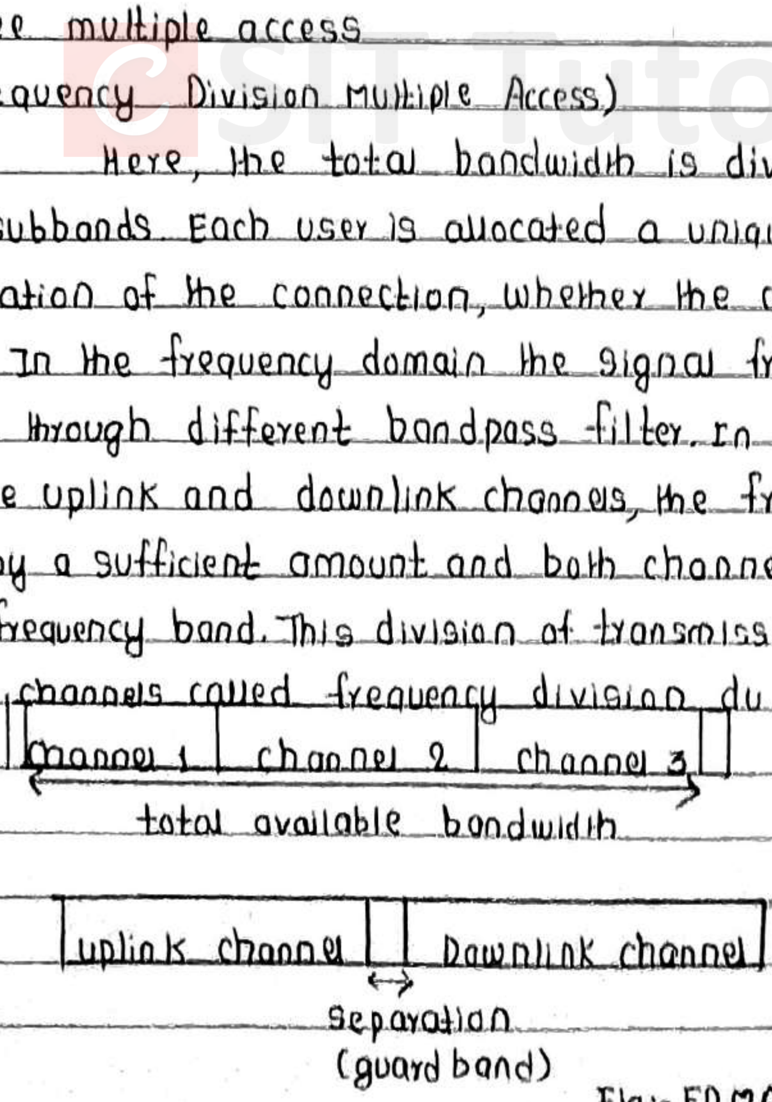


Fig:- FDMA

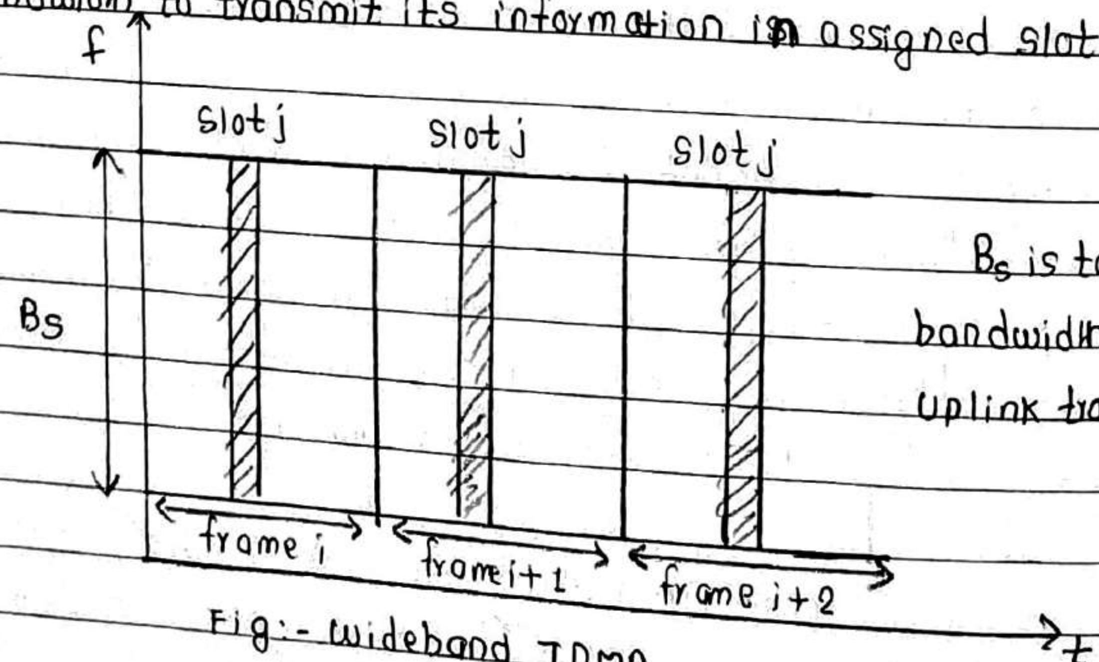
* Time Division multiple Access (TDMA)

Here, the overall available channel time is partitioned into frames and the length of frame is long enough so that every user in services has an oppurtunity to transmit once per frame. To achieve this, a TDMA frame is further partitioned into time slots where user has to transmit their data in their assigned slots The assigment of slot can be fixed or dynamic. If the assigned slot is fixed from frame to frame for the duration of connection, the user have to synchronize to their assigned ~~sett~~ slots and is called synchronous TDMA. If the users can transmit data through any one of the slot in a frame, only after it has the packet ready to send, then such type of TDMA is asynchronous TDMA.

• STDMA

Here, the channel time is divided into contiguous slot, each of which is long enough to transmit or receive one information unit. STDMA operates on a frame-by-frame basis. Depending on the manner in which frequency is allocated, STDMA can be wideband or narrowband.

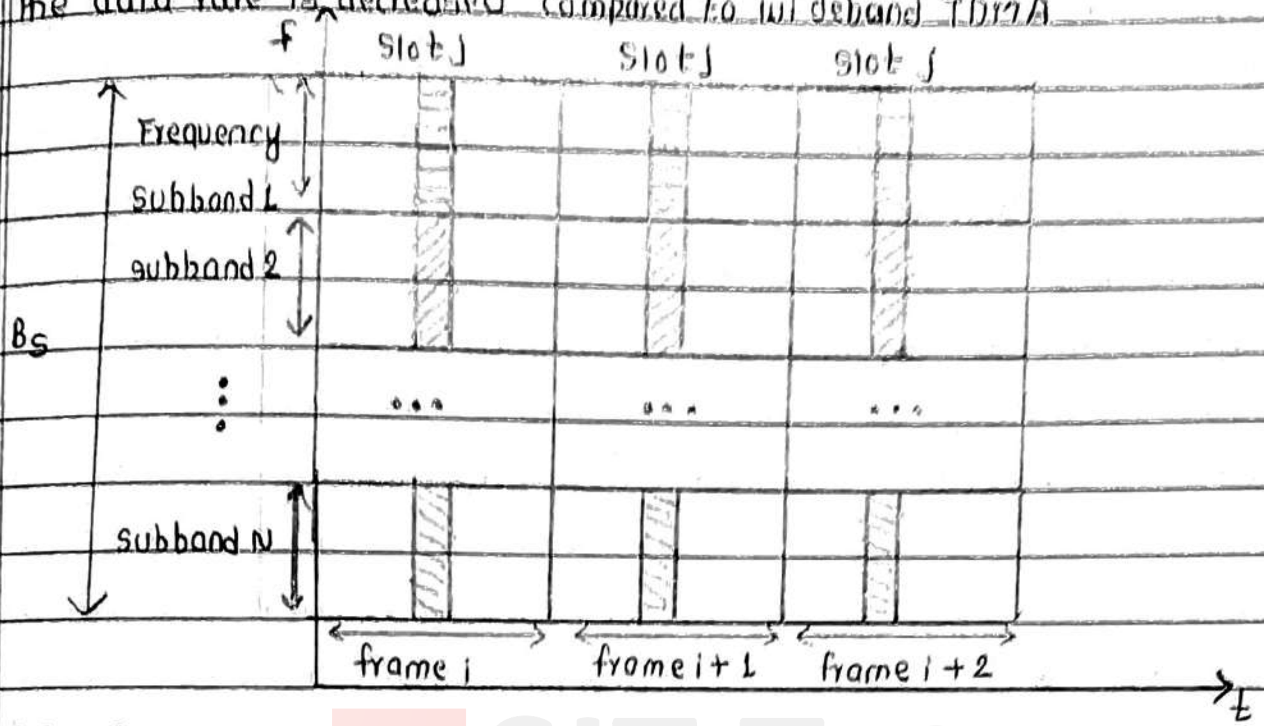
It is wideband TDMA if the channel time is divided into slots and an individual user is allowed to use the entire available channel bandwidth to transmit its information in assigned slot.



B_s is total frequency bandwidth allocated to uplink transmission.

Fig:- Wideband TDMA

It is narrowband TDMA if the overall bandwidth is first divided into frequency bands and the channel time corresponding to each frequency band is divided into time slots for packet transmission. Here, the data rate is decreased compared to wideband TDMA.



- ATDMA

It is implemented using reservation access mechanism. It's frame contains two segments

↳ The leading segment that contains mini request slots for active users to submit requests for slot allocation.

↳ The trailing segment that contains information slots for packet transmission.

Request submitted in request slot in current uplink frame will be received by BS. Based on information contain in request vector, BS assigns information slots to requesting users in next downlink frame. So, request submitted in current frame will be accomodated in next frame.

To structure ATDMA frame one can allocate a request slot to each user on permanent basis. If there are N users the request segment has to contain N request slots.

* Code Division Multiple Access (CDMA)

CDMA is a spread spectrum multiple access method, which the bandwidth of baseband information carrying signal from different users is spreaded by different signals with a bandwidth much larger in comparison to that of baseband signal. These extra spreading signal used for different users are orthogonal to each other and have noise like behavior that are generated by a device called pseudo random generator and spreading signal is called pseudo random sequence or PN sequence. While this system is very bandwidth inefficient in case of single user, the advantage of spread spectrum is that many users can simultaneously use the same bandwidth without significantly interfering with one another. In case of multiuser scenario, the system is very bandwidth efficient and is the basic technology used in CDMA. The spreaded signals are demodulated at the receiver to remove PN sequence and extract the original message.

- Direct sequence spread spectrum (DS-SS)

A direct sequence spread spectrum system spreads the baseband data by directly multiplying the baseband data pulses with a pseudo noise sequence that is generated by the pseudo noise code generator. A single pulse or symbol of PN waveform is called chip. Following are the diagram of DS-SS system transmitter and receiver.

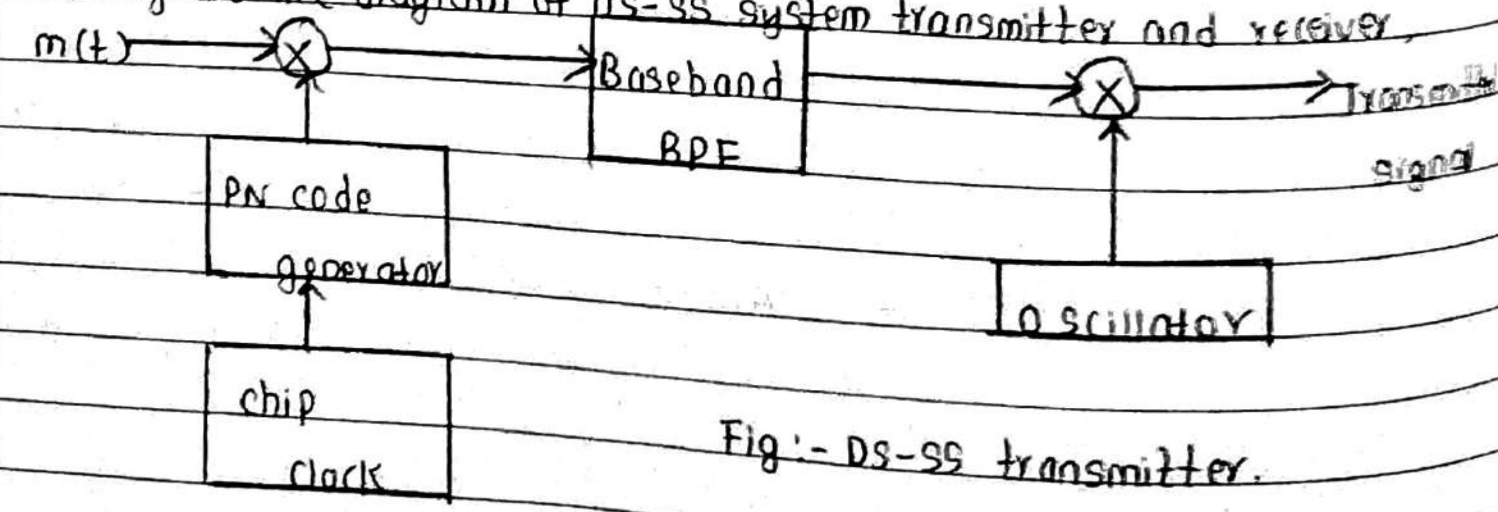


Fig:- DS-SS transmitter.

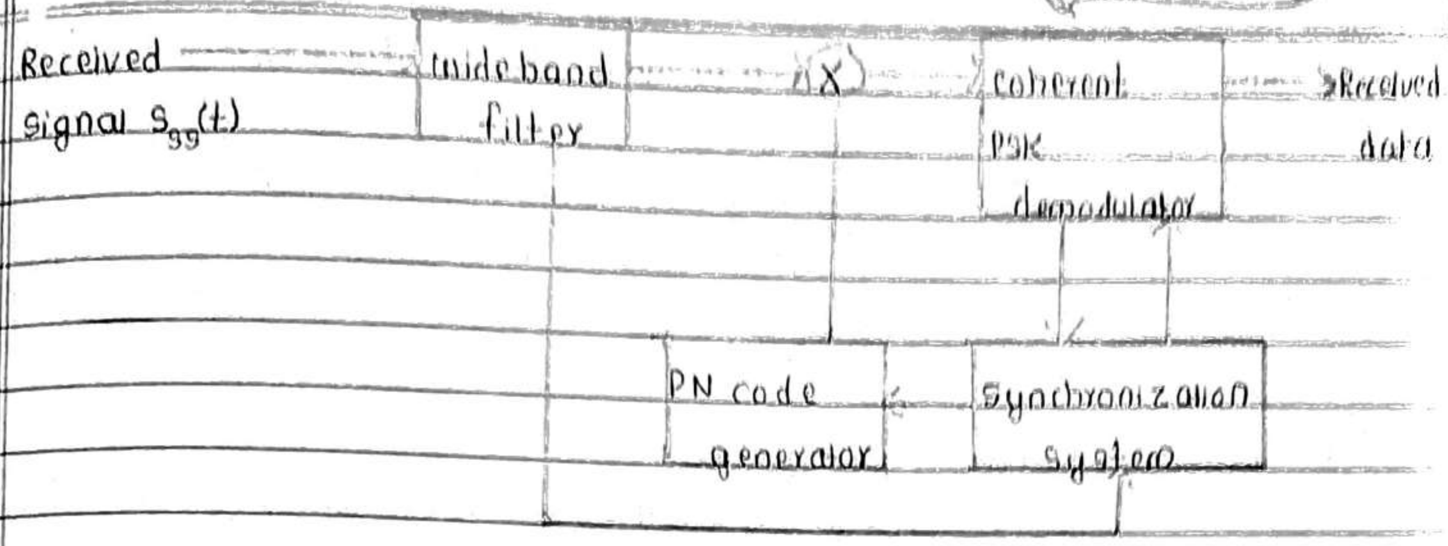


Fig:- DS-SS Receiver.

* spectral efficiency

Spectral effeciency is the measurement of how datas are transmitted through the available bandwidth and with what rate. The spectral efficiency (η) can mathematically defined as

$$\eta = \frac{\text{total no. of channels available for data in system}}{(\text{system bandwidth}) \times (\text{total coverage area})}$$

unit of η = channels/ MHz Km^2

It is also expressed as

$$\eta = \frac{\text{total traffic carried by system}}{(\text{system bandwidth}) \times (\text{total coverage area})}$$

- spectral efficiency of FDMA

If B_s is total available bandwidth, B_g is guard bandwidth of each channel then total no. of available channels will be

$$N_s = \frac{B_s - 2B_g}{B_c} \Rightarrow B_s = N_s B_c + 2B_g$$

~~where~~,

Let N_{ctc} be no. of allocated control channel and N_{data} be no. of data channels then,

$$N_s = N_{ctc} + N_{data}$$

so, $B_s = N_{ctc} B_c + N_{data} B_c + 2B_g$

where, $N_{data} B_C$ is total B.W available for data transmission Now, spectral efficient of FDMA is

$$\eta_{FDMA} = \frac{\text{B.W available for data transmission}}{\text{system B.W}}$$

$$= \frac{N_{data} * B_C}{B_S}$$

* Spectral efficiency of system

consider N_S be total available channels. In case of frequency reuse ~~ratio~~ scenario this N_S is distributed in cluser of size N to increase the system capacity so, total number of channels per cluster is

$$N_S = N_{ch/cluster} = \frac{B_S - 2B_g}{B_C}$$

Now,

If $N_{data/cluster}$ & $N_{ch/cluster}$ represents the total no. of data and control channel respectively in cluster

$$N_{ch/cluster} = N_{data/cluster} + N_{ctc/cluster}$$

So,

$$N_{data/cluster} = \frac{B_S - 2B_g}{B_C}$$

Now, spectral efficiency is

$$\eta = \frac{\text{total number of data channels in cluster}}{\text{system B.W} \times \text{Area of cluster}}.$$

Also note that in any multiple access system there is a finite probability of blocking of some traffic so we have to define a factor called trunk efficience (η_t) which is a function of blocking probability and total no. of available channels, then efficiency is

$$\eta = \frac{\eta_t \times N_{data/cluster}}{B_S \times A_{cluster}} \quad \text{Erlangs/MHz/km}^2$$

* Spectral efficiency of WTDMA

It is defined as the percentage of time required to transmit information data symbol in each frame

Let, T_p = time duration for preamble

T_t = time duration for trailer

T_f = frame duration

L_d = no of data symbols in each slot

L_s = total no. of data symbols

spectral efficiency $(\eta_{WTDMA}) = \frac{T_f - T_t - T_p}{T_f} \times \frac{L_d}{L_s}$

* Spectral efficiency of NTDMA

Let, B_c = Bandwidth of individual user (sub-band)

N_u = Number of subbands

B_g = Guard band / spacing

No. of subbands $(N_u) = \frac{B_s - 2B_g}{B_c}$

η_{NTDMA} is directly proportional to η_{WTDMA} and proportionality constant which is $\alpha = \frac{N_u B_c}{B_s}$

So, $\eta_{NTDMA} = \eta_{WDTMA} \times \alpha$

$= \frac{T_f - T_t - T_p}{T_f} \times \frac{L_d}{L_s} \times \frac{N_u B_c}{B_s}$

$\therefore \eta_{NTDMA} = \frac{T_f - T_t - T_p}{T_f} \times \frac{B_s - 2B_g}{B_s} \times \frac{L_d}{L_s}$

* cell capacity of TDMA

Cell capacity is defined as maximum number of mobile users that can be supported simultaneously in each cell.

Total no. of users in system.

$$N_s = N_u \times N_{slot} \quad \text{where, } N_u = \begin{cases} 1 & \text{for WTDMA} \\ \frac{B_s - 2B_g}{B_c} > 1 & \text{for NTDMA} \end{cases}$$

So, the total no. of TDMA channel per cell with frequency reuse factor 'N' will be $N_c = \frac{N_u \times N_{slot}}{N}$ is capacity of cellular system.

Usually user can't send data in its slot and hence there is chance of empty slot in each frame. Let 'S_f' denote source activity factor defined as percent time that a connected mobile user is actually generating information data for transmission. cell capacity is

$$N_c = \frac{N_u \times N_{slot}}{N \times S_f}$$

value of S_f is taken as finite value less than 1 in ATDMA and 1 is STDMA.

* DS-SS

Received spread sprectrum for single user is

$$s_{ss}(t) = \sqrt{\frac{2E_s}{T_s}}\, m(t)\, p(t) \cos(2\pi f_c t + \theta)$$

$m(t)$ = data sequence

$p(t)$ = PN spreading sequence

since, $p(t) = \pm 1$, $p[(t)]^2 = 1$

output of multiplier $= \sqrt{\frac{2E_s}{T_s}}\, m(t) \cos(2\pi f_c t + \theta)$ and in DS-SS message signal is digitalized $m(t) = \pm 1$

so output $= \pm \sqrt{\frac{2E_s}{T_s}} \cos(2\pi f_c t + \theta)$

which is BPSK signal so, it is denomulated using BPSK demodulator.

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Mobility management in wireless networks.

* call Admission control (CAC)

The total capacity of any cell is used to handle the ongoing connections and to admit ~~new~~ new call request along with some handoff request. The mechanism that supervises the admission of new and handoff call should protect the integrity of network and satisfy the user QoS requirement, this mechanism is refferred to as call admission control (CAC).

CAC ensures the network integrity by restricting the access to the network so as to avoid overload and congestion, suppose that N-1 calls are ongoing in a cell having 'N' no. of channels within CAC mechanism when a new request (either new call / handoff) is encountered it calculates the amount of resources first. If there are enough resources to admit the n^{th} call such that its QoS requirement and those of ongoing calls are satisfied then the new call is admitted. As it is more important to maintain the continuity of ongoing call rather than the admission of new calls the priority is given to handoff calls to new call request of such situation arises.

Let P_n be new call blocking probability and P_h be the handoff call dropping probability. The GoS can be defined by jointly taking both P_h and P_n as

$$GoS = P_n + \alpha P_h$$

Where $\alpha > 1$ is a balancing factor for higher probability.

* Prioritized call admission

The handoffs are given more priority than new call for the management of admission of request based on priority, the certain no. of channels are already reserved for handoff calls. This concept is guard channel concept of CAC.

consider N_g be total reserved no. of guard channels and N_{UU} be total no. of unallocated channels in a cell; then prioritized call admission rule is following

↳ If $N_{UU} > N_g$ admit new or handoff request

↳ If $N_{UU} \leq N_g$ admit handoff request only

* Handoff strategies

MSC collects all the information concerning the channel occupancy status of all the base stations of every cluster. When the mobile moves between cells in the cluster overseen by same MSC, handoff initiation and execution are handled by MSC and there is no need to copy connection identification states. This mode of operation is "intraswitch handoff."

When the mobile moves from one cell to another cell overseen by another MSC connection identification states has to be transferred from the current serving MSC to new MSC. This mode of operation is "interswitch handoff."

Depending on the information used and action taken to initiate the handoff, the methods of handoff are

(i) Mobile controlled handoff (MCHO)

Here all the handoff related works are done by mobile stations. It reduces burden to the network. However this will increase the complexity of mobile terminal.

(ii) Network controlled handoff (NCHO)

Here, BS or APs monitor the signal quality information from mobile and report it to the MSC. MSC then monitor the reliable BS for handoff.

(iii) Mobile Assisted handoff (MAHO)

Here, the mobile terminal itself collects the signal quality information from serving and surrounding base station and transform it to the MSC through serving BS.