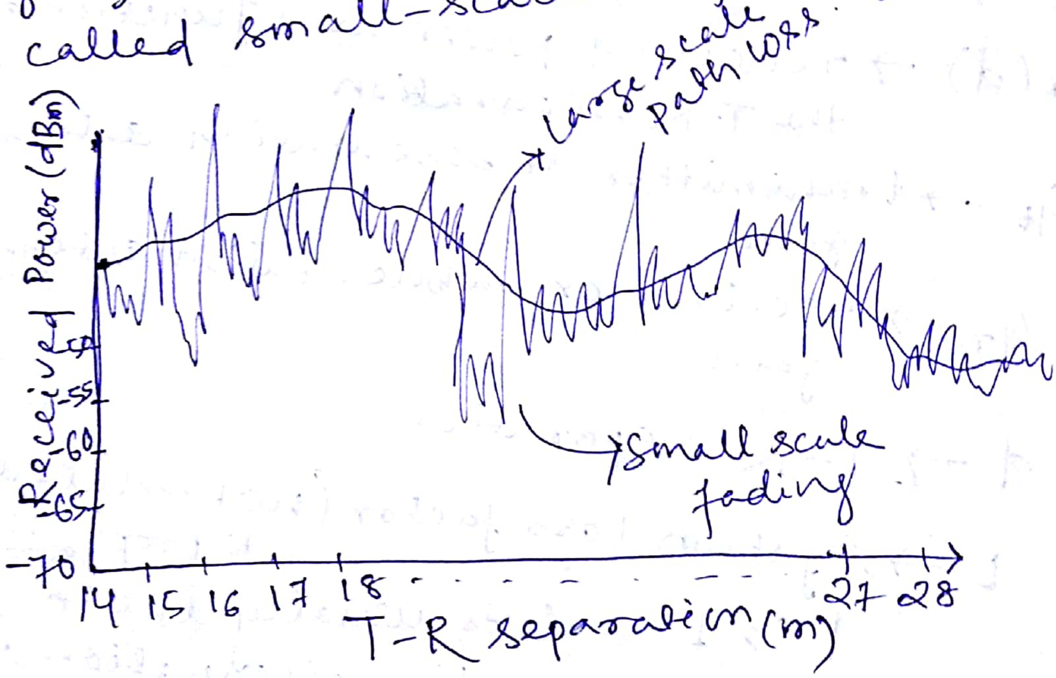
Large Scale path Loss

→ Propagation models that predict the mean signal strength for an arbitary (large) T-R (transmitter-receiver) separation distance are called the large-scale propagation models. These are useful in estimating the radio coverage area of a transmitter.

→ Where, the propagation models that characterize the rapid fluctuations of the received signal strength over very short distances (a few wave-lengths) or short time durations (few seconds) are called small-scale or fading models.



Small scale fading → rapid fluctuations
Large scale path loss → more gradual variations.

Free Space Propagation model

→ The free space propagation model is used to predict received signal strength when the Tx and Rx have a clear, unobstructed line-of-sight path between them.

→ Satellite communication systems and microwave line-of-sight radio links undergo free space propagation.

→ The free space power received by a receiver antenna which is separated from a radiating transmitter antenna by a distance, d, is given by

$$P_r(d) = \frac{P_t \cdot G_b \cdot G_m}{L}\left(\frac{\lambda}{4\pi d}\right)^2$$

$$\text{or } \frac{P_t \cdot G_t \cdot G_r \cdot \lambda^2}{(4\pi)^2 d^2 L}$$ → Friis free space equation.

P_t → transmitted power

$P_r(d)$ → received power as a function of the T-R separation.

G_b or G_t → transmitter or base station antenna gain.

G_m or G_r → receiver or mobile station antenna gain.

d → T-R separation.

L → system loss factor (not related to propagation)

$L \geq 1$ and is usually due to transmission line attenuation, filter losses, antenna losses.

$L = 1$ → no loss in the system hardware.

λ → wavelength in meters.

→ The gain of antenna is related to its effective aperature, A_e, by

$$G = \frac{4\pi A_e}{\lambda^2}$$

$$\lambda = \frac{c}{f} = \frac{2\pi c}{\omega_c}$$

$f \rightarrow$ carrier freq. in Hz

$\omega_c \rightarrow$ carrier freq. in radians/sec.

→ From the Friis free space equation, it is ~~pr~~ known that the received power decays with distance at a rate of 20 dB/decade.

→ The antenna gains are usually referenced with an isotropic radiator which have unity gain. The isotropic radiator radiates same power in all directions (G_t or $G_r = 1$). The effective isotropic radiated power (EIRP) is defined as

$$\boxed{EIRP = P_t \cdot G_t \text{ or } P_t \cdot G_t}.$$

It represents the maximum radiated power available from a transmitter in the direction of maximum antenna gain, as compared to an isotropic radiator.

→ The path loss, which represents signal attenuation ~~as~~ as a positive quantity measured in dB, is defined as the difference (in dB) between the effective transmitted power and the received power. This may or may not include the effect of the antenna gains.

→ The path loss for the free space model when antenna gains are included is given by

$$PL(dB) = 10 \log \frac{P_t}{P_r} = -10 \log \left[\frac{G_t \cdot G_r \lambda^2}{(4\pi)^2 d^2} \right]$$

If the antennas are assumed to have unity gain,

Q. If a transmitter produces 50 W of power, express the transmitted power in units of (a) dBm (b) dBW. If 50W is applied to a unity gain antenna with a 900 MHz carrier frequency, find the received power in dBm at a free space distance of 100m from the antenna. What is $P_r(10\,km)$? Assume unity gain for the receiver antenna.

$$\lambda = \frac{c}{f} = \frac{3\times10^8}{900\times10^6} = \frac{1}{3}\,m.$$

$P_t = 50\,W$

$$P_t(dBW) = 10\log_{10}\left[\frac{P_t}{1\,W}\right] = 10\log\left[50/1\right]$$

$$= 17\,dBW.$$

$$P_t(dBm) = 10\log\left[P_t(mW)/1\,mW\right].$$

$$= 10\log\left[P_t(W)/0.001\,W\right]$$

$$= 10\log\left[50\times10^3\right] = 47.0\,dBm.$$

$$P_r(100m) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L} = \frac{50\times1\times1\times \frac{1}{3}^2}{(4\times3.14)^2\times100^2\times1}$$

$$= 3.5\times10^{-6}\,W = 3.5\times10^{-3}\,mW.$$

$$P_r(dBm) = 10\log\left(3.5\times10^{-3}_{mW}\right) = -24.5\,dBm.$$

Ans. The received power at 10km can be expressed in terms of dBm where $d_0 = 100\,m$ and $d = 10\,km$

$$P_r(10Km) = P_r(100) + 20\log\left[\frac{100}{10000}\right].$$

$$= -24.5\,dBm - 40$$

$$= -64.5\,dBm.$$

The three basic Propagation Mechanisms-

→ The signal propagation in mobile communication systems is mostly affected by reflection, diffraction and scattering.

→ Reflection occurs when a propagating electro-magnetic wave impinges upon an object which has very large dimensions when compared to the wavelength of the propagating wave.
e.g. earth surface, buildings, walls.

→ Diffraction occurs when the radio path betn the Tx and Rx is obstructed by a surface that has sharp irregularities (edges). It causes bending of waves around the obstacle. At high frequencies diffraction depends on the geometry of the object as well as the amplitude, phase and polarization of the incident wave at the point of diffraction.

→ Scattering occurs when the propagation medium consists of objects with dimensions that are small compared to the signal wavelength.
e.g. Street signs, lamp posts, foliage induce scattering in mobile comm. system.

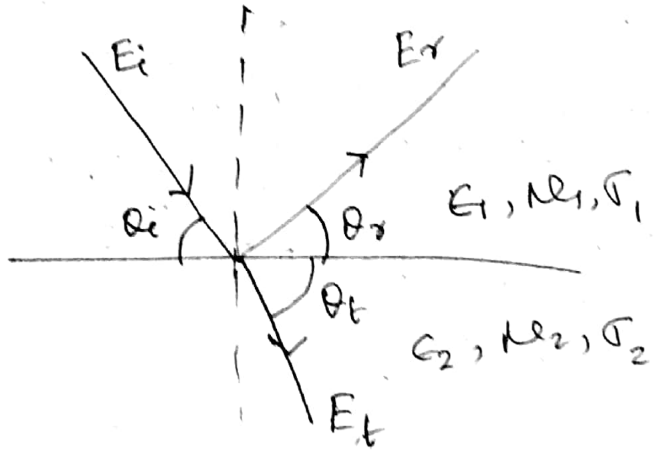
Reflection

→ When a radio wave propagating in one medium impinges upon another medium having different electrical properties, the wave is partially reflected and partially transmitted

→ If the wave is incident on a perfect dielectric, part of the energy is transmitted into the second medium and part of the energy is reflected back into the first medium

→ If the second medium is a perfect conductor, then all incident energy is reflected back into the first medium without loss of energy.

→ The electric field intensity of the reflected and transmitted waves may be related to the incident wave through the Fresnel reflection co-efficient (Γ).
This is a function of the material properties and generally depends on the wave polarization, angle of incidence and the frequency of the propagating wave.



$$\theta_i = \theta_r$$

$$E_r = \Gamma E_i$$

$$E_t = (1+\Gamma) E_i$$

Γ is either $\Gamma_{\parallel}$ or $\Gamma_{\perp}$, depending on ~~the~~ whether the E field is in (vertical) ~~and~~ or normal (horizontal) to the plane of incidence.

Brewster Angle

The Brewster angle is the angle at which no reflection occurs in the medium of origin. It occurs when the incident angle θ_B is such that the reflection co-efficient $\Gamma_{\parallel} = 0$. The Brewster angle is given by the value of θ_B which satisfies

$$\sin(\theta_B) = \sqrt{\frac{\varepsilon_1}{\varepsilon_1 + \varepsilon_2}}$$

If the first medium is free space and the second medium has a relative permittivity ε_r, then

$$\sin(\theta_B) = \frac{\sqrt{\varepsilon_r - 1}}{\sqrt{\varepsilon_r^2 - 1}}$$

Brewster angle occurs only for parallel polarization

Electric field is in the plane of incidence → vertical polarization and $\Gamma = \Gamma_{\parallel}$

Electric field is normal to the plane of incidence → horizontal polarization
$\Gamma = \Gamma_{\perp}$

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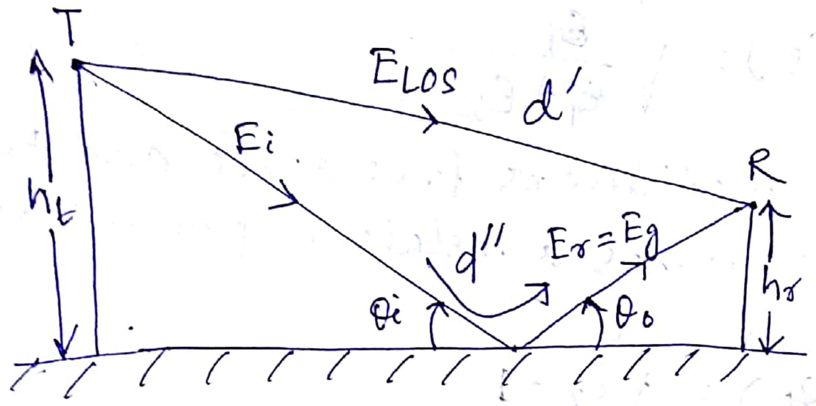
Ground Reflection (Two-Ray) Model

→ The two-ray model considers both the direct path and a ground reflected propagation path between transmitter and receiver.



→ In most mobile communication systems the max^m T–R separation is at most only a few tens of kms and the earth can be assumed to be flat.

→ The total received E-field, E_{TOT} is

$$E_{TOT} = E_{LOS} + E_g.$$

$h_t \rightarrow$ height of the transmitter

$h_r \rightarrow$ height of the receiver.

If E_0 is the free-space E-field (V/m) at a reference distance d_0 from the Tx, then for $d > d_0$, the free space propagating E-field is given by

$$E(d,t) = \frac{E_0 d_0}{d} \cos\left(\omega_c\left(t - \frac{d}{c}\right)\right) \quad (d > d_0)$$

where $E(d,t) = E_0 d_0 / d$ represents the envelope of the E-field at d meters from the transmitter.

Two propagating waves arrive at the receiver: the direct wave that travels a distance d' and the reflected wave that travelled a distance d''. The E-field due to these are given as,

$$E_{LOS}(d', t) = \frac{E_0 d_0}{d'} \cos\left(w_c\left(t - \frac{d'}{c}\right)\right)$$

and the E-field for the ground reflected wave, which has a propagation distance of d'', can be expressed as

$$E_g(d'', t) = \Gamma \cdot \frac{E_0 d_0}{d''} \cos\left(w_c\left(t - \frac{d''}{c}\right)\right)$$

According to laws of reflection,

$$\theta_i = \theta_0$$

$$E_g = \Gamma E_i$$

$$E_t = (1+\Gamma) E_i$$

where Γ is the reflection co-efficient for ground. For small values of θ_i, the reflected wave is equal in magnitude and 180° out of phase with the incident wave.

-> Assuming perfect horizontal polarization (i.e. the E-field is parallel to the reflecting surface and normal to the plane of incidence), and ($\Gamma = -1$ and $E_t = 0$) the resultant E-field ground reflection model, is given as:

$$\cancel{|E_{TOT}| = |E_{LOS} + E_g|}$$

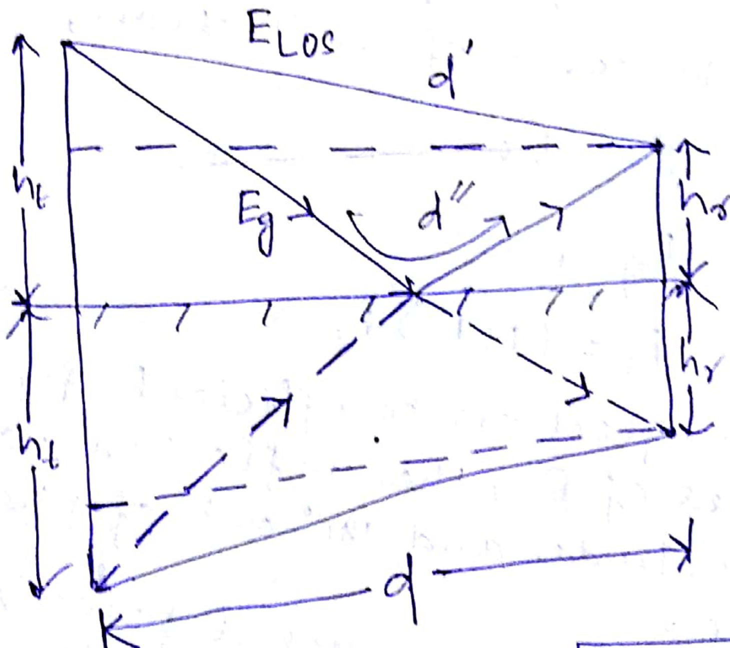
The resultant E-field is,

$$|E_{TOT}| = |E_{LOS} + E_g|$$

$$\Rightarrow \dot{E}_{TOT}(d,t) = \frac{E_0 d_0}{d'} \cos\left(\omega_c\left(t - \frac{d'}{c}\right)\right) + (-1)\frac{E_0 d_0}{d''} \cos\left(\omega_c\left(t - \frac{d''}{c}\right)\right)$$

Phase difference and delay between the two components:

The path difference, $\Delta = d'' - d'$, is determined from the method of images.



$$\Delta = \underbrace{\sqrt{(h_t + h_r)^2 + d^2}}_{d''} - \underbrace{\sqrt{(h_t - h_r)^2 + d^2}}_{d'}$$

if $d >> h_t + h_r$, using Taylor series ~~apporix~~ approximation, the path difference can be simplified as

$$\Delta = d'' - d' \approx \frac{2 h_t h_r}{d}$$

The phase difference,

$$\theta_\Delta = \frac{2\pi\Delta}{\lambda} = \frac{\Delta \cdot w_c}{c} \qquad \lambda = \frac{c}{f} = \frac{c \cdot 2\pi}{w}$$

and time delay, $\tau_d = \frac{\Delta}{c} = \frac{\theta_\Delta}{w_c} = \frac{\theta_\Delta}{2\pi f_c}$

As d becomes large $\Rightarrow \Delta = d'' - d'$ becomes small.

So the amplitudes of E_{LOS} and E_g are nearly identical and differs only in phase.

So $\left|\frac{E_0 d_0}{d}\right| \approx \left|\frac{E_0 d_0}{d'}\right| \approx \left|\frac{E_0 d_0}{d''}\right|$

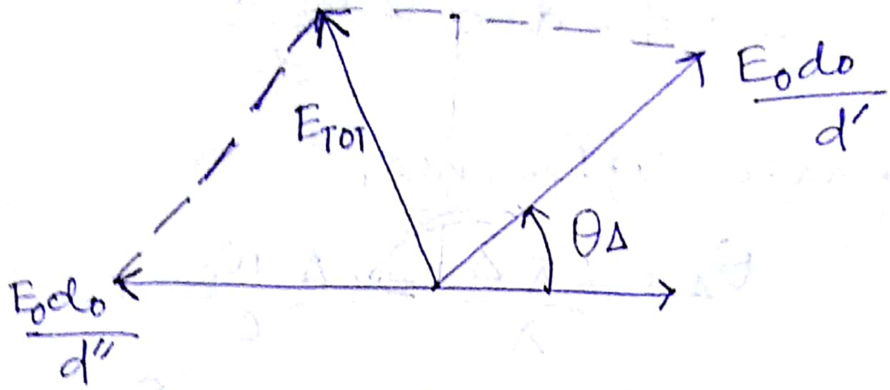
If we calculate the E field at the time when, the reflected wave arrives at the receiver $(t = d''/c)$

$$E_{TOT}\left(d, t = \frac{d''}{c}\right) = \frac{E_0 d_0}{d'} \cos\left(w_c\left(\frac{d'' - d'}{c}\right)\right) - \frac{E_0 d_0}{d''} \cos 0^\circ$$

$$\approx \frac{E_0 d_0}{d}\left[\cos\left(\frac{\Delta w_c}{c}\right) - 1\right]$$

$$= \frac{E_0 d_0}{d}\left[\cos(\theta_\Delta) - 1\right]$$

In order to determine the exact E-field for the two-ray model at distance d,



From the above phasor diagram,

$$|E_{TOT}(d)| = \sqrt{\left(\frac{E_0 d_0}{d''}\right)^2 + \left(\frac{E_0 d_0}{d'}\right)^2 - 2\frac{(E_0 d_0)^2}{d''.d'}.\cos\theta_\Delta}$$

$\approx$ As $d'' = d' \approx d$,

$$= \sqrt{\left(\frac{E_0 d_0}{d}\right)^2 + \left(\frac{E_0 d_0}{d}\right)^2 - 2\left(\frac{E_0 d_0}{d}\right)^2 \cos\theta_\Delta}$$

$$= \sqrt{2\left(\frac{E_0 d_0}{d}\right)^2 (1 - \cos\theta_\Delta)}$$

$$= \sqrt{2\left(\frac{E_0 d_0}{d}\right)^2 . 2\sin^2\frac{\theta_\Delta}{2}} \qquad \left(1 - 2\sin^2\theta = \cos 2\theta\right)$$

$$|E_{TOT}(d)| = 2 \cdot \frac{E_0 d_0}{d} . \sin\frac{\theta_\Delta}{2}$$

For very small values of $\frac{\theta_\Delta}{2}$,

$$\sin\frac{\theta_\Delta}{2} \approx \frac{\theta_\Delta}{2} \quad \text{for } \frac{\theta_\Delta}{2} < 0.3 \text{ rad}$$

$$\Rightarrow \frac{\theta_\Delta}{2} = \frac{2\pi h_t h_r}{\lambda d} < 0.3 \text{ rad}$$

$$\Rightarrow \boxed{d > \frac{20\pi h_t h_r}{3\lambda} \approx \frac{20 h_t h_r}{\lambda}}$$

As long as d satisfies the above equation,

$$\boxed{E_{TOT}(d) \approx 2 \cdot \frac{E_0 d_0}{d} \cdot \frac{2\pi h_t h_r}{\lambda d} = \frac{4\pi E_0 d_0 h_t \cdot h_r}{\lambda d^2}}.$$

Free space power received at distance d is related to the square of the electric field as

$$\boxed{P_r(d) = \frac{|E|^2 A_e}{120\pi} = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2}} \quad \text{——— (1)}$$

$$\Rightarrow P_r(d) = \frac{(4\pi)^2 E_0^2 d_0^2}{\cancel{\lambda^2}} \cdot \frac{h_t^2 h_r^2}{d^4} \times \frac{G_r \cancel{\lambda^2}}{4\pi \times 120\pi}$$

$$= \frac{E_0^2 d_0^2 \cdot G_r}{30} \frac{h_t^2 h_r^2}{d^4} \quad \text{——— (2)}$$

$$P_r(d_0) = \frac{E_0^2 \cdot G_r \cdot \cancel{\lambda^2}}{30 \times \cancel{(4\pi)^2}} = \frac{P_t G_t \cdot \cancel{G_r} \cancel{\lambda^2}}{\cancel{(4\pi)^2} d_0^2}$$

$$\Rightarrow \frac{E_0^2 d_0^2}{30} = P_t \cdot G_t .$$

Putting this value in eq^n (2),

$$\boxed{P_r(d) = P_t \cdot G_t \cdot G_r \cdot \frac{h_t^2 \cdot h_r^2}{d^4}.} \qquad L = 1$$

At ~~v~~ large distances, $d >> \sqrt{h_t h_r}$, the received power falls off with distance raised to the 4th power, or at a rate of 40 dB/decade. Also the received power and path loss is independent of frequency.

$$PL(dB) = 40 \log d - (10 \log G_t + 10 \log G_r + 20 \log h_t + 20 \log h_r)$$

Q. A mobile is located 5 km away from a base station and uses a vertical $\lambda/4$ monopole antenna with a gain of 2.55 dB to receive cellular radio signals. The E-field at 1 km from the transmitter is measured to be 10^{-3} V/m. The carrier frequency used for this system is 900 MHz.

(a) Find the length and the effective aperature of the receiving antenna.

(b) Find the received power at the mobile using the two-ray ground reflection model assuming the height of the transmitting antenna is 50 m and the receiving antenna is 1.5 m above ground.

Ans – T-R separation distance, $d = 5$ km.

E-field at a distance of 1 km $= 10^{-3}$ V/m

Frequency of operation, $f = 900$ MHz.

$G_r = 2.55$ dB.

$$\lambda = \frac{c}{f} = \cancel{900 \times 10} \; \frac{3 \times 10^8}{900 \times 10^6} = \frac{1}{3} = 0.333 \text{ m}$$

Length of the antenna, $L = \lambda/4$

$= 0.333/4$

$= 0.0833$ m

Effective aperature, $A_e = \frac{G_r \cdot \lambda^2}{4\pi}$

$= 0.01588$ or $0.016\,m^2$

Since $d >> \sqrt{h_t h_r}$, the electric field is given by

$$E_R(d) = \frac{2 E_0 d_0 \cdot 2\pi h_t h_r}{\lambda d^2} = \frac{2\times10^{-3} \times 1\times10^{3} \times 2\pi \times 50 \times 1.5}{0.333 \times (5\times10^{3})^2}$$

$$= 113.1 \times 10^{-6}\ V/m$$

$$P_r(d) = \frac{|E|^2 A_e}{120\pi} = \frac{(113.1\times10^{-6})^2 \times 0.016}{120 \times 3.141}$$

$$= 0.543 \times 10^{-12}\ W$$

$$= -122.68\ dBW \text{ or } -92.68\ dBm$$

Q. With $h_b = 100\,ft$ (or h_t) and h_m (or h_r) $= 5\,ft$ and a frequency of $881.52\ MHz$, calculate signal attenuation at a distance of $5000\,ft$.

$G_t = 8\,dB$, $G_r = 0\,dB$.

What are the free space and reflected surface attenuations?

Ans - $d = 5000\,ft$

$$\bar{d} = \frac{4 h_t h_r}{\lambda} = \frac{4\times100\times5}{1.116} = 1792\,ft,$$

$d > \bar{d}$.

$G_t = 8\,dB = 10^{0.8}$, $G_r = 0\,dB = 1$.

Free space attenuation:

$$\frac{P_t}{P_r} = \left(\frac{4\pi d}{\lambda}\right)^2 \cdot \frac{1}{G_t G_r}$$

Attenuation on reflecting surface:

$$\frac{P_t}{P_r} = \left(\frac{d^2}{h_t h_r}\right)^2 \cdot \frac{1}{G_t \cdot G_r}$$

Log distance path Loss model

→ Both theoretical and measurement-based propagation models indicate that average received signal power decreases logarithmically with distance.

→ The average large-scale path loss for an arbitrary T-R separation is expressed as a function of distance by using a path loss exponent γ.

→ The value of γ depends on the specific propagation environment.

For e.g. in free space, $\gamma = 2$.

when obstructions are present, γ will have a larger value.

$$P_r = P_0 (d_0/d)^{\gamma}$$

or the path Loss (d) $\alpha \left(\frac{d}{d_0}\right)^{\gamma}$

$$P_r(d) = 10 \log[P_0(d_0)] + 10\gamma \log(d_0/d) \text{ dBm}$$

$$\overline{PL}(dB) = \overline{PL}(d_0) + 10\gamma \log\left(\frac{d}{d_0}\right) \quad \text{——— (1)}$$

(The bar on the PL denote the ensemble average of all possible path loss values for a given value of d.)

→ The accuracy of P_r predicted by the above model can be improved by taking into account the random shadow effect caused by obstructions such as buildings or mountains.

→ Shadow effect is described by a zero-mean Gaussian random variable, X_σ, with standard deviation σ (dB). So the path loss is given as

$$PL(d) = \overline{PL}(d) + X_\sigma$$

$$= \underbrace{\overline{PL}(d_0)}_{\text{reference value of path loss}} + 10\gamma \log(d/d_0) + X_\sigma (dB).$$

→ X_σ is often based on measurement made over a wide range of locations and transmitter-receiver separation. An average value of 8 dB for σ is often used giving X_σ as 10.5 dB.

→ Also, whenever relative motion exists between transmitter and receiver, there is a change in frequency of the wave for the receiver which is known as Doppler shift in the received signal.

The maximum Doppler shift in frequency is given as:

$$f_m = \frac{v_m}{\lambda_c} = \frac{v_m}{c/f_c}$$

v_m → velocity of the moving vehicle

f_c → frequency of the carrier

Q. Calculate the received power at a distance of 3 km from the transmitter if the path loss exponent γ is 4. Assume the transmitting power of 4W at 1800 MHz, a shadow effect of 10.5 dB and the power at reference distance, $d_0 = 100m$ of -32 dBm. What is the allowable path loss?

$$P_r(d) = 10 \log P_0(d_0) + 10\gamma \log(d_0/d) + X_\sigma$$

$$= -32 + 10 \times 4 \times \log\left(\frac{100}{3000}\right) + 10.5$$

$$= -80.5 \text{ dBm}$$

Allowable path loss $= P_t - P_r$ (dB)

$$= 36 - (-80.5)$$

$$= 116.5 \text{ dB}$$

Q. What is the separation distance between the Tx and the Rx with an allowable path loss of 150 dB and shadow effect of 10 dB? The path loss is given as

$$PL = 133.2 + 43 \log d.$$

Allowable path loss = Path loss + shadow effect

$$\Rightarrow 150 = 133.2 + 43 \log d + 10$$

$$\Rightarrow \log d = \frac{6.8}{40} = 0.17$$

$$\Rightarrow d = 10^{0.17} = 1.48 \text{ Km}.$$

Okumara / Hata Model

→ Okumara's model is one of the most widely used models for signal prediction in urban areas which is based on a large amount of experimental data collected around Tokyo, Japan.

→ Okumara developed a set of curves giving the average path-loss relative to free space (Amu), in an urban area over quasi-smooth terrain with a base station antenna height (hte) of 200m and a mobile antenna height (hre) of 3m.

→ To determine path loss using Okumara's model, the free space path ~~ba~~ loss is first determined and then the value of Amu(f,d) is added to it along with correction factors to account for the type of terrain

The model is expressed as:

$$L_{50}(dB) = L_F + A_{mu}(f,d) - G(h_{te}) - G(h_{re}) - G_{AREA}$$

where L_{50} is the median value of propagation path loss.

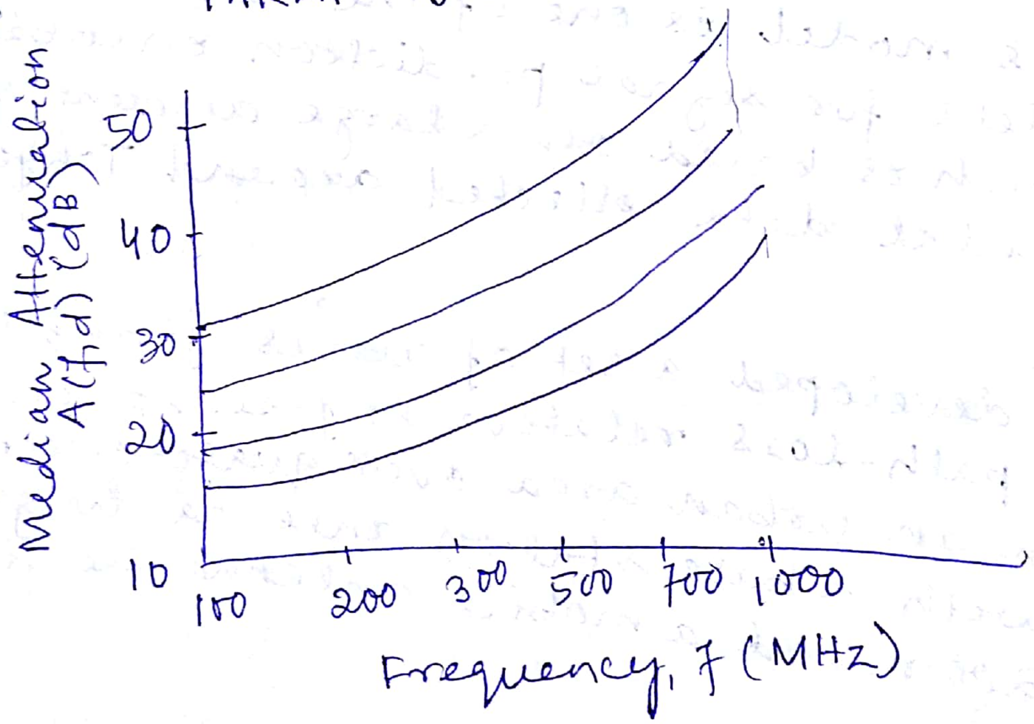
L_F : free space propagation loss.

A_{mu} : median attenuation relative to free space

$G(h_{te})$: base station antenna height gain factor

$G(h_{re})$: mobile antenna height gain factor

G_{AREA} : gain due to the type of environment.



$$G(h_{te}) = 20 \log\left(\frac{h_{te}}{200}\right) \quad 1000\,m > h_{te} > 30\,m.$$

$$G(h_{re}) = 10 \log\left(\frac{h_{re}}{3}\right) \quad h_{re} \leq 3\,m$$

$$= 20 \log\left(\frac{h_{re}}{3}\right) \quad 10\,m > h_{re} > 3\,m.$$

→ The Okumara's model is very practical and has become a standard for system planning in modern land mobile radio systems in Japan.

→ The major disadvantage with the model is its slow response to rapid changes in terrain. Therefore it is fairly good in urban and suburban areas, but not in rural areas.

→ ~~Hata~~ The Hata model is an empirical formulation of the graphical path loss data provided by Okumara. It is valid from 150 MHz to 1500MHz. It is given for three ~~condition~~ categories:

(i) Typical urban

The standard formula for median path loss in urban areas is given by

$$L_{50}(urban) = 69.55 + 26.16 \log f_c - 13.82 \log h_{te} - a(h_{re}) + (44.9 - 6.55 \log h_{te}) \log d$$

where

$a(h_{re})$ is the correction factor for effective mobile antenna height which is a function of the size of the coverage area.

For large cities,

$$a(h_{re}) = 8.29(\log 1.54 h_{re})^2 - 1.1 \text{ dB}$$

for $f_c \leq 300$ MHz.

$$a(h_{re}) = 3.2(\log 11.75 h_{re})^2 - 4.97 \text{ dB}$$

for $f_c \geq 300$ MHz.

For small to medium sized cities,

$$a(h_{re}) = [1.1 \log f_c - 0.7] h_{re} - (1.56 \log f_c - 0.8) \text{ dB}$$

(ii) Suburban area

To obtain the path loss in a suburban area, the standard Hata model is modified as:

$$L_{50}(\text{dB}) = L_{50}(\text{urban}) - 2[\log(f_c/28)]^2 - 5.4.$$

(iii) Rural area

For path loss in rural areas, the formula is modified as

$$L_{50}(\text{dB}) = L_{50}(\text{urban}) - 4.78(\log f_c)^2 + 18.33 \log f_c - 40.94.$$

The predictions of the Hata model compare very closely with the original Okumara model for the following range of parameters:

$150 \leq f_c \leq \cancel{2200}\ 1500$ MHz

$30 \leq h_{te} \leq 200$ m

$1 \leq h_{re} \leq 10$ m

$1 \leq d \leq 20$ km

Cost 231 Model

→ An extension to the Okumara-Hata model was developed by the European Cooperative for Scientific and Technical Research (EURO-COST) called the ~~cos~~ COST-231 model.

→ It ~~works~~ is applicable for frequencies from 1500 MHz to 2 GHz (2000 MHz).

→ The path loss according to this model is

$$L_{50}(\text{urban}) = 46.3 + 33.9 \log f_c - 13.82 \log h_{te} - a(h_{re}) + (44.9 - 6.55 \log h_{te}) \log d + C_M .$$

where $a(h_{re})$ is same as defined by the Hata model.

$C_M = 0$ dB for medium sized city and suburban areas

3 dB for metropolitan centers.

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Small-Scale Fading and Multipath

Small scale fading or simply fading is used to describe the rapid fluctuations of the amplitudes, phases or multipath delays of a radio signal over a short period of time or travel distance.

Fading is caused by interference between two or more versions of the transmitted signal which arrive at the receiver at slightly different times. These signals are called multipath signals.

The three most important fading effects are:

- Rapid changes in signal strength over a small distance or time interval.
- Random frequency modulation due to varying Doppler shifts on different multipath signals.
- Time dispersion (echoes) caused by multipath propagation delays.

Factors influencing Small-scale fading

- Multipath propagation
  The random phase and amplitudes of different multipath components (resulting from reflecting objects and scattering) cause fluctuations in signal strength, inducing small scale fading, signal distortion.

- Speed of the mobile

The relative motion between the base station and the mobile results in random frequency modulation due to different Doppler shifts on each of the multipath components. Doppler shift will be +ve or -ve depending on whether the mobile receiver is moving towards or away from the base station.

- Speed of surrounding objects

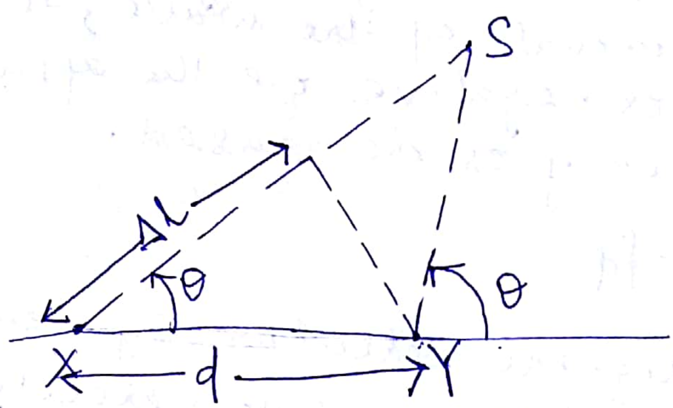
If objects in the radio channel are in motion, they induce a time varying Doppler shift on multipath components

- The transmission bandwidth of the signal

If the transmitted radio signal bandwidth is greater than the bandwidth of the ~~mult~~ multipath channel, the received signal will be distorted.

Doppler Shift -

Let us consider a mobile moving at a constant velocity v along a path segment having length d bet^n points X and Y. It receives signals from a remote source S.



The difference in path lengths travelled by the wave from source S to the mobile at X and Y is

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

$\Delta t \rightarrow$ time required for the mobile to travel from X to Y.

~~If~~ Since the source is assumed to be far away from points X and Y, θ is assumed to be same at points X and Y.

The phase change in the received signal due to the difference in path lengths is,

$$\Delta\phi = \frac{2\pi \Delta l}{\lambda} = \frac{2\pi v \Delta t \cos\theta}{\lambda}.$$

Hence the apparent change in frequency or Doppler shift is given by f_d where

$$\boxed{f_d = \frac{1}{2\pi} \cdot \frac{\Delta\phi}{\Delta t} = \frac{v}{\lambda} \cos\theta}$$

From the equation, it can be seen that,

→ if the mobile is moving towards the direction of arrival of the wave, the Doppler shift is positive i.e. the apparent received frequency is increased.

$$f = f_c + f_d$$

→ if the mobile is moving away from the direction of arrival of the wave, the Doppler shift is negative i.e. the apparent received frequency is decreased.

$$f = f_c - f_d$$

Q. In the U.S. digital cellular system, if $f_c = 900$ MHz and the mobile velocity is 70 km/hr, calculate the received carrier frequency if the mobile is moving

(a) directly toward the transmitter

(b) directly away from the transmitter

(c) in a direction perpendicular to the direction of the arrival of the transmitted signal.

Ans - $f_c = 900$ MHz.

$$\lambda = \frac{c}{f_c} = \frac{3 \times 10^8}{900 \times 10^6} = 1/3 = 0.33\,m.$$

$$\text{Vehicle speed} = \frac{70 \times 1000}{60 \times 60} = 19.44\ m/sec$$

(a) ~~The receivd frequency~~

(a) The vehicle is moving directly towards the transmitter. The received frequency is

$$f = f_c + f_d = 900 \times 10^6 + \frac{19.44}{0.33} = 900.0000589 \text{ MHz}$$

(b) $$f = f_c - f_d = 900 \times 10^6 - \frac{19.44}{0.33} = 899.9999411 \text{ MHz}$$

(c) $\theta = 90^\circ \Rightarrow \cos\theta = 0$ and hence there is no Doppler shift. The received signal freq is the same as the transmitted frequency of 900 MHz.

Types of Small-Scale fading

The time dispersion and frequency dispersion mechanisms in a mobile radio channel lead to four possible different types of fading effects.

Small Scale fading
(Based on multipath time delay spread)

- Flat fading
  1. BW of signal $\ll$ BW of channel
  2. Delay spread < symbol period
- Frequency selective fading
  1. BW of signal > BW of channel
  2. Delay spread > Symbol period.

Small Scale Fading (Based on Doppler spread)

Fast Fading

1. High Doppler spread
2. Coherence time < Symbol period
3. Channel variations faster than baseband signal variations.

Slow fading

1. Low Doppler spread
2. Coherence time > Symbol period
3. Channel variations slower than baseband signal variations.

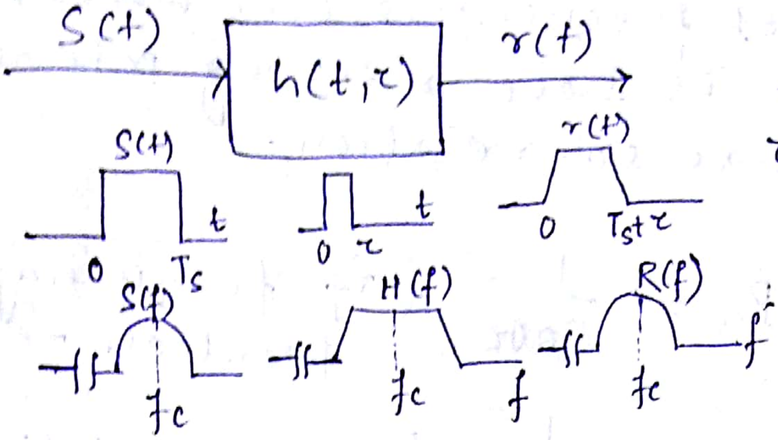
Fading effects due to Multipath time delay Spread

Flat fading

→ If the mobile radio channel has a constant gain and linear phase response over a bandwidth which is greater than the bandwidth of the transmitted signal, then the received signal will undergo flat fading.

→ In flat fading, the multipath structure of the channel is such that the spectral characteristics of the transmitted signal are preserved at the receiver.

→ But the strength of the received signal changes with time, due to fluctuations in the gain of the ~~bandwidth~~ channel caused by multipath.



$\tau << T_s$

→ In a flat fading channel, the reciprocal bandwidth of the transmitted signal is much larger than the multipath time delay spread of the channel

→ Flat fading channels are also known as amplitude varying channels and also referred to as narrowband channels, since the BW of the signal is narrow as compared to the channel flat fading bandwidth.

→ Flat fading can lead to deep fades of more than 30 to 40 dB

→ A signal undergoes flat fading if

$$B_s << B_c$$

$$\text{and } T_s >> \sigma_\tau$$

T_s → reciprocal bandwidth (e.g. symbol period)

B_s → ~~BS~~ bandwidth of the signal

σ_τ → rms delay spread of the channel

B_c → coherence bandwidth of the channel.

→ Coherence bandwidth is the range of frequencies over which the channel can be considered "flat". The frequency band over which the attenuation remains constant i.e. all frequency components behave identically.

→ fast fading occurs for very low data rates

→ In a flat fading, fast fading channel, the amplitude of the channel impulse response varies faster than the rate of change of transmitted signal.

→ In a frequency selective, fast fading channel, the amplitudes, phases and time delays of any of the multipath components vary faster than the rate of change of transmitted signal.

<u>Slow fading</u>

→ The channel impulse response changes at a rate much slower than the transmitted signal $s(t)$.

$$T_S << T_C$$

$$B_S >> B_D$$

→ Doppler spread is much less than the BW of the baseband signal.

<u>Multipath Delay Spread, Coherence Bandwidth and Coherence Time</u>

→ Multipath delay spread is the time dispersion characteristic of the channel.

→ τ_d → multipath rms delay spread

f_m → maximum spread in frequency due to Doppler shift.

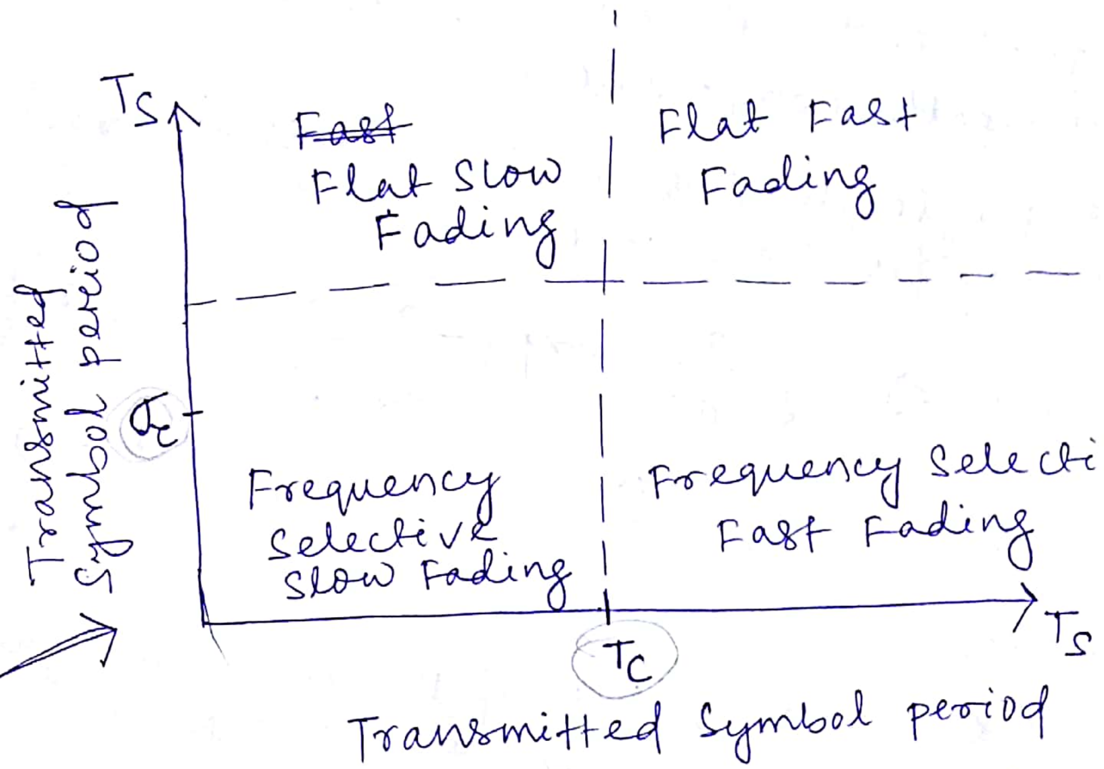
The coherence bandwidth is given by ~~B_d~~

$$B_C \simeq \frac{1}{2\pi \tau_d}$$

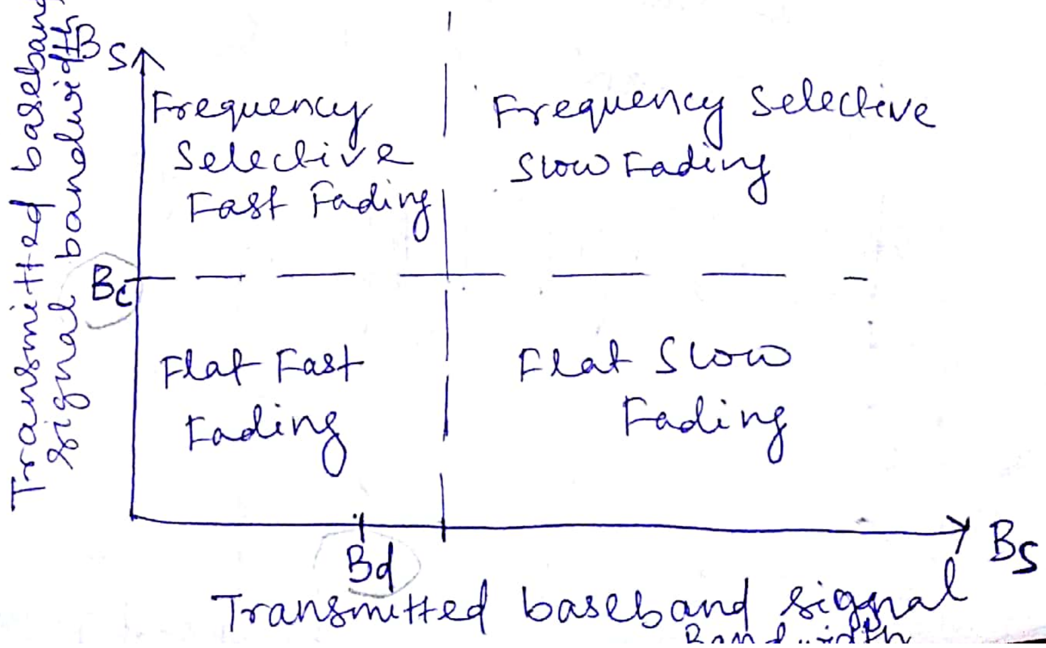
→ The coherence time is approximately inversely proportional to Doppler spread.

$$T_c \approx \frac{1}{2\pi f_m}$$

$f_m = \frac{V}{\lambda}$ → maximum Doppler spread.



Type of fading experienced by a signal as a function of symbol period.



Rayleigh and Ricean Distributions (Signal Fading Statistics)

Rayleigh Fading Distribution

→ The Rayleigh distribution is commonly used to describe the statistical time varying nature of the received envelope of a flat fading signal or the envelope of an individual multipath component.

The Rayleigh distribution has a probability density function given by

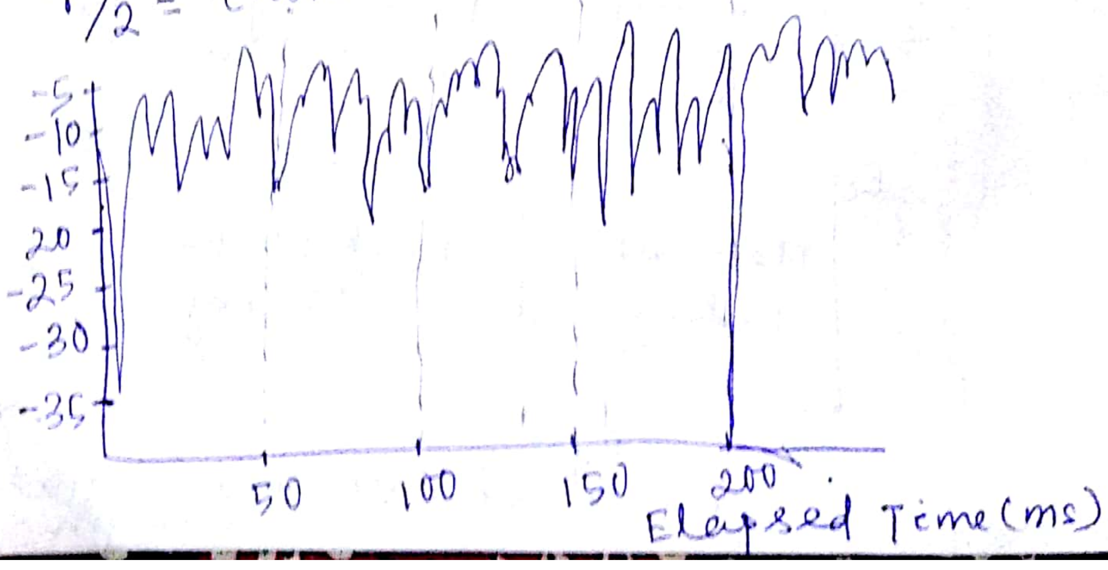
$$p(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) & (0 \le r \le \infty) \\ 0 & (r < 0) \end{cases}$$

where

σ = rms value of the received signal before envelope detection.

σ^2 = time-average power of the received signal before envelope detection.

$r^2/2$ = instantaneous power



The probability that the envelope of the received signal does not exceed a specified value R is given by the corresponding cumulative distribution function (CDF)

$$P(R) = P(r \le R) = \int_0^R p(r)\,dr = 1 - \exp\left(-\frac{R^2}{2\sigma^2}\right)$$

The mean value of the Rayleigh distribution is given by,

$$r_{mean} = E(r) = \int_0^\infty r\,p(r)\,dr$$

$$= \sigma\sqrt{\pi/2} = 1.2533\,\sigma.$$

The variance of the Rayleigh distribution is given by σ_r^2 (which represent the ac power in the signal envelope)

$$\sigma_r^2 = E[r^2] - E^2[r] = \int_0^\infty r^2 p(r)\,dr - \frac{\sigma^2\pi}{2}$$

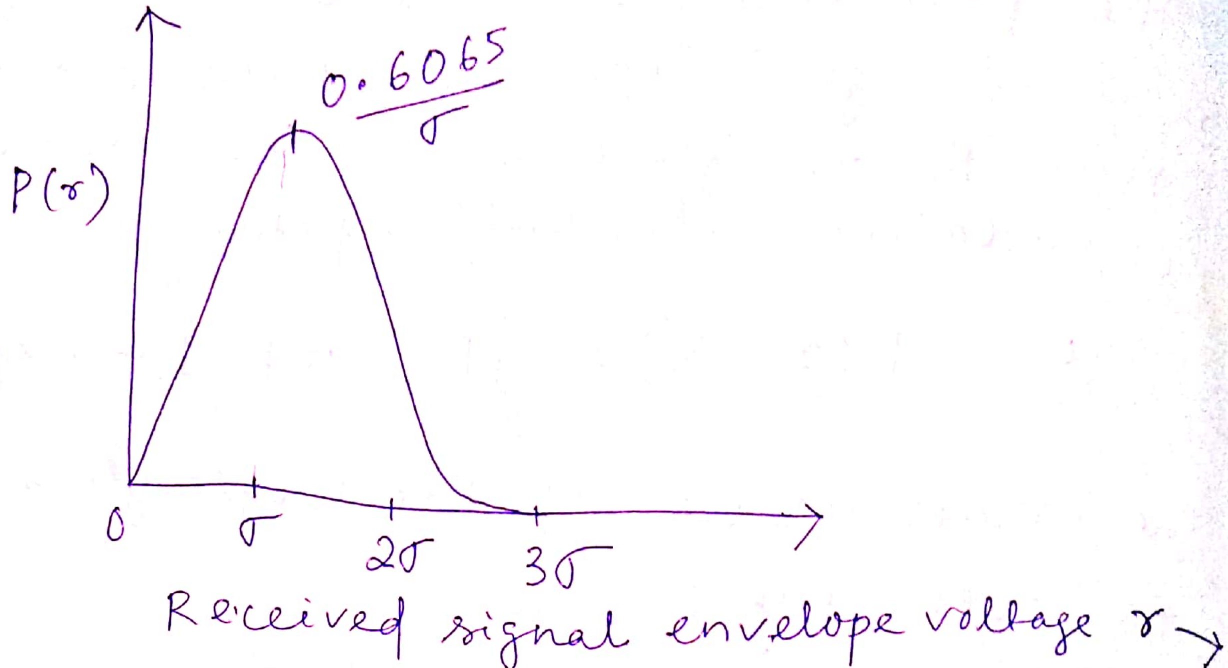
$$= \sigma^2\left(2 - \frac{\pi}{2}\right) = 0.4292\,\sigma^2.$$

The rms value of the envelope is the square root of the mean square or $\sqrt{2}\,\sigma$

The median value of r is found by solving

$$\frac{1}{2} = \int_0^{r_{median}} p(r)\,dr$$

$$\Rightarrow r_{median} = 1.177\,\sigma.$$



→ Thus the mean and the median differ by only 0.55 dB in a Rayleigh fading signal.

Ricean Fading Distribution

→ When there is a dominant stationary (nonfading) signal component present, such as a line of sight propagation path, the small scale fading envelope distribution is Ricean.

→ In this case, random multipath components arriving at different angles are superimposed on a stationary dominant signal.

→ At the output of an envelope detector, this has the effect of adding a dc component to the random multipath.

→ As the dominant signal becomes weaker, the composite signal resembles a noise signal which follows Rayleigh distribution.

Thus, the Ricean distribution degenerates to a Rayleigh distribution when the dominant component fades away.

The Ricean distribution is given by

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

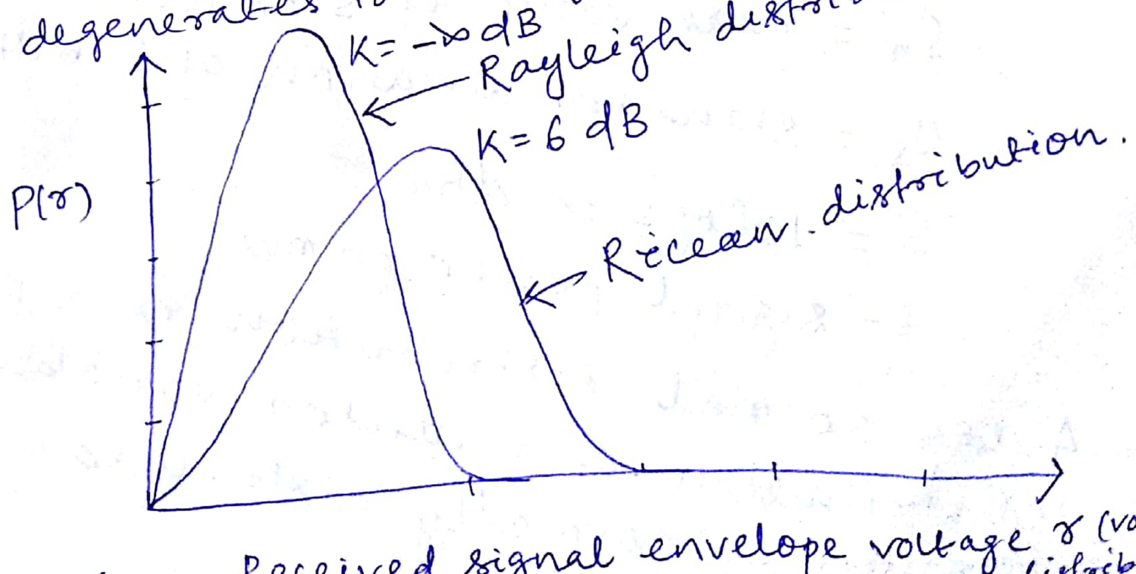
Here $A \rightarrow$ Peak amplitudes of the dominant signal.

$I_0(\cdot\cdot) \rightarrow$ modified Bessel function of the first kind and zero order.

The Ricean distribution factor, K is expressed as:

$$K_{(dB)} = 10 \log \frac{A^2}{2\sigma^2} \text{ dB}.$$

As $A \rightarrow 0$, $K \rightarrow -\infty$ dB and as the dominant path decreases in amplitude, the Ricean distribution degenerates to a Rayleigh distribution.



(Probability density function of Ricean distribution)

Lognormal Distribution

→ Log normal distribution describes the random shadowing effects which occur over a large number of measurement locations which have the same transmitter and receiver separation, but have different levels of clutter on the propagation path.

→ ~~The signal, S(t) typically follows~~

→ Log normal shadowing implies that the measured signal levels at a specific T-R separation have a Gaussian (normal) distribution where the measured signal levels have values in dB units. The standard deviation of the Gaussian distribution that describes the shadowing has also units in dB.

The lognormal distribution is given by

$$P(S) = \frac{1}{\sqrt{2\pi\sigma_s^2}} e^{-\left[\frac{(S-S_m)^2}{2\sigma_s^2}\right]}$$

S_m = mean value of S in dBm

σ_s = standard deviation of S in dB

$S = 10 \log s$ in dBm

s = signal power in mW.

A log-normal distribution is a probability distribution of a random variable whose logarithm is normally distributed.

i.e. Here $S = 10 \log s$ is normally distributed.