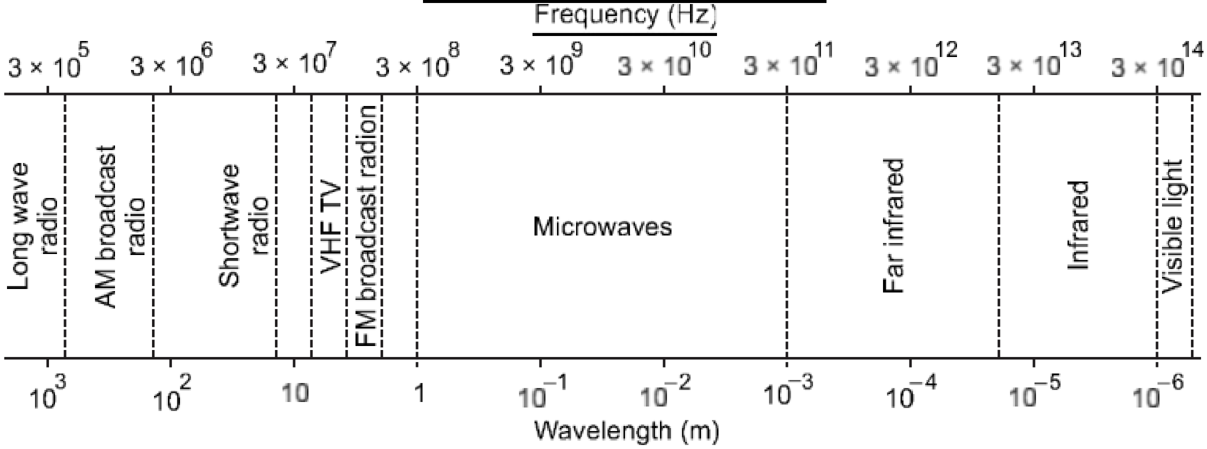
UNIT - 1

Introduction to Microwave Systems and Antennas

Microwave Frequency Bands



Physical Concepts of radiation

For wireless communication systems, the antenna is one of the most critical components. A good design of the antenna can relax system requirements and improve system performance. A example is TV for which the overall typical overall. broadcast reception can be improved by utilizing a high performance antenna.

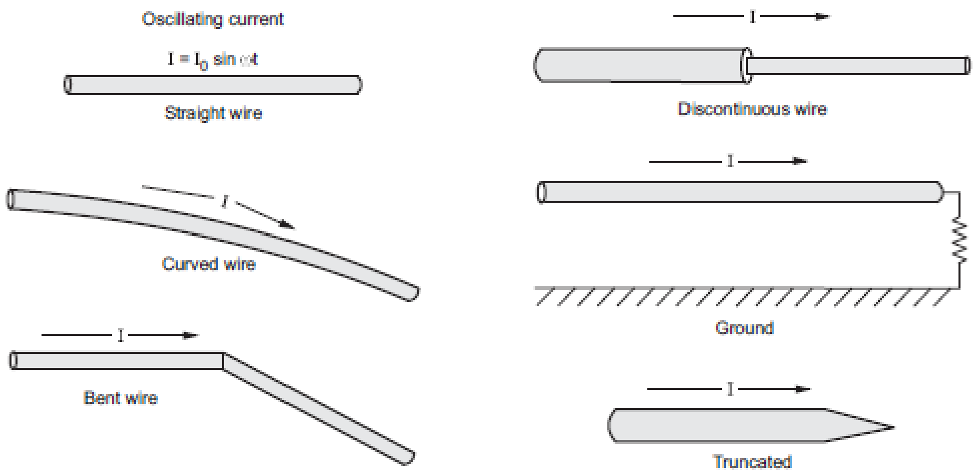
An antenna is the system component that is designed to radiate or receive electromagnetic waves. In other words, the antenna is the electromagnetic transducer which is used to convert, in the transmitting mode, guided waves within a transmission line to radiate free-space waves or to convert, in the receiving mode, free-space waves to guided waves. In a modern wireless system, the antenna must also act as a directional device to optimize or accentuate the transmitted or received energy in some directions while suppressing it in others. The antenna serves to a communication system the same purpose that eyes and eye glasses serve to a human.

Radiation from Single Wire

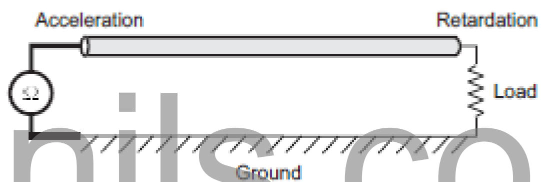
The conducting wire can radiate electromagnetic energy if there is :

- Oscillating current in wire

- Steady current in curved, bent, discontinuous, terminated or truncated wire



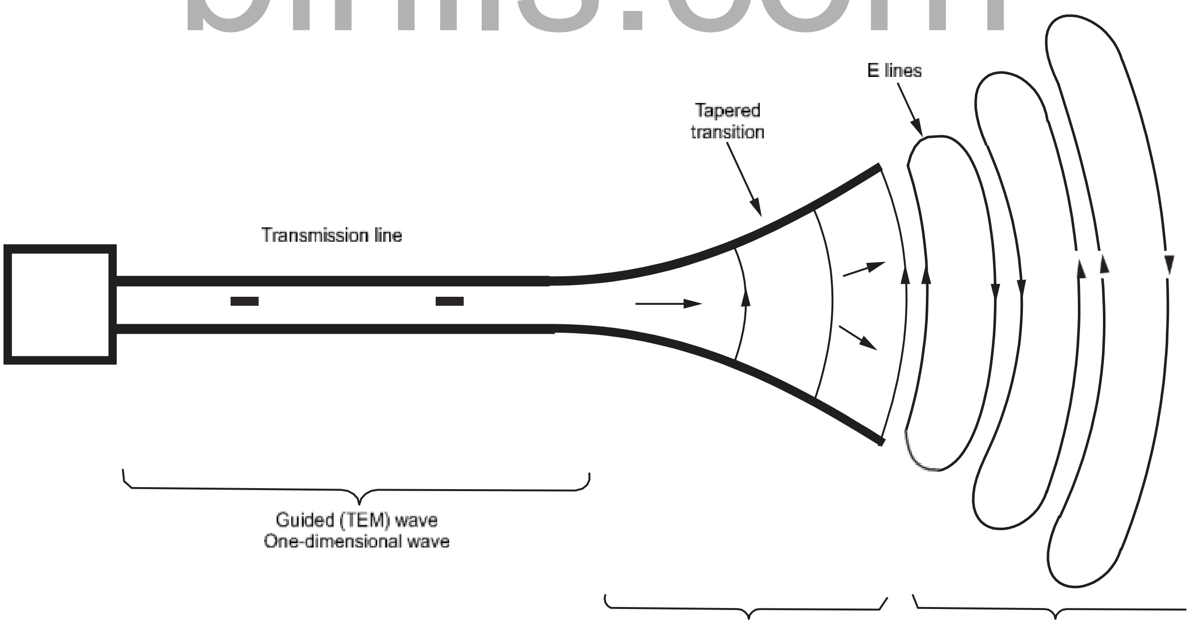
- Consider a pulse source is applied to the end of the wire and the other end is connected to the ground via load



- The free electrons are accelerated from the source end and are retarded at the load end due to the build-up of the electrons there. The electromagnetic radiations are produced at the ends and along the length of the wire. Since the magnitude of acceleration or retardation is not uniform throughout, as a result there is a broad frequency spectrum (frequency is proportional to acceleration or retardation of electric charge). The band width depends upon the pulse width.
- For AC current, ideally there is single frequency of radiation. The moving charge through curved or bent wire experiences centripetal acceleration, which also products radiation. For discontinuous wire impedance change rapidly at the point of discontinuity which is also responsible for radiation.

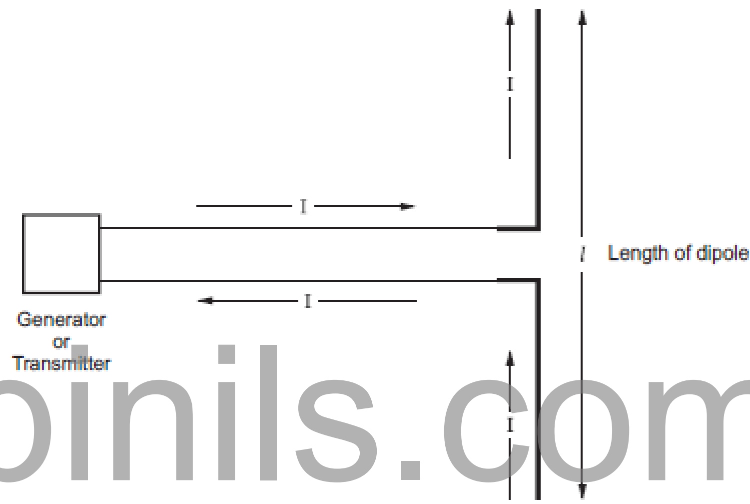
Radiation from Two Wires

- Consider two straight conducting wires connected through generator or transmitter as shown in Fig. 1.2.4. The AC current in the two wires is same but their directions are opposite. If the separation between the conductors is very small as compared to the wavelength, then the electromagnetic fields of both the wires cancel each other and as a result there is no net radiation. However when the open end of the wires are tapered which results in the increased separation between the wires, secondly the directions of currents in the two wires now are not exactly opposite which results in no net canceling of electromagnetic fields and the structure starts radiating.

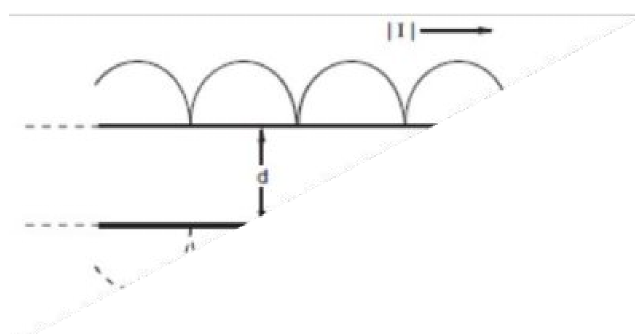


Radiation from Dipole

- If the transition region of the two conductor wires is bent through 90°, the two currents become exactly parallel to each other In the next quarter period, the original field line travels additional distance of and the total distance becomes simultaneously the charge on the dipode becomes zero. This can be through of as being neutralized by the appearance of opposite charges. The field lines by this opposite charges are shown dashed (anticlockwise sense). Since there is no net charge on the dipole and the existence of such-line is only possible when they form closed loops. These closed loops propagate away resulting in electromagnetic radiation.



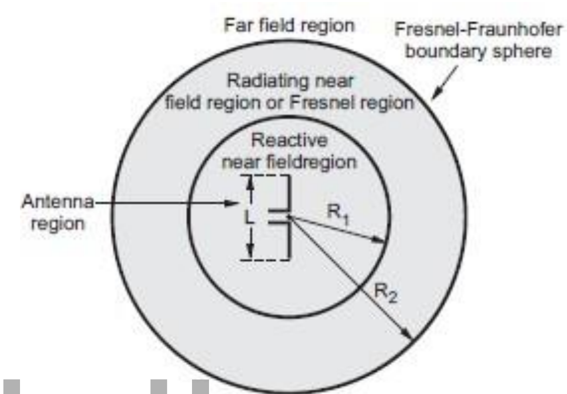
Current Distribution on Thin Wire Antenna



Consider two wire transmission line which is open at load side. The source sends traveling current wave in the wires. The current at the each end reflects back with phase changes of 180°. The incident and reflected current combine to produce a standing wave pattern of sinusoidal form. If the separation between the lines is very small as compared to wavelength, there will not be any radiation. If the open and region is bent to 90°, the currents in the two vertical sections become in the same direction.

Near and Far Field Radiation

There are mainly two principle regions or zones of the antenna fields namely near field or Fresnel region and far field or Fraunhofer region. The near field region may be further classified as **reactive near field region** and **radiating near field region.**



Reactive near field region

- This region is the portion of near field region very close to antenna where the reactive field is dominating.
- For most of the antennas used, the boundary of reactive near field exists at But for the shorter dipole, the boundary exists at

Radiating near field or Fresnel region

- This region exists in between reactive near field and far field regions.
- For this region, the distance from the antenna is taken to
- In this region radiating field dominates reactive field. Also the antenna pattern starts taking shape in this region but it is not completed. The distance from the antenna decides the angular field distribution.
- This Region may not exist if the maximum dimension of antenna is not large compared to wavelength.

Far field or Fraunhofer region

- In this region, the angular field distribution is independent of distance from antenna.
- the boundary for this region exists at
- In this region, the radiation pattern is completely formed. The electric and magnetic field vectors are orthogonal to each other and wavefront becomes planer approximately.

Antenna gain and Efficiency

Resistive losses, due to nonperfect metals and dielectric materials, exist in all practical antennas. Such losses result in a difference between the power delivered to the input of an antenna and the power radiated by that antenna. As with many other electrical components, we can define the **radiation efficiency** of an antenna as the ratio of the desired output power to the supplied input power:

$$\eta_{rad} = P_{rad} / P_{in} = (P_{in} - P_{loss}) / P_{in} = 1 - (P_{loss} / P_{in})$$

P_{rad} is the power radiated by the antenna,

P_{in} is the power supplied to the input of the antenna, and

P_{loss} is the power lost in the antenna

there are other factors that can contribute to the effective loss of transmit power, such as impedance mismatch at the input to the antenna, or polarization mismatch with the receive antenna. However, these losses are external to the antenna and could be eliminated by the proper use of matching networks, or the proper choice and positioning of the receive antenna. Therefore losses of this type are usually not attributed to the antenna itself, as are dissipative losses due to metal conductivity or dielectric loss within the antenna.

we define **antenna gain** as the product of directivity and efficiency:

$$G = \eta_{rad} D.$$

gain is always less than or equal to directivity.

Aperture efficiency and effective area

- Many types of antennas can be classified as *aperture antennas*, meaning that the antenna has a well-defined aperture area from which radiation occurs.
- the maximum directivity that can be obtained from an electrically large aperture of area A is given as

$$D_{max} = 4\pi A / \lambda^2.$$

- the directivity of an aperture antenna as

$$D = \eta_{ap}(4\pi A / \lambda^2)$$

- Aperture efficiency is always less than or equal to unity.
- The above definitions of antenna directivity, efficiency, and gain were stated in terms of a transmitting antennas, but they apply to receiving antennas as well. For a receiving antenna it is also of interest to determine the received power for a given incident plane wave field.

The received power will be proportional to the power density, or Poynting vector, of the incident wave.

$$P_r = A_e \cdot S_{avg}$$

A_e is defined as the *effective aperture area* of the receive antenna. The effective aperture area has dimensions of m^2

The maximum effective aperture area of an antenna can be shown to be related to the directivity of the antenna as

$$A_e = D\lambda^2 / 4\pi$$

where λ is the operating wavelength of the antenna

UNIT - 1

Introduction to Microwave Systems and Antennas

Fields and Power radiated by an antenna

Consider an antenna located at the origin of a spherical coordinate system. At large distances, where the localized near-zone fields are negligible, the radiated electric field of an arbitrary antenna can be expressed as,

$$\bar{E}(r,\theta,\phi) = \left[\hat{\theta}F_\theta(\theta,\phi) + \hat{\phi}F_\phi(\theta,\phi)\right]\frac{e^{-jk_0 r}}{r} \text{ V/m},$$

where $\bar{E}$ is the electric field vector, $\hat{\theta}$ and $\hat{\phi}$ are unit vectors in the spherical coordinate system, r is the radial distance from the origin, and $k_0 = 2\pi/\lambda$ is the free-space propagation constant, with wavelength $\lambda = c/f$. The pattern functions, $F_\theta(\theta, \varphi)$ and $F_\varphi(\theta, \varphi)$. Electric field propagates in the radial direction with a phase variation of $e^{-jk_0 r}$ and an amplitude variation with distance of $1/r$. The electric field may be polarized in either the $\hat{\theta}$ or $\hat{\phi}$ direction, but not in the radial direction, since this is a TEM wave. The magnetic fields associated with the electric field of can be found,

$$H_\phi = \frac{E_\theta}{\eta_0}$$

$$H_\theta = \frac{-E_\phi}{\eta_0}$$

where η_0 = 377 _, the wave impedance of free-space.

The Poynting vector for this wave is given by,

$$\bar{S} = \bar{E} \times \bar{H}^* \text{ W/m}^2$$

the time-average Poynting vector is,

$$\bar{S}_{avg} = \frac{1}{2}\text{Re}\{\bar{S}\} = \frac{1}{2}\text{Re}\{\bar{E} \times \bar{H}^*\} \text{ W/m}^2$$

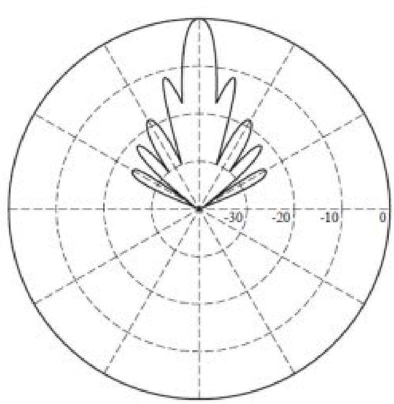
We mentioned earlier that at large distances the near fields of an antenna are negligible. We can give a more precise meaning to this concept by defining the far-field distance as the distance where the spherical wave front radiated by an antenna becomes a close approximation to the ideal planar phase front of a plane wave. This approximation applies over the radiating aperture of the antenna, and so it depends on the maximum dimension of the antenna. If we call this maximum dimension D, then the far-field distance is defined as,

$$R_{ff} = \frac{2D^2}{\lambda} \text{m.}$$

This result is derived from the condition that the actual spherical wave front radiated by the antenna departs less than $\pi/8 = 22.5^\circ$ from a true plane wave front over the maximum extent of the antenna. For electrically small antennas, such as short dipoles and small loops, this result may give a far-field distance that is too small; in this case, a minimum value of $R_{ff} = 2\lambda$ should be used.

Antenna pattern characteristics

The radiation pattern of an antenna is a plot of the magnitude of the far-zone field strength versus position around the antenna, at a fixed distance from the antenna. Thus the radiation pattern can be plotted from the pattern function $F_\theta(\theta, \varphi)$ or $F_\varphi(\theta, \varphi)$, versus either the angle θ (for an elevation plane pattern) or the angle φ (for an azimuthal plane pattern). The choice of plotting either F_θ or F_φ is dependent on the polarization of the antenna.



pattern is plotted in polar form, versus the elevation angle, θ, for a small horn antenna oriented in the vertical direction. The plot shows the relative variation of the radiated power of the antenna in dB, normalized to the maximum value. Since the pattern functions are proportional to voltage, the radial scale of the plot is computed as 20 log $|F(\theta, \varphi)|$; alternatively, the plot could be computed in terms of the radiation intensity as 10 log $|U(\theta, \varphi)|$. The pattern may exhibit several distinct lobes, with different maxima in different directions. The lobe having the maximum value is called the main beam, while those lobes at lower levels are called sidelobes.

A fundamental property of an antenna is its ability to focus power in a given direction, to the exclusion of other directions. Thus an antenna with a broad main beam can transmit (or receive) power over a wide angular region, while an antenna having a narrow main beam will transmit (or receive) power over a small angular region. One measure of this focusing effect is the 3 dB beamwidth of the antenna, defined as the angular width of the main beam at which the power level has dropped 3 dB from its maximum value.

Antennas having a constant pattern in the azimuthal plane are called omnidirectional, and are useful for applications such as broadcasting or for hand-held wireless devices, where it is desired to transmit or receive equally in all directions. Patterns that have relatively narrow main beams in both planes are known as pencil beam antennas, and are useful in applications such as radar and point-to-point radio links.

Another measure of the focusing ability of an antenna is the directivity, defined as the ratio of the maximum radiation intensity in the main beam to the average radiation intensity over all space:

$$D = \frac{U_{\max}}{U_{\text{avg}}} = \frac{4\pi U_{\max}}{P_{\text{rad}}} = \frac{4\pi U_{\max}}{\int_{\theta=0}^{\pi} \int_{\phi=0}^{2\pi} U(\theta, \phi) \sin\theta d\theta d\phi}$$

An antenna that radiates equally in all directions is called an *isotropic* antenna.

$$\int_{\theta=0}^{\pi} \int_{\phi=0}^{2\pi} \sin\theta d\theta d\phi = 4\pi$$

Since the minimum directivity of any antenna is unity, directivity is sometimes stated as relative to the directivity of an isotropic radiator, and written as dBi. Beamwidth and directivity are both measures of the focusing ability of an antenna: an antenna pattern with a narrow main beam will have a high directivity, while a pattern

with a wide beam will have a lower directivity. We might therefore expect a direct relation between beamwidth and directivity, but in fact there is not an exact relationship between these two quantities. This is because beamwidth is only dependent on the size and shape of the main beam, whereas directivity involves integration of the entire radiation pattern. Thus it is possible for many different antenna patterns to have the same beamwidth but quite different directivities due to differences in sidelobes or the presence of more than one main beam. With this qualification in mind, however, it is possible to develop approximate relations between beamwidth and directivity that apply with reasonable accuracy to a large number of practical antennas. One such approximation that works well for antennas with pencil beam patterns is the following:

$$D \cong \frac{32{,}400}{\theta_1 \theta_2},$$

where $\theta 1$ and $\theta 2$ are the beamwidths in two orthogonal planes of the main beam, in degrees.

Antenna noise temperature and *G/T*

If a receiving antenna has dissipative loss, so that its radiation efficiency ηrad is less than unity, the power available at the terminals of the antenna is reduced by the factor $\eta_{rad}$from that intercepted by the antenna (the definition of radiation efficiency is the ratio of output to input power). This reduction applies to received noise power, as well as received signal power, so the noise temperature of the antenna will be reduced from the brightness temperature by the factor ηrad. In addition, thermal noise will be generated internally by resistive losses in the antenna, and this will increase the noise temperature of the antenna. In terms of noise power, a lossy antenna can be modeled as a lossless antenna and an attenuator having a power loss factor of $L = 1/\eta_{rad}$. The resulting noise temperature seen at the antenna terminals as,

$$T_A = \frac{T_b}{L} + \frac{(L-1)}{L} T_p = \eta_{\text{rad}} T_b + (1 - \eta_{\text{rad}}) T_p.$$

The equivalent temperature TA is called the antenna noise temperature, and is a combination of the external brightness temperature seen by the antenna and the thermal noise generated by the antenna.

it is important to realize the difference between radiation efficiency and aperture efficiency, and their effects on antenna noise temperature. While radiation efficiency accounts for resistive losses, and thus involves the generation of thermal noise, aperture efficiency does not. Aperture efficiency applies to the loss of directivity in aperture antenas, such as reflectors, lenses, or horns, due to feed spillover or suboptimum aperture excitation (e.g., a nonuniform amplitude or phase distribution), and by itself does not lead to any additional effect on noise temperature that would not be included through the pattern of the antenna.

The antenna noise temperature defined above is a useful figure of merit for a receive antenna because it characterizes the total noise power delivered by the antenna to the input of a receiver. Another useful figure of merit for receive antennas is the G/T ratio, defined as,

$$G/T\,(\text{dB}) = 10 \log \frac{G}{T_A}\ \text{dB/K},$$

where G is the gain of the antenna, and TA is the antenna noise temperature.The signal-to-noise ratio (SNR) at the input to a receiver is proportional to G/TA. The ratio G/T can often be maximized by increasing the gain of the antenna, since this increases the numerator and usually minimizes reception of noise from hot sources at low elevation angles.

Impedance matching

Impedance matching, which is often an important part of a larger design process for a microwave component or system.

an impedance matching network placed between a load impedance and a transmission line. The matching network is ideally lossless, to avoid unnecessary loss of power, and is usually designed so that the impedance seen looking into the matching network is *Z*0. Then reflections will be eliminated on the transmission line to the left of the matching network, although there will usually be multiple reflections between the matching network and the load. This procedure is sometimes referred to as *tuning*. Impedance matching or tuning is important for the following reasons:

_ Maximum power is delivered when the load is matched to the line (assuming the generator is matched), and power loss in the feed line is minimized.

_ Impedance matching sensitive receiver components (antenna, low-noise amplifier, etc.) may improve the signal-to-noise ratio of the system.

_ Impedance matching in a power distribution network (such as an antenna array feed network) may reduce amplitude and phase errors

As long as the load impedance, *ZL*, has a positive real part, a matching network can always be found. Many choices are available, however, and we will discuss the design and performance of several types of practical matching networks. Factors that may be important in the selection of a particular matching network include the following:

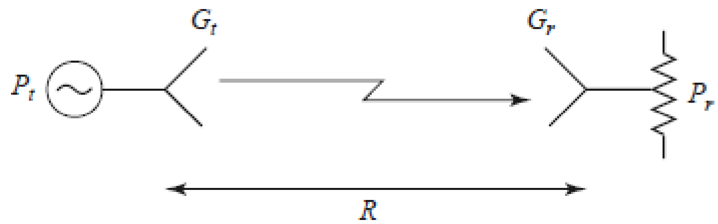
_ *Complexity*—As with most engineering solutions, the simplest design that satisfies the required specifications is generally preferable. A simpler matching network is usually cheaper, smaller, more reliable, and less lossy than a more complex design.

_ *Bandwidth*—Any type of matching network can ideally give a perfect match (zero reflection) at a single frequency. In many applications, however, it is desirable to match a load over a band of frequencies. There are several ways of doing this, with, of course, a corresponding increase in complexity.

Implementation—Depending on the type of transmission line or waveguide being used, one type of matching network may be preferable to another. For example, tuning stubs are much easier to implement in waveguide than are multisection quarter-wave transformers.

_ *Adjustability*—In some applications the matching network may require adjustment to match a variable load impedance. Some types of matching networks are more amenable than others in this regard.

Friis transmission equation



In a general radio system link, where the transmit power is Pt, the transmit antenna gain is Gt , the receive antenna gain is Gr , and the received power (delivered to a matched load) is Pr . The transmit and receive antennas are separated by the distance R.

the power density radiated by an isotropic antenna at a distance *R* is given by

$$S_{\text{avg}} = \frac{P_t}{4\pi R^2} \text{ W/m}^2.$$

This result reflects the fact that we must be able to recover all of the radiated power by integrating over a sphere of radius R surrounding the antenna; since the power is distributed isotropically, and the area of a sphere is 4π / R2 .

If the transmit antenna has

a directivity greater than 0 dB, we can find the radiated power density by multiplying by the directivity, since directivity is defined as the ratio of the actual radiation intensity to the equivalent isotropic radiation intensity. In addition, if the transmit antenna has losses, we can include the radiation efficiency factor, which has the effect of converting directivity to gain. Thus, the general expression for the power density radiated by an arbitrary transmit antenna is

$$S_{\text{avg}} = \frac{G_t P_t}{\;}$$

If this power density is incident on the receive antenna, we can use the concept of effective aperture area, to find the received power:

$$P_r = A_e S_{\text{avg}} = \frac{G_t P_t A_e}{4\pi R^2} \text{ W.}$$

the possibility of losses in the receive antenna can be accounted for by using the gain (rather than the directivity) of the receive antenna. Then the final result for the received

power is

$$P_r = \frac{G_t G_r \lambda^2}{(4\pi R)^2} P_t \text{ W.}$$

This result is known as the Friis radio link formula.

As can be seen from the Friis formula, received power is proportional to the product Pt.Gt . These two factors—the transmit power and transmit antenna gain—characterize the transmitter, and in the main beam of the antenna the product PtGt can be interpreted equivalently as the power radiated by an isotropic antenna with input power Pt.Gt . Thus, this product is defined as the effective isotropic radiated power (EIRP):

$$\text{EIRP} = P_t G_t$$

Link budget and link margin

In a link budget, where each of the factors can be individually considered in terms of its net effect on the received power. Additional loss factors, such as line losses or impedance mismatch at the antennas, atmospheric attenuation (see Section 14.5), and polarization mismatch can also be added to the link budget. One of the terms in a link budget is the path loss, accounting for the free-space reduction in signal strength with distance between the transmitter and receiver. From (14.24), path loss is defined (in dB) as,

$$L_0(\text{dB}) = 20 \log\left(\frac{4\pi R}{\lambda}\right) > 0.$$

Note that path loss depends on wavelength (frequency), which serves to provide a normalization for the units of distance. With the above definition of path loss, we can write the remaining terms of the Friis formula as shown in the following link budget:

Transmit power	P_t
Transmit antenna line loss	$(-)L_t$
Transmit antenna gain	G_t
Path loss (free-space)	$(-)L_0$
Atmospheric attenuation	$(-)L_A$
Receive antenna gain	G_r
Receive antenna line loss	$(-)L_r$
Receive power	P_r

We have also included loss terms for atmospheric attenuation and line attenuation. we can write the receive power as

If the transmit and/or receive antenna is not impedance matched to the transmitter/ receiver (or to their connecting lines), impedance mismatch will reduce the received power by the factor $(1 - |\Gamma|^2)$, where Γ is the appropriate reflection coefficient. The resulting impedance mismatch loss,

$$L_{imp}(\text{dB}) = -10 \log(1 - |\Gamma|^2)$$

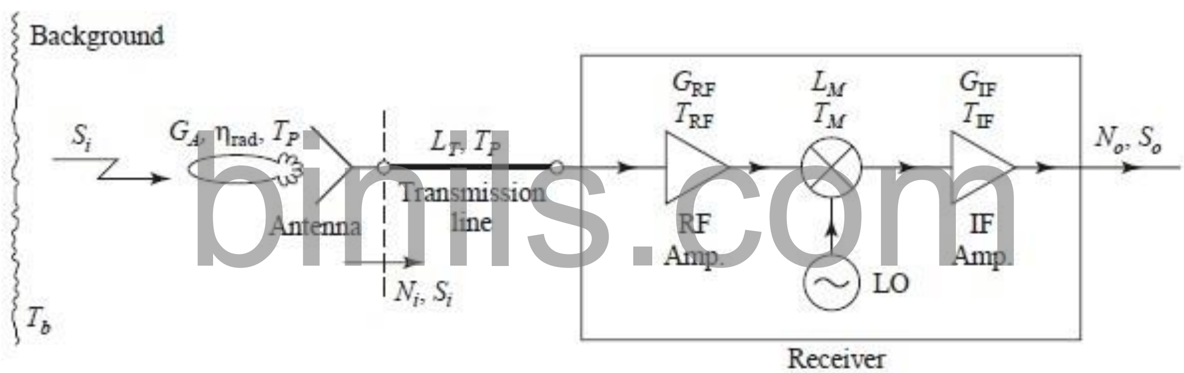
can be included in the link budget to account for the reduction in received power. In practical communications systems it is usually desired to have the received power level greater than the threshold level required for the minimum acceptable quality of service (usually expressed as the minimum carrier-to-noise ratio (CNR), or minimum SNR). This design allowance for received power is referred to as the link margin, and can be expressed as the difference between the design value of received power and the minimum threshold value of receive power:

$$\text{Link margin (dB)} = \text{LM} = P_r - P_{r(\text{min})} > 0,$$

where all quantities are in dB. Link margin should be a positive number; typical values may range from 3 to 20 dB. Having a reasonable link margin provides a level of robustness to the system to account for variables such as signal fading due to weather, movement of a mobile user, multipath propagation problems, and other unpredictable effects that can degrade system performance and quality of service. Link margin that is used to account for fading effects is sometimes referred to as fade margin.

Noise characterization of a microwave receiver

Analyze the noise characteristics of a complete antenna–transmission line–receiver front end. In this system the total noise power at the output of the receiver, No, will be due to contributions from the antenna pattern, the loss in the antenna, the loss in the transmission line, and the receiver components. This noise power will determine the minimum detectable signal level for the receiver and, for a given transmitter power, the maximum range of the communication link.



The receiver components, consist of an RF amplifier with gain GRF and noise temperature TRF, a mixer with an RF-to-IF conversion loss factor LM and noise temperature TM, and an IF amplifier with gain GIF and noise temperature TIF. The noise effects of later stages can usually be ignored since the overall noise figure is dominated by the characteristics of the first few stages. The component noise temperatures can be related to noise figures as T = (F − 1)T0. The equivalent noise temperature of the receiver can be found as,

$$T_{REC} = T_{RF} + \frac{T_M}{G_{RF}} + \frac{T_{IF} L_M}{G_{RF}}.$$

The transmission line connecting the antenna to the receiver has a loss LT , and is at a physical temperature Tp. The equivalent noise temperature is,

$$T_{TL} = (L_T - 1)T_p.$$

we find that the noise temperature of the transmission line (TL) and receiver (REC) cascade is,

$$T_{TL+REC} = T_{TL} + L_T T_{REC} = (L_T - 1)T_p + L_T T_{REC}.$$

This noise temperature is defined at the antenna terminals, we can assume that all noise power comes via the main beam, so that the noise temperature of the antenna is given by,

$$T_A = \eta_{rad} T_b + (1 - \eta_{rad})T_p,$$

where ηrad is the efficiency of the antenna, Tp is its physical temperature, and Tb is the equivalent brightness temperature of the background seen by the main beam. The noise power at the antenna terminals, which is also the noise power delivered to the transmission line is,

$$N_i = kBT_A = kB[\eta_{rad} T_b + (1 - \eta_{rad})T_p],$$

where *B* is the system bandwidth. If *Si* is the received power at the antenna terminals, then the input SNR at the antenna terminals is *Si /Ni* . The output signal power is

$$S_o = \frac{S_i G_{RF} G_{IF}}{[illegible]}$$

where G_{SYS} has been defined as a system power gain. The output noise power is

$$N_o = (N_i + kBT_{TL+REC})\, G_{SYS} = kB(T_A + T_{TL+REC})G_{SYS}$$

$$= kB[\eta_{rad} T_b + (1 - \eta_{rad})T_p + (L_T - 1)T_p + L_T T_{REC}]G_{SYS} = kBT_{SYS}G_{SYS}$$

where T_{SYS} has been defined as the overall system noise temperature. The output SNR is

$$\frac{S_o}{N_o} = \frac{S_i}{kBT_{\mathrm{SYS}}} = \frac{S_i}{kB[\eta_{\mathrm{rad}}T_b + (1-\eta_{\mathrm{rad}})T_p + (L_T - 1)T_p + L_T T_{\mathrm{REC}}]}$$

It may be possible to improve this SNR by various signal processing techniques. Note that it may appear to be convenient to use an overall system noise figure to calculate the degradation in SNR from input to output for the above system, but one must be very careful with such an approach because noise figure is defined only for $Ni = kT_0B$.