NEAR EAST UNIVERSITY

Faculty of Engineering

Department of Electrical and Electronic Engineering

GAIN AND POLARIZATION MEASUREMENTS IN ANTENNA

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EE-400

Student : Naji SAYADI (20002353)

Supervisor: Assoc.Prof.Dr Sameer IKHDAIR

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To all of them, all my love.

ABSTRACT

The term antenna is defined by the dictionary as a usually metallic device (as a rod or wire) for radiating or receiving radio waves. The official definition of the Institute of Electrical and Electronics Engineers (IEEE) is simply as, a means for radiating or receiving radio waves. The ideal antenna is, in most applications, one that will radiate all the power delivered to it by a transmitter in the desired direction or directions and with the desired polarization.

The antenna measurements almost lie within two basic categories: impedance measurements and pattern measurements. Polarization measurements are important only in special cases.

Polarization measurement is common to use the antenna under test to transmit and to use certain standard antennas, or one antenna whose orientation is varied, as receivers. The antenna measurements are very expensive and need gigantic instruments to pursue this work; so that, we decided to search about this subject to make these measurements cheaper and easier ways for finding these results.

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Introduction

The term antenna is defined by the dictionary [1] as a usually metallic device (as a rod or wire) for radiating or receiving radio waves. The official definition of the Institute of Electrical and Electronics Engineers (IEEE) [2] is simply as, a means for radiating or receiving radio waves. The ideal antenna is, in most applications, one that will radiate all the power delivered to it by a transmitter in the desired polarization. Practical antennas can never fully achieve this ideal performance, but their merit is conveniently described in terms of the degree to which they do so. For this purpose, certain parameters of antenna performance are defined.

The antenna measurements are very expensive and need gigantic instruments to pursue this work; so that, we decided to search about this subject to make these measurements cheaper and easier ways for finding these results.

Chapter one is primary concerned with antenna parameters. The principal parameters of antennas are associated with the radiation pattern and efficiency, input impedance, gain, and bandwidth.

Chapter two, discusses gain and directivity and the manner of measuring this parameters. We studied the most practical and most economical way to make these measurement in addition to the problems we can face while making these measurement.

In Chapter Three, we discussed Linear, circular and Elliptical polarization in addition to the main measurement method for polarization: Polarization pattern method, Linear component method and circular component method.

Chapter four was measurement application on Yagi-Uda antenna. What is a Yagi-Uda antenna, performance of Yagi-Uda antenna and laboratory measurement on Yagi Uda antenna.

CHAPTER ONE

ANTENNA PARAMETERS

An antenna is characterized by a number of relevant parameters. These are bandwidth, input impedance, polarization, antenna gain, radiation efficiency and antenna size.

1.1 Bandwidth

The bandwidth, expressed in Hz, is the frequency range over which an antenna exhibit a specified behavior with respect to a relevant antenna parameter. Because there is in general more than one relevant antenna parameter, namely antenna gain, input impedance, polarization, cross polarization, radiation efficiency, the specification of bandwidth must be accompanied by the antenna parameter for which it is specified. For example, bandwidth is very often specified with respect to input impedance, which then leads to bandwidth specified with respect to a certain level of return loss that is still acceptable. A typical bandwidth specification is given as the frequency range where return loss is equal to or better than 10 dB.

1.2 Radiation Resistance

If the power radiated by the antenna is P and the antenna current is I, the radiation resistance is defined as

$$R_r = \frac{P}{I^2} \tag{1.1}$$

This concept is applicable only to antennas in which the radiation is an associated with a definite current in a single linear conductor.

In this limited application, the definition is ambiguous as it stands, because the current is not the same everywhere even in a linear conductor, it is therefore necessary to specify the point in the conductor at which the current will be measured. Two points sometimes specified are the point at which the current has its maximum value and the feed point (input terminals). These two points are sometimes one and the same points, as center-fed in a dipole, but they are not always the same. The value obtained for the radiation resistance of the antenna depends on which point is specified; this value of the radiation resistance referred to that point. The current maximum of a standing wave pattern is known as a current loop, so the radiation resistance referred to the current maximum is sometimes called the loop radiation resistance.

The word maximum here refers to the effect current rms in that part of the antenna where it has its greatest value. In some texts, however, formulas for radiation resistance are written in terms of this peak value, which is the amplitude of the current sine wave. The formula in terms of the current amplitude Io is

$$I_0 = \sqrt{2} \times I_{rms}. \qquad (1.2a)$$

$$R_r = 2P / I_0^{\,2}. \qquad (1.2b)$$

The radiation resistance of some types of antennas can be calculated, when there is clearly defined current value to which it can be referred, but for other types the calculation cannot be made practically, and the value must be obtained by measurement. The typical value of the loop radiation resistance of actual antennas range from a fraction of an ohm to several hundred ohms. The very low values are undesirable because they imply large antenna current, and therefore the possibility of considerable ohmic loss of power, that is , dissipation of the power as heat rather than as radiation. An excessively high value of radiation resistance would also be undesirable because it would require a very high voltage to be applied to the antenna. Very high voltage value do not occur in practical antennas, because there is always some ohmics resistance whereas very low value sometimes do occur unavoidably.

1.3 Radiation Efficiency

The radiation efficiency is defined as the ratio of the power that is radiated by an antenna to the power that is accepted by the antenna. The power accepted by the antenna is equal to the total power fed to the antenna through signal lines minus the power that is reflected by the antenna due to impedance mismatch.

Antennas always do have some ohmic resistance, although sometimes it may be so small as to be negligible. The ohmic resistance is usually distributed over the antenna, and since the antenna current varies, the resulting loss can be considered to be equivalent to the loss in a ficitious lumped resistance placed in series with the radiation resistance. If this equivalent ohmic loss resistance is denoted by Ro, the full power (dissipated plus radiated) is $I^2 \times R_r$. Hence the antenna radiation efficiency ξ given by

$$\xi_r = \frac{R_r}{R_0 - R_r}. \tag{1.3}$$

This formula is not really very useful because both Ro and Rr are fictitious quantities, derived from measurements of current and power;

$$R_r = \frac{P}{I^2} \tag{1.4a}$$

$$R_0 = \frac{P_0}{I^2} \tag{1.4b}$$

$$\xi_r = \frac{P_r}{P_0 + P_r} \tag{1.4c}$$

1.4 Input Impedance

An antenna whose radiation results directly from the flow of RF current in a wire or other linear conductor must somehow have this current introduced into it from a source of RF power transmitters. The current is usually carried to the antenna through a transmission line. To connect the line to the antenna, a small gap is made in the antenna conductor, and the two wires of the transmission line are connected to the terminals of the gap at antenna input terminals. At this point of connection the antenna presents load impedance to the transmission line. This impedance is also the input impedance of the antenna and it is equal to the characteristic of the line Zo.

The impedance match between the antenna and the transmission line is usually expressed in terms of the standing wave ratio (SWR) or the reflection coefficient of the antenna when connected to a transmission line of given impedance. The reflection coefficient expressed in decibels is called return loss.

The input impedance determines how large a voltage must be applied at the antenna input terminals to obtain the desired current flow and hence the desired amount of radiated power. Thus, the impedance is equal to the ratio of the input voltage E_i to the input current Ii and it can be written as

$$Z = \frac{E_i}{I_i}. \tag{1.5}$$

Which is in general complex. If the gap in the antenna conductor (feed point) is at a current maximum, and if there is no reactive component to the input impedance, it will be equal to the sum of the radiation resistance and the loss resistance;

that is

$$Z_i = R_i = R_r + R_0. \tag{1.6}$$

If this reactance has a large value, the antenna input voltage must be very large to produce an appreciable input current. If in addition the radiation resistance is very small, the input current must be very large to produce appreciable radiated power. Obviously this combination of circumstances, which occurs with the short dipole antenna that must be used at very low frequencies, results in a very difficult feed problem or impedance matching problem, they are usually fed by waveguides rather than by transmission line. The equivalent of input impedance can be defined at the point of connection of the waveguide to the antenna, just as waveguides have a characteristic wave impedance analogous to the characteristic impedance of a transmission line. For some types of antennas consisting of current carrying conductors this is difficult, and it may even be difficult to define input impedance. This is true, as an example, for an array of dipoles, when each dipole is fed separately; sometimes each dipole, or groups of dipole, will be connected to separate transmitting amplifiers and receiving amplifiers. The input impedance of each dipole or group may then be defined, but the concept becomes meaningless for the antenna as a whole, as does also for simple linear current radiation elements; but they comprise a very large class of antennas.

1.5 Polarization

The polarization of an antenna is defined as the polarization of the electromagnetic wave radiated by the antenna along a vector originating at the antenna and pointed along the primary direction of propagation. The polarization state of the wave is described by the shape and orientation of an ellipse formed by tracing the extremity of the electromagnetic field vector versus time. A brief explanation about polarization will be discussed in the next chapter.

1.6 Principal Patterns

Antenna performance is often described in terms of its principal E and H plane patterns. For a linearly polarized antenna, the E plane pattern is defined as " the plane containing the electric field vector and the direction of maximum radiation" and the H plane as "the plane containing the magnetic-field vector and the direction of maximum radiation."

1.6.1 Radiation Pattern

The radiation pattern describes the relative strength of the radiated field in various directions from the antenna, at a fixed or a constant distance.

Antenna radiation patterns are graphical representations of the distribution of radiated energy as a function of direction about an antenna. Radiation patterns can be plotted in terms of field strength, power density, or decibels. They can be absolute or relative to some reference level, with the peak of the beam often chosen as the reference. Radiation patterns can be displayed in rectangular or polar format as functions of the spherical coordinates θ and ϕ.

An antenna is supposed to be located at the center of a spherical coordinate system, its radiation pattern is determined by measuring the electric field intensity over the surface of a sphere at some fixed distance, R. Since the field E is then a function of the two variables ϕ and θ, so it is written $E(\theta,\phi)$ in functional notation.

A measurement of the electric field intensity $E(\theta,\phi)$ of an electromagnetic field in free space is equivalent to a measurement of the magnetic field intensity $H(\theta,\phi)$, since the magnitudes of the two quantities are directly related by

$$E = \eta_0 \times H. \tag{1.7}$$

(of course , they are at right angles to each other and their phase angles are equal) where η_0 =377 Ω for air. Therefore the pattern could equally be given in terms of E or H. the power density of the field, $P(\theta,\phi)$, can be computed when $E(\theta,\phi)$ is known, the relation being

$$P = \frac{E^2}{\eta_0}. \tag{1.8}$$

Therefore a plot of the antenna pattern in terms of $P(\theta,\phi)$ conveys the same information as a plot of the magnitude of $E(\theta,\phi)$. In some circumstances, the phase of the field is of some interest, and plot may be made of the phase angle of $E(\theta,\phi)$ as well as its magnitude. This plot is called the phase polarization of the antenna. But ordinary the term antenna pattern implies only the magnitude of E or P. Sometimes the polarization properties of E may also be plotted, thus forming a polarization pattern.

If the radiation pattern is plotted in terms of the field strength in electrical units, such as volts per meter or the power density in watts per square meter, it is called an absolute pattern. An absolute pattern actually describes not only the characteristics of an antenna but also those of the associated transmitter, since the absolute field strength at a given point in space depends on the total amount of power radiated as well as on the directional properties of the antenna.

Often when the pattern is plotted in relative terms, that is, the field strength or power density is represented in terms of its ratio to some reference value. The reference usually chosen is the field level in the maximum field strength direction. This type of pattern provides as much information about the antenna as does an absolute pattern, and therefore relative patterns are usually plotted when it is desired to describe only the properties of the antenna, without reference to an associated transmitter (or receiver).

It is also fairly common to express the relative field strength or power density in decibels. This coordinate of the pattern is given as $20 \log (E / E_{Max})$ or $10 \log(P / P_{Max})$. The value at the maximum of the pattern is therefore zero decibels, and at other angles the decibel values are negative.

Finally, we should mention that the antenna patterns are usually given for the free space condition, it being assumed that the user of the antenna will calculate the effect of ground reflection on this pattern for the particular antenna height and ground conditions that apply in the particular antenna height and ground conditions that apply in the particular case. Some types of antenna are basically dependent on the presence of the ground for their operation, for example, certain types of vertical antennas at low frequencies. The ground is in fact an integral part of these antenna systems. In these cases, the pattern must include the effect of the earth.

1.6.2 Radiation pattern lobes

Various part of a radiation pattern are referred to as lobes, which may be sub classified into major, minor, side, and back lobes.

A radiation lobe is a "portion of the radiation pattern bounded by regions of relatively weak radiation intensity." Figure 1.1 demonstrates a symmetrical three-dimensional polar pattern with a number of radiation lobes. Some are of greater radiation intensity than others, but all are classified lobes. Figure 1.2 illustrates a linear two dimensional pattern where the same pattern characteristics are indicated.

A major lobe (also called main beam) is defined as "the radiation lobe containing the radiation of maximum radiation." In figure1.1 the major lobe is pointing in the $\theta = 0$ direction. In some antennas, such as split-beam antennas, there may exist more than one major lobe.

A minor lobe is any lobe except a major lobe.

A side lobe is "a radiation lobe in any direction other than the intended lobe." Usually a side lobe is adjacent to the main lobe and occupies the hemisphere in the direction of the main beam.

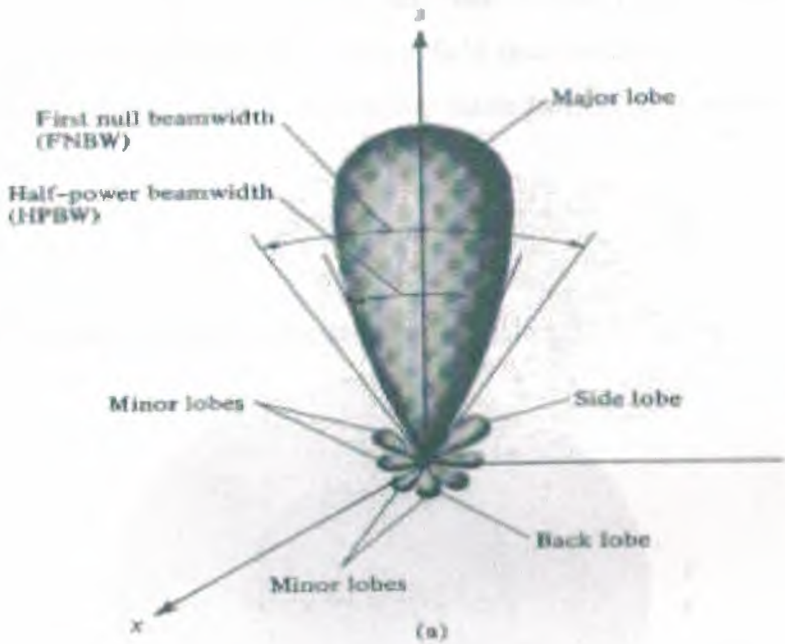


Figure 1.1: Radiation lobes and beamwidths of an antenna.

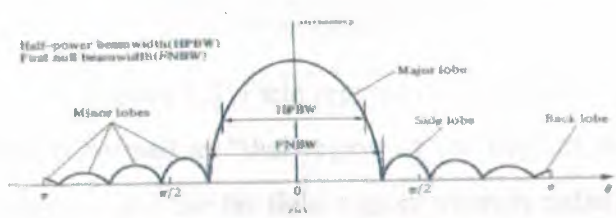


Figure 1.2: Linear plot of power pattern and its associated lobes and beamwidths

A back lobe usually refers to a minor lobe that occupies the hemisphere in a direction opposite to that of the major lobe.

Minor lobes usually represent radiation in undesired directions, and they should be minimized. Side lobes are normally the largest of the minor lobes. The level of minor lobes is usually expressed as a ratio of the power density in the lobe in question to that of the major lobe. This ratio is often termed the side lobe ratio or side lobe level. Side lobe levels of -20 dB or smaller are usually not very harmful in most applications. Attainment of a side lobe level smaller than -30 dB usually requires very careful design and construction.

1.6.3Near and Far Field Patterns

The space surrounding an antenna is usually subdivided into three regions:

(a) reactive near field, (b) radiating near field and (c) far field regions.

Reactive near field region is defined as "that region of the field immediately surrounding the antenna wherein the reactive field predominates." For most antennas, the outer boundary of this region is commonly taken to exist at a distance R from the antenna surface.

$$R < 0.62\sqrt{\frac{D^3}{\lambda}}. \tag{1.9}$$

where λ is the wavelength and D is the largest dimension of the antenna.

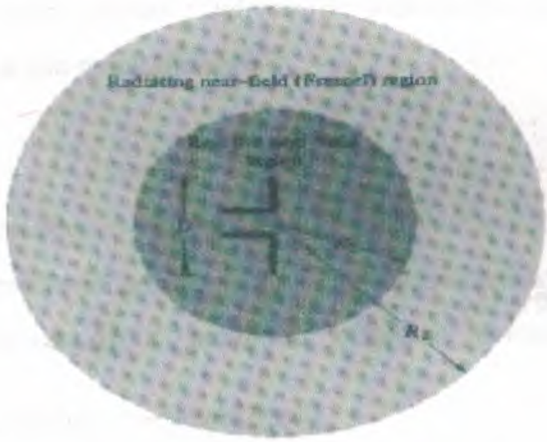


Figure 1.3: Field regions of an antenna.

Radiating near field is defined as "that region of the field of an antenna between the reactive near field region and the far field region wherein radiation fields predominate and wherein the angular field distribution is dependent upon the distance from the antenna. For an antenna focused at infinity, the radiating near field region is sometimes referred to as the Fresnel region on the basis of analogy to optical terminology. If the antenna has a maximum overall dimension, which is very small compared to the wavelength, this field region may not exist.

The inner boundary is taken to be the distance

$$R \geq 0.62\sqrt{\frac{D^3}{\lambda}}. \qquad (1.10)$$

and the outer boundary the distance

$$R < 2\frac{D^2}{\lambda}. \qquad (1.11)$$

where D is the largest dimension of the antenna. In addition D must be large compared with the wavelength.

Far field region is defined as "that region of the field of an antenna where the angular field distribution is essentially independent of the distance from the antenna. If the antenna has a maximum overall dimension D, the far field region is sometimes referred to as the Fraunhofer region on the basis of analogy to optical terminology." In this region, the field components are essentially transverse and the angular distribution is independent of the radial distance where the measurements are made. The inner boundary is taken to be the radial distance

$$R < 2\frac{D^2}{\lambda}.$$

and the outer one at infinity.

To illustrate the pattern variation as a function of radial distance, in Figure iii we have included three patterns of a parabolic reflector calculated at distances of $.R = 2\frac{D^2}{\lambda}$, $4\frac{D^2}{\lambda}$ *and* infinity.

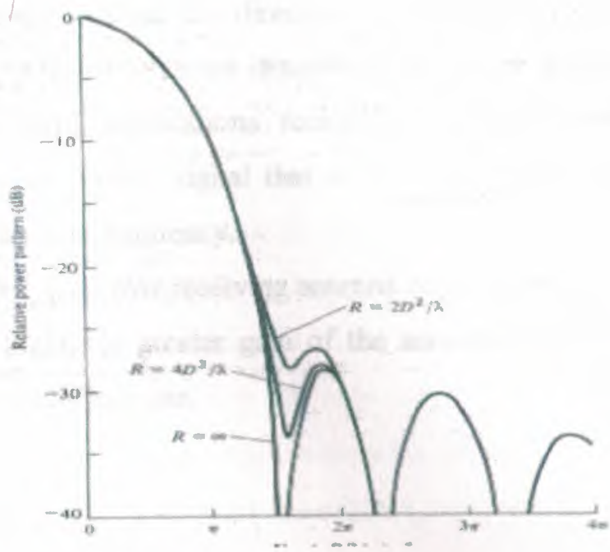


Figure 1.4: Calculated radiation patterns of a paraboloid antenna for different distances from the antenna.

It is observed that the patterns are almost identical, except for differences in the pattern structure around the first null and at a level below 25 dB. Because infinite distances are not realizable in practice, the most commonly used criterion for minimum distance of far field observation is $2D^2/\lambda$.

1.7 Beamwidth

When the radiated power of an antenna is concentrated into a single major lobe, the angular width of this lobe is the beamwidth. It is logical to define the width of a beam in such a way that it indicates the angular range within which radiation of useful strength is obtained, or over which good reception may be expected. From this point of view the convention has been adopted of measuring beamwidth between the points on the beam pattern at which the power density is half the value at the maximum. In a plot of the electric intensity is equal to 0.707 of the maximum value. The angular width of the beam between these points is called the half-power beamwidth, when a beam pattern is plotted with the ordinate scale in the minus 3dB points. For this reason the half power beamwidth is often referred to as the -3dB beamwidth on a rectangular pattern plot.

If an antenna has a narrow beam and is used for reception, it can be used to determine the direction from which the received signal is arriving, and consequently it provides information on the direction of the transmitter. To be useful for this purpose, the antenna beam must be steerable; that is, capable of being pointed in various directions.

It is intuitively apparent that for this direction-finding application, a narrow beam is desirable and the accuracy of direction determination will be inversely proportional to the beamwidth. In some applications receiving may be unable to discriminate completely against an unwanted signal that is either at the frequency as the desired signal or on nearly the same frequency.

In such a case, pointing a narrow receiving antenna beam in the direction of the desired signal is helpful; resulting in greater gain of the antenna for the desired signal, and reducing gain for the undesired one.

1.8 Antenna Gain

The gain, or power gain, is a measure of the ability to concentrate in a particular direction the net power accepted by the antenna from the connected transmitter. When the direction is not specified, the gain is usually taken to be its maximum value. A brief discussion about antenna gain and measurement of antenna gain will be shown in the next chapter.

1.9 Antenna Size, Feed Line and Insulators

The geometrical size of an antenna is always related to the wavelength of the signal that the antenna must transmit or receive. It ranges from micro miniature to gigantic. Typically, the relevant characteristic size of an antenna is half the wavelength of the signal. The large antennas are used for low frequencies (high wavelength), and vice versa, small antennas are used for high frequencies (low wavelength), but sometimes-large antennas are used at short wavelength (high frequencies) to obtain a highly directional radiation pattern and high gain in a preferred direction. Very small antennas can be used at long wavelength, when efficiency is not important.

Feed lines, usually called transmission line, are used to connect the transmitter or receiver to the antenna. The design of the feed lines and any necessary impedance-matching or power dividing devices associated with it is one of the most important problems in the calculation of antenna design. At the very lowest frequencies the earth is a part of the antenna electrical system. One terminal of the antenna input is a rod driven into the ground or a wire leading to a system of buried conductors, especially if the earth is dry in the vicinity of the antenna. The other terminal is then usually the base of a tower or other vertically rising conductor.

At some higher frequencies, up to 30MHz, the antenna may be a horizontal wire strung between towers, or other supports (from which it is insulated). The feed line is then often a two wire balanced line connected at the center of the antenna, either to the two terminals provided by a gap in the antenna wire (series feed). For upper high frequencies (up to 1 GHz), coaxial feed lines are commonly used. Coaxial line diameters range from a fraction of an inch up to 9 inches or more. Above 1GHz, waveguides are commonly used.

The conducting portions of an antenna not only carry RF currents but also have RF voltages between their different parts and between the conductors and ground. So that to avoid the short-circuiting these voltages, insulators must sometimes be used between the antenna and its supports.

CHAPTER TWO

ANTENNA GAIN MEASUREMENT

2.1 Power Gain

Antenna gain is independent of reflection losses resulting from impedance mismatch.

Any directional antenna will radiate more power in its direction (or directions) of maximum radiation than an isotrope would, with both radiating the same total power. It is intuitively apparent that this should be so, since the directional antenna sends less power in some directions than an isotrope does, it follows that it must send more power in other directions, if the total powers radiated are to be the same. This conclusion will now be demonstrated more rigorously.

If an isotrope radiates a total power P_t and is located at the center of a transparent (or imaginary) sphere of radius R meters, the power density over the spherical surface is shown bellow

$$P_{isotrope} = \frac{P_t}{4\pi \times R^2}(w/m^2). \qquad (2.1)$$

Since the total P_t is distributed uniformly over the surface area of the sphere, which is $\left(4 \times \pi \times R^2\right)$

Imagine that in some way it is possible to design an antenna that radiates the same total power uniformly through one half of the same spherical surface, with no power radiated to the other half. Such a fictitious radiator may be called a semi isotrope. Since the half sphere has a surface area (2π R^2) square meters, the power density is

$$P_{semi\_isotrope} = \frac{P_t}{2 \times \pi \times R^2}\left(w/m^2\right) \qquad (2.2)$$

therefore, we get

$$\frac{P_{semi\_isotrope}}{P_{isotrope}} = 2. \qquad (2.3)$$

The last result shows that at any distance, R, the power density radiated by the semi isotrope is twice as great as that radiated by the isotrope, in the half-shpere within which the semi-isotrope radiates.

In this region, therefore, the semi-isotrope is said to have a directive gain of 2. It is fairly apparent that if the radiation were confined to smaller portions of the total imaginary spherical surface, the resulting directive gain would be greater. For example, if the power Pt uniformly into only on fourth of the spherical surface, the directive gain would be 4, and so on.

2.2 Directive Gain

The directive gain D, of an antenna is defined, in a particular direction, as the ratio of the power density radiated in that direction, at a given distance, to the power density that would be radiated at the same distance by an isotrope in the hemisphere into which it radiates is 2; its directive gain in the other hemisphere (where no power is radiated) is zero.

Thus D of an antenna is defined as a quantity that may be different in different directions. In fact the relative power density pattern of an antenna becomes a directive gain pattern if the power density reference value is taken as the power density of an isotrope radiating the same total power (instead of using as a reference the power density of the antenna in its maximum radiation direction). In this case, we define the direction gain of the antenna as

$$D = \frac{P_{antena}}{P_{isotrope}}. \qquad (2.4)$$

Where P(antenna) is the antenna power density.

Substituting Eqs 1.8 and 2.1 into Eq 2.4 we get

$$D = \frac{4 \times \pi \times R^2 \times E^2}{377 P_t} = \frac{4 \times \pi \times R^2 \times P_{antenna}}{P_t}. \qquad (2.5)$$

Where P_t is the total radiation power.

If Pt represents the input power to the actual antenna rather than the power radiated, G should be substituted for D on the left hand side of this equation, that is, give the power gain rather than the directive gain. The efficiency factor ξ is the ratio of the power radiated by the antenna to the total input power, it is a number between zero to unity, and it connects the direction gain D with the power gain G in

$$G = \xi \times D. \qquad (2.6)$$

This value can be calculated from

$$D_{Max} = \frac{4\times\pi}{\int_0^{2\pi}\int_0^{\pi}[E(\theta,\phi)/E_{Max}]^2 \sin\theta d\theta d\phi}. \qquad (2.7)$$

once the directivity Dmax has been calculated from the relative pattern, the directive gain in any other direction θ, ϕ can also be simply determined from the following relationship

$$D_{(\theta_1,\phi_1)} = D_{Max}\left[\frac{E(\theta_1,\phi_1)}{E_{Max}}\right]^2. \qquad (2.8)$$

2.3 Absolute Field Strength Method

This method of gain measurement is based on Eq 2.5 which is rewritten here for reference.

$$D = \frac{4\times\pi\times R^2\times E^2}{377P_t} = \frac{4\times\pi\times R^2\times P_{antenna}}{P_t}. \qquad (2.9)$$

This method requires an absolute measurement of the field intensity E or power density at distance R from the antenna when it is radiating a total power P_t , the measurement being made in the direction of maximum radiation. If this method is to give the direction of the antenna itself, using Eq2.9 , the measurement must be made under free space propagation condition that is, with no multipath interference due to the earth reflection, or any other factors that modify the free space. Otherwise, we should take the propagation factor F into the consideration,

$$F = \frac{E}{E_d}. \qquad (2.10)$$

Where Ed is the field strength in the free space, and E is the measured field strength. On the other hand, if the measurement is made using Eq 2.9 with the antenna in its operating location, the gain measured is the effective gain of the antenna in combination with its environment. When earth reflection is involved, this gain will depend on the elevation angle of the measuring point, as well as on the antenna height and the reflection coefficient of the earth.

If these factors are known or can be measured, the gain of the antenna by itself can be deduced. If a value of field intensity is actually measured by analysis of the reflection interference effect it may be calculated that the field density is great or less than the value that would have been measured if free space propagation existed, by the propagation factor F, as defined by Eq 2.10 , in term of this factor. Eqation2.9 can be rewritten so that it expressed the free space gain of the antenna even if the field intensity E or the power density P is measured under nonfree space conditions.

$$D = \frac{4\pi R^2 E^2}{377 P_t F^2} = \frac{4\pi R^2 P_{antenna}}{P_t F^2}. \tag{2.11}$$

Equation 2.11 conforms with Eq 2.9 when F=1 (free space). The absolute field intensity E can be measured at low frequencies. At higher frequencies, it is more convenient to make the measurement in terms of the received power P_r . This quantity is related to the receiving antenna capture cross section A_r by

$$P_i = \frac{P_r}{A_r} = \frac{4\pi P_r}{\xi D_r \lambda^2}.. \tag{2.12}$$

This formula can be used only if the effective area A_r of the receiving antenna is known and if the received P_r can be measured.

2.4 Gain by Comparison

Gain may be measured with respect to a comparison or reference antenna whose gain has been determined by other means. A λ/2 dipole antenna or a horn antenna are commonly used as references.

The gain G is then given by

$$G = \frac{P_1}{P_2} = \left(\frac{V_1}{V_2}\right)^2. \qquad 2.13$$

where P_1 is the power received with antenna under test, P_2 is the power received with reference antenna, V_1 is the voltage received with antenna under test and V_2 is the voltage received with reference antenna.

It is assumed that both antennas are properly matched. If both are also lossless and the reference is a λ/2 dipole, the gain G_0 over a lossless isotropic source is

$$G_0 = 1.64G = 10\log(1.64G) \quad \text{dBi.} \qquad (2.14)$$

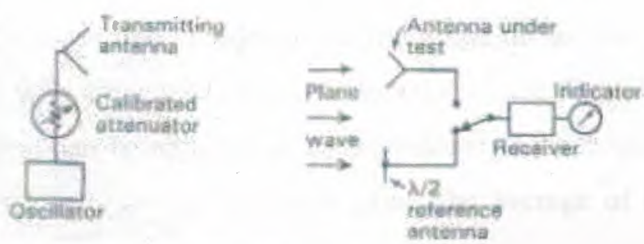


Figure 2-1. Gain measurement by comparision.

The comparison should be made with both antennas in a suitable location where the wave from a distant source is substantially plane and of constant amplitude.

Both antennas may be mounted side by side as in Fig 2.1 and the comparison made by switching the receiver from one antenna to the other. The ratio V_1/V_2 is observed on an output indicator calibrated attenuator so that the received indications the same for both antennas. The ratio P_1/P_2 is then obtained from the attenuator settings.

Mounting both antennas side by side as in Fig 2.1 but in too close proximity may affect the measurements because of coupling between the antennas. To avoid such coupling, a direct substitution may be made with the ideal antenna removed to some distance. If the antennas are of unequal gain, it is more important that the high gain antenna be thus removed.

If the gain of the antenna under test is large, it is often more convenient to use a reference antenna of higher gain than that of a λ/2 dipole. At microwave frequencies electromagnetic horns are frequently employed for this purpose.

Short wave directional antenna arrays, such as used in transoceanic communication, are situated at a fixed height above the ground. The gain of such antennas is customarily referred to either a vertical or a horizontal λ/2 antenna placed at a height equal to the average height of the array. This gain comparison is at the elevation angle of the down coming wave. If the directional antenna is a high gain type and any mutual coupling exists between it and the antenna, the directional antenna can be rendered completely inoperative by lowering it to the ground or sectionalizing its elements when receiving with the λ/2 antenna.

In the above discussion it has been assumed that the antennas are perfectly matched. It is not always practical to provide such matching. This is particularly true with wideband receiving antennas that are only approximately matched to the transmission line. In general, another mismatch may occur between the transmission line and the receiver. In such cases the measured gain is a function of the receiver input impedance and the length of the transmission line. To determine the range of fluctuation of gain of such wideband antennas with a given receiver as a function of the frequency and line length, the length of the line can be adjusted at each frequency to a length giving maximum gain and then to a length giving minimum gain. The average of this maximum and minimum may be called the average gain.

2.5 Gain Measurement by Using Standard Antennas

A gain standard antenna is one whose gain is accurately known so that it can be used in measurement of other antennas. Certain simple forms of antenna can be constructed to have gain of known amount.

Alternatively, a standard antenna can be obtained by a gain measurement, which does not require two antennas that are identical. One is used as a transmitting antenna and the other for receiving, separated by a distance R.

The transmitted power Pt and the receiving power Pr are both measured. The directivity of the antennas can then be calculated by an application of Eq 2.11 and 2.12. If the second expression given for P in Eq 2.12 is substituted into Eq 2.9, then the result is

$$D_t = \left(\frac{4\pi R^2}{P_t F^2}\right)\left(\frac{4\pi P_r}{\xi D_r \lambda^2}\right). \tag{2.15}$$

Where the transmitting antenna directivity denoted by D_t, the quantity Pt has been defined as the radiated power. If now it is instead regarded as the power delivered to the transmitting antenna terminals, D_t must be replaced by $G_t = \xi D_t$, and Dr by $G_r = \xi D_r$. Since it has been stipulated that $G_t = G_r$ and the equation can then be solved for G, the power gain of the two identical antennas

$$G = \frac{41R}{\frac{7}{8}F}\sqrt{\frac{P_r}{P_t}}. \qquad (2.16)$$

This procedure is likely to be successful when F is approximately equal to one, that is, under effectively free space conditions or no earth reflection interference effects.
It can also be applied successfully under conditions that permit accurate calculation of F, as an example, when reflection occurs from a smooth water surface between the two antennas.

2.6 Absolute Gain of Identical Antennas.

The gain can also be measured by a so called absolute method in which two identical antennas are arranged in free space as in Fig 2.2 One antenna acts as a transmitter and the other as a receiver. By the Friis transmission formula

$$\frac{P_r}{P_t} = \frac{A_{er}\, A_{et}}{\lambda^2\, r^2} \quad \text{(dimensionless)}. \qquad (2.17)$$

Where P_r is the received power (W), P_t is the transmitted power (W), A_{er} is the effective aperture of receiving antenna (m^2), A_{et} is the effective aperture of receiving antenna (m^2), λ is the wavelength (m) and r is the distance between antennas (m).

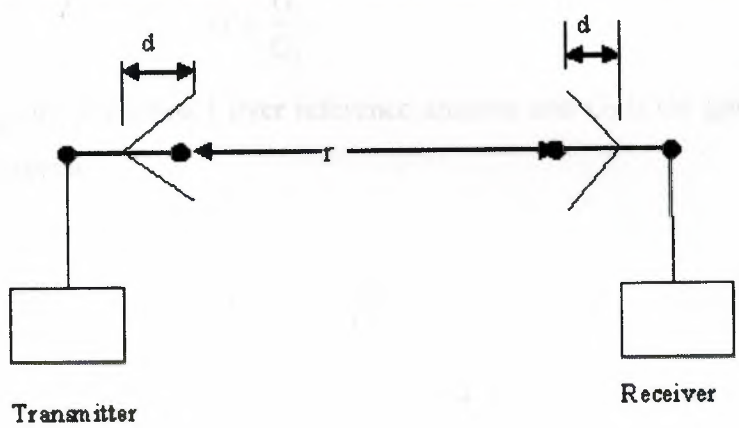


Figure2.2 Absolute gain of identical antennas.

If r is large compared to the depth d of the antenna, the precise points on the antennas between which r is measured will not be critical. Since

$$A_{er} = G_0 \frac{\lambda^2}{4\pi}. \tag{2.18}$$

Where G_0 = gain of antenna over an isotropic source
and since it is assumed that $A_{er}=A_{et}$, Eq (2.17) becomes

$$\frac{P_r}{P_t} = \frac{G_0{}^2\lambda^2}{(4\pi)^2 r^2}. \tag{2.19}$$

and

$$G_0 = \frac{4\pi\,\mathrm{r}}{\lambda}\sqrt{\frac{P_r}{P_t}}. \tag{2.20}$$

Thus, by measuring the ratio of the received to transmitted power, the distance r and the wavelength , the gain of either antenna can be determined. Although it may have been intended that the antenna be identical, they may actually differ in gain by an appreciable amount. The gain measured in this case is

$$G_0 = \sqrt{G_{01}G_{02}}. \tag{2.21}$$

where G_{01} is gain of antenna 1 of the "identical " pair and G_{02} is the gain of antenna 2 of the "identical" pair.

With both gains referred to an isotropic source. To find G_{01} and G_{02}, the above measurement is supplemented by a comparison of each of the antennas with a third reference antenna whose gain need not be known. This gives a gain ratio between "identical" antennas of

$$G' = \frac{G_1}{G_2}. \tag{2.22}$$

where G_1 is the gain of antenna 1 over reference antenna and G_2 is the gain of antenna 2 over reference antenna

Then since

$$G' = \frac{G_1}{G_2} = \frac{G_{01}}{G_{02}}. \tag{2.23}$$

we have

$$G_{01} = G_0\sqrt{G}. \tag{2.24}$$

and

$$G_{02} = \frac{G_0}{\sqrt{G}}. \tag{2.25}$$

2.7 Absolute Gain of single Antenna.

2.7.1 By flat sheet reflector.

By replacing the second antenna of Fig2.2 with a sufficiently large, flat, perfectly reflecting sheet, as in Fig2.3 , the gain of the single(transmission, receiver) antenna is given by (2.20) where r now equals the distance from the antenna to its image behind the reflector. This distance must meet the far_field requirement and this may require a very large flat sheet reflector.

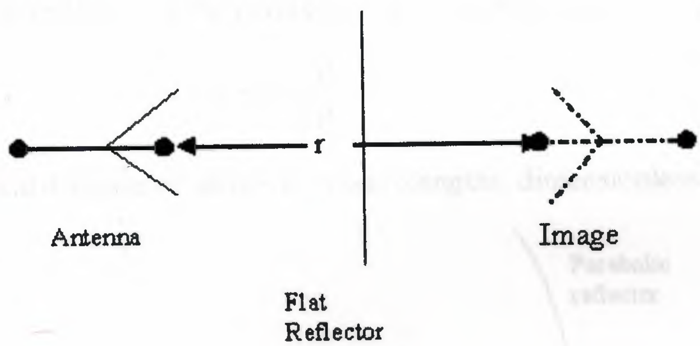


Figure 2.3 Absolute gain of a single antenna by flat sheet reflector method.

2.7.2 By reflecting sphere.

The radar cross section σ of a perfectly reflecting sphere is equal to its physical cross section(π a^2) when its radius a >> λ. With a sphere as the radar target, as in Fig 2.4, we have from the radar equation that the antenna gain

$$G = \frac{8\pi r^2}{\lambda a}\sqrt{\frac{P_r}{P_t}}. \tag{2.26}$$

where r is the distance from antenna to sphere (m) and a is the radius of the sphere (m).

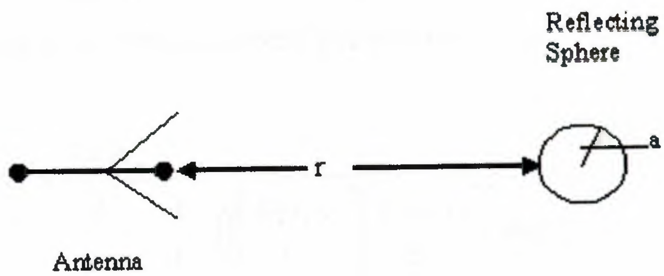


Figure2.4 Absolute gain of a single antenna by reflecting sphere method.

2.7.3.By parabolic reflector.

A more compact configuration involves the use of a parabolic reflector as in Fig 2.5 with the antenna at the focus of the parabola. For this configuration the gain

$$G = 4\pi r_\lambda \sqrt{\frac{P_r}{P_t}} \qquad (2.27)$$

where r_λ is the focal distance of parabola in wavelengths, dimensionless.

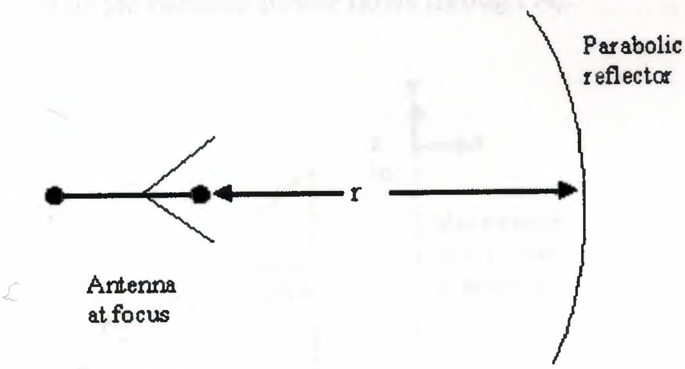


Figure2.5 Absolute gain of a single antenna by parabolic reflected method.

2.8 Gain by Near_Field Measurements.

Referring to Fig 2.6, measurements of the near field of a large antenna with a probe can be used to obtain the gain from Bracewell's relation as

$$G = \frac{4\pi\ A_p}{\lambda^2} \frac{1}{\frac{1}{A_p} \iint_{av} \left[\frac{E(x,y)}{E_{av}}\right]\left[\frac{E(x,y)}{E_{av}}\right]^{*} dxdy}. \tag{2.28}$$

where

E(x,y)=electric field at any point x, y in the aperture, V/m

$E_{av} = \frac{1}{A_p} \iint_{A_p} E(x,y)dxdy = average$ electric field

over the aperture, vm^{-1}

A_p = area of (aperture) plane over

which measurements are made, m^2

It is assumed that all of the radiated power flows through A_p.

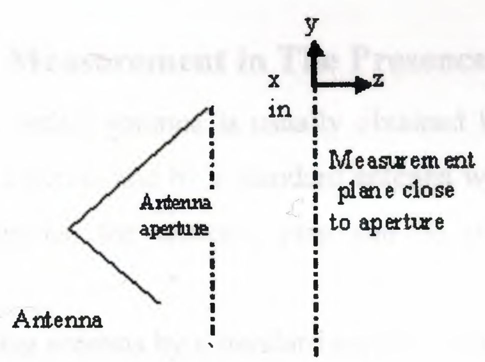


Figure 2.6 Gain by near field method.

This general method is employed by the US Nation Bureau of Standards for gain measurements to an overall accuracy of the order of ±0.2 dB. In addition, far field patterns are obtained using the Fourier transform.

2.9 Gain and Aperture Efficiency from Celestial Source Measurements

For gain measurements using a celestial radio source, an accurate flux density of the source is required and, generally, the source should be essentially unpolarized. Since flux densities are given at only discrete frequencies it may be necessary to interpolate the fluxes at other frequencies.

The effective aperture A_e of an antenna is related to the known flux density S and measured incremental antenna temperature ΔT_A as given by

$$A_e = \frac{2k\,\Delta T_A}{S}. \tag{2.29}$$

From which the gain is

$$G = \frac{4\pi\,A_e}{\lambda^2} = \frac{8\pi\,k\,\Delta T_A}{S\lambda^2}. \tag{2.30}$$

where k is the Bolzman's constant ($1.38*10^{-23}$ JK^{-1}), ΔT_A is the measured source temperature (K), S is the source flux density ($Wm^{-2}\,Hz^{-1}$) and λ is wavelength (m). Thus, knowing S and λ, a measurement of ΔT_A `determines the gain. This measurement includes the effect of any (ohmic) loss in the antenna and any mismatch.

2.10 Antenna Gain Measurement in The Presence of Multipaths .

The antenna gain of a testing antenna is usually obtained by comparing the voltages received by the testing antenna and by a standard antenna with a known gain value. In the presence of multipaths, the antenna gain can be obtained by the following procedure.

1. Replace the testing antenna by a standard antenna with a known gain value.
2. Measure the frequency response. The bandwidth and the frequency points must be the same as those used in measuring the testing antenna. Apply the inverse Fourier transform on the frequency response to obtain the range profile.
3. Record the peak value of the desired path.
4. The gain of the testing antenna is obtained by taking the ratio of the peak value obtained I the case of testing antenna over the peak value obtained in the case of testing antenna over the peak value obtained in the case of standard antenna.

5. Or apply the same window function to retain the desired path and eliminate all other paths, and then take the inverse Fourier transform to obtain the filtered frequency response. The gain of the testing antenna at a certain frequency is obtained by comparing the ratio of the two filtered responses at the frequency.

If the testing antenna is a narrowband antenna, there will be a mismatch between the antenna and the receiver over the bandwidth. It is also noted that the received voltage $V_l(k)$ is a function of the antenna input impedance $Z_{in}(k)$. A mismatch in impedance will reduce the load voltage. If we apply the Fourier transform to the frequency response $V_l(k)$ to obtain the range profile, the range resolution will become poorer and the peak value will decrease. If the bandwidth is too narrow, it may not be able to resolve the desired path and the derived antenna gain value can be inaccurate.

The reflection coefficient | Γ(k) | of a testing antenna usually can be measured. It is known that the power delivered to the receiver will be reduced by a factor of 1- | Γ(k) |^2 due to a mismatched impedance. If we correct the measured $V_l(k)$ with a factor of 1/ [1-| Γ(k) | ^2]^(1/2) and then follow the procedure described in the previous section, we can obtain a more accurate measurement of the range profile and the antenna pattern and the gain value.

2.11 Practical Significance of Power Gain

It is apparent for a given amount of input power in antenna; the power density at a given point in space is proportional to the power gain of the antenna in that direction. Therefore the signal available to a receiving antenna at that location can be increased by increasing the power gain of the transmitting antenna, without increasing the transmitting power. A transmitter with a power output of 1000watts and antenna with a power gain of 10 (10dB) will provide the same power density at a receiving point as will a transmitter of 5oowatts power and an antenna power gain of 20 (13dB) than it would be to double the transmitter power (though in other cases the converse maybe true). But generally speaking it is desired to provide the maximum possible field strength in a particular direction.

CHAPTER THREE
POLARIZATION MEASUREMENT

3.1 Polarization

The polarization of an antenna is defined as the polarization of the electromagnetic wave radiated by the antenna along a vector originating at the antenna and pointed along the primary direction of propagation. The polarization state of the wave is described by the shape and orientation of an ellipse formed by tracing the extremity of the electromagnetic field vector versus time. Although all antennas are elliptically polarized, most antennas are specified by the ideal polarization conditions of circular or linear polarization.

3.2 Wave Polarization

With some antennas it is of interest to measure the nature of the polarization. This may be measured at one frequency as a function of the space angles θ and ϕ. Or it may be measured at one angular position (θ_0,ϕ_0) as a function of the frequency. Such measurements are desirable where the dominant radiation is circularly or elliptically polarized. It is convenient to consider linear polarization and circular polarization as special cases of elliptical polarization.

3.3 Linear Polarization

The electric field vectors for a linearly polarized wave are shown in Figure 3.1a. The magnitude and direction of the electric field E are indicated as a function of distance for a given instant of time. In Fig 3.1b the wave is viewed from the direction of the positive z axis. The electric field E varies in magnitude between positive and negative E, the direction of E being confined to the y direction.

The simplest antennas radiate (and receive) linearly polarized wave. They are usually oriented so that the polarization (direction of the electric vector) is either horizontal or vertical. For example at the very low frequencies it is practically difficult to radiate a horizontally polarized wave successfully because it will be virtually cancelled by radiation from the image of the antenna in the earth, also vertically polarized waves propagate much more successfully at these frequencies (eg, below 1000KHz). Therefore vertical polarization is practically required at these frequencies.

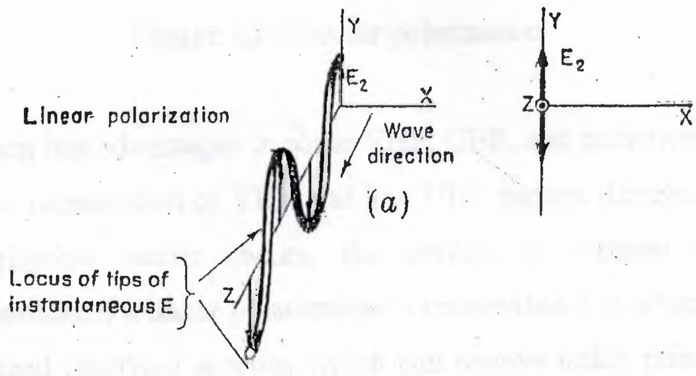


Figure 3.1 Linear polarization.

At the frequencies of television broadcasting (54 to 890 MHz) horizontal polarization has been adopted as standard. The standard frequency is very important to determine the type of polarization. Otherwise, we have to design an antenna such has both polarizations, thus greatly complicating design problem and increasing the received noise level.

At the microwave frequencies (above 1GHz) there is little basis for a choice of horizontal or vertical polarization. Also in specific applications there may be some possible advantages in one or the other. Of course in communication it is essential that the transmitting and receiving antennas have the same polarization.

3.4 Circular Polarization

On the other hand, when $E_1 = E_2$, the ellipse becomes a circle and we have another special case of elliptical polarization called circular polarization. The variation of E for a circularly polarized wave is illustrated by Figure 3.2a and b .

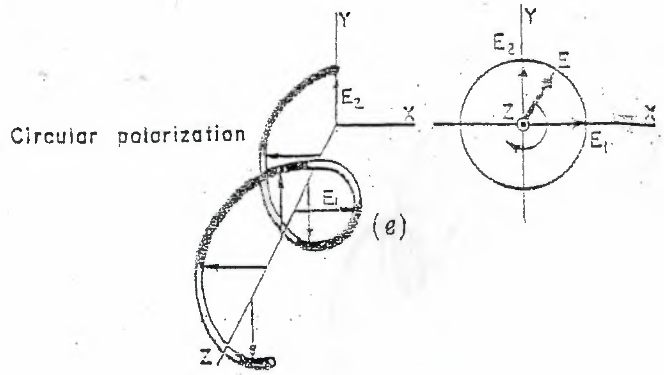


Figure 3.2 Circular polarization

Circular polarization has advantages in some VHF, UHF, and microwave applications . As an example, in transmission of VHF and low UHF signals through the ionosphere, rotation of polarization vector occurs, the amount of rotation being generally unpredictable. Therefore if a linear polarization is transmitted it is advantageous to have a circularly polarized receiving antenna which can receive either polarization, or vice versa. The maximum efficiency is realized if both antennas are circularly polarized.

The ratio of the major axis to the minor axis of the polarization ellipse defines the magnitude of the axial ratio. The tilt angle describes the orientation of the ellipse in space. The sense of polarization is determined by observing the direction of rotation of the electric field vector from a point behind the source. Right-hand and left-hand polarizations correspond to clockwise and counterclockwise rotation respectively.

3.5 Elliptical Polarization

In Figure 3.3a the instantaneous space distribution of E is presented for an elliptically polarized wave traveling in the positive z direction. As viewed from the positive z axis, the tip of the electric field vector E at a fixed position z describes an ellipse with major and minor semi axes E_2 and E_1 as shown in Figure 3.3b. The special case of the linearly polarized wave of Figure 3.1a and b occurs when $E_1 = 0$.

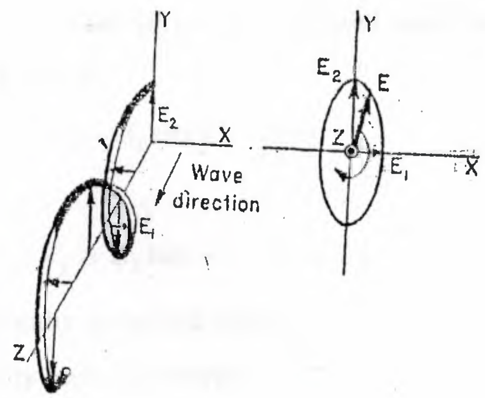


Figure 3.3 Elliptical polarization.

An elliptically polarized wave may be regarded from two points of view:
(1) as the resultant of two linearly polarized waves of the same frequency and (2) as the resultant of two circularly polarized waves of the same frequency but having opposite rotation directions .Both points of view will be discussed, the former being taken up first.

3.5.1 Elliptical polarization as produced by two linearly polarized waves.

In this section an elliptically polarized wave is considered as the resultant of two linearly polarized waves of the same frequency. Assume that both waves are traveling in the positive z direction and that the plane of polarization of one wave is in the x direction and the other in the y direction as in Figure 3.4. If x is horizontal, the wave with E in the x direction may also be called a horizontally polarized wave and the wave with E in the y direction a vertically polarized wave.

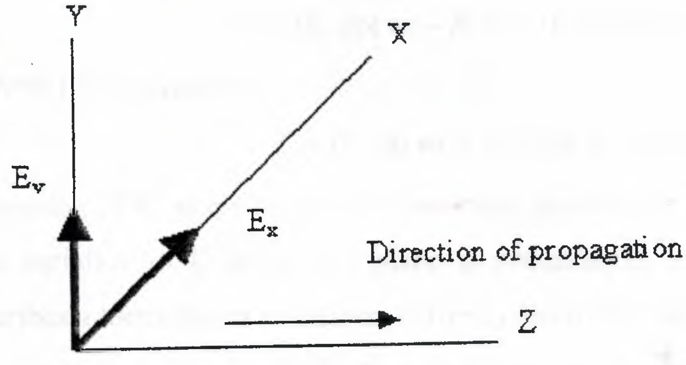


Figure 3-4. Linearly polarized components of an elliptically polarized wave.

Let the instantaneous electric field of the horizontally polarized wave be designated E_x and the instantaneous electric field of the vertically polarized wave be designated as E_y. Then as a function of time and distance,

$$E_x = E_1 \sin(wt - \beta z). \tag{3.1}$$

and

$$E_y = E_2 \sin(wt - \beta z + \delta). \tag{3.2}$$

where E_1 = Amplitude of horizontally polarized wave

E_2 = Amplitude of vertically polarized wave

δ = Time phase angle by which E leads E (the horizontally polarized wave is taken as the reference for phase)

The component of the field in the z direction is everywhere zero (E_z=0).

The instantaneous values of the fields may also be expressed as the imaginary part (Im) of a complex function. Thus,

$$E_x = \operatorname{Im} \dot{E}_x = E_1 \operatorname{Im} e^{j(wt-\beta z)} = E_1 \sin(wt - \beta z) \tag{3.3}$$

and

$$E_y = \operatorname{Im} \dot{E}_y = E_2 \operatorname{Im} e^{j(wt-\beta z+\delta)} = E_2 \sin(wt - \beta z + \delta) \tag{3.4}$$

where

$$\dot{E}_x = E_1 e^{j(wt-\beta z)}. \tag{3.5}$$

and

$$\dot{E}_y = E_2 e^{j(wt-\beta z+\delta)}. \tag{3.6}$$

The instantaneous value of the total field E resulting from the two linearly polarized waves is

$$E = iE_1 \sin(wt - \beta z) + jE_2 \sin(wt - \beta z + \delta). \tag{3.7}$$

At Z=0, (3.7) reduces to

$$E = iE_1 \sin wt + jE_2 \sin(wt - \beta z). \tag{3.8}$$

Evaluating (3.8) as a function of time t and plotting the values of the total field E, the time variation of E in the x-y plane is obtained. In general the tip of the vector E describes a locus that is an ellipse. If E_1=E_2 and δ=90, the ellipse becomes a circle.

The fact that, in general, the locus is an ellipse may be shown in another way by proving that with z=0 are the parametric equations of an ellipse. Thus, we have

$$E_x = E_1 \sin wt. \tag{3.9}$$

and

$$E_y = E_2 \sin(wt + \delta). \tag{3.10}$$

Where wt is the independent variable. The procedure used in the proof will be to eliminate wt and rearrange the resulting expression into the form of the equation for an ellipse. First we expand (3.10).that is,

$$E_y = E_2(\sin wt.\cos\delta + \cos wt.\sin\delta) \tag{3.11}$$

from (3.9)

$$\sin wt = \frac{E_x}{E_1} \tag{3.12}$$

we can also write

$$\cos wt = \sqrt{1-\sin^2 wt} = \sqrt{1-\left(\frac{E_x}{E_1}\right)^2}. \tag{3.13}$$

substituting 3.12 and 3.13 in 3.11 and rearranging and squaring yields,

$$\frac{E_x^2}{E_1^2} - \frac{2E_xE_y\cos\delta}{E_1E_2} + \frac{E_y^2}{E_2^2} = \sin^2\delta. \tag{3.14}$$

which can be reduced to

$$aE_x^2 - bE_xE_y + cE_y^2 = 1. \tag{3.15}$$

where

$a = 1/E_1^2 \sin^2\delta$

$b = 2\cos\delta / E_1E_2 \sin^2\delta$

$c = 1/E_2^2 \sin^2\delta$

Equation (3.15) may be recognized as the equation for an ellipse in its most general form, the axes of the polarization ellipse not, in general, coinciding with the x and y axes (Fig 3-5). This is the general case of elliptical polarization. The line segment OA is the semi major axis, and the line segment OB is the semi minor axis of the ellipse.

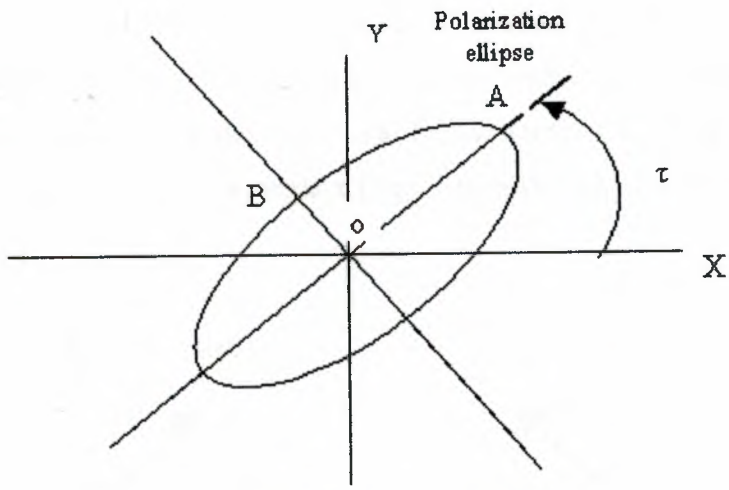


Figure 3-5. Polarization ellipse.

The ratio OA to OB is called the axial ratio (AR) of the polarization ellipse or simply the axial ratio. Thus

$$\text{axial ratio} = \frac{\text{OA}}{\text{OB}}. \tag{3.16}$$

Returning now to, three special cases will be considered.

Case 1. First consider the case where E_y is either exactly in phase or 180 out of phase with E_x. Then $\delta = k\pi$, where k= 0,1,2,3,... and Eq (3-14) Then reduces to

$$\frac{E_x^2}{E_1^2} \pm \frac{2E_xE_y\cos\delta}{E_1E_2} + \frac{E_y^2}{E_2^2} = 0. \tag{3.17}$$

which may be rewritten as

$$\left(\frac{E_x}{E_y} \pm \frac{E_y}{E_2}\right)^2 = 0. \qquad (3.180$$

or

$$E_y = \pm\frac{E_2}{E_1}E_x. \tag{3.19}$$

Equation (3-19) is the equation of a straight line of the form

$$E_y = mE_x. \tag{3.20}$$

Where m = the slope equal to $\pm E_2/E_1$ when k is even (δ=0,2π,4 π, etc.), the slope is positive, and when k is odd (δ= π,3 π,5π,etc.) the slope is negative.

Thus , when the two linearly polarized component waves are exactly in phase or 180 out of phase, the resultant wave is linearly polarized with E, in general, not in the x or y direction. However, if E_2=0, E is in the x direction and the resultant wave is horizontally polarized. If E1=0, E is in the y direction and the resultant wave is vertically polarized. If E_1=E_2 and δ=0, then m= +1 and E is at 45 angle with respect to the positive x axis (Fig 3-6a). if E_1=E_2 and δ= π, then m=-1 and E is at a negative 45 angle with respect to the positive x axis (Fig3-6b).The angle τ (Fig. 3-6a and 3-6b) is related to the slope m by τ = arctan m.

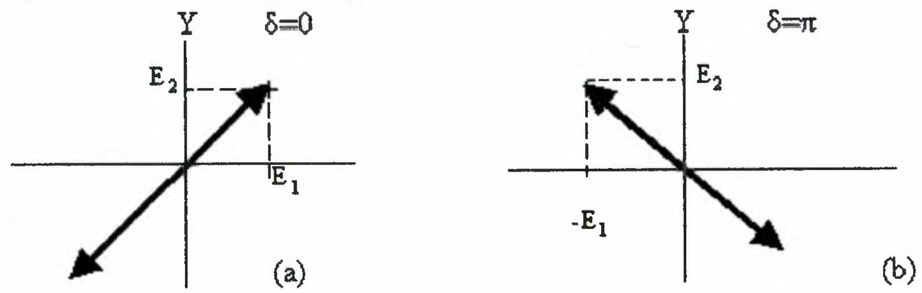


Figure 3.6 Example of linearly polarized waves.

Case 2. Next consider the situation where E_x and E_y are in time phase quadrature. That is,

$$\delta = \frac{1+2k}{2}\pi. \quad (3.21)$$

where k= 0,1,2,3,….
Then the cross-product term in (3-14) disappears and (3-14) reduces to

$$\frac{E_x^2}{E_1^2} + \frac{E_y^2}{E_2^2} = 1. \quad (3.22)$$

this is the standard form of the equation for an ellipse, that is, an ellipse with its axes coincident with the coordinate axes. This is a special case of elliptical polarization. For example if E_2=1/2 E_1 the polarization ellipse is as shown in Fig 3.7.

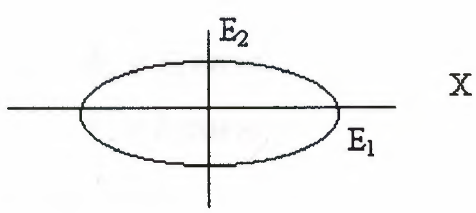


Figure 3.7 Example of elliptically polarized wave.

Case3. Finally consider case 2 for the special condition of $E_1=E_2$. Then (3-22) becomes

$$E_x^2 + E_y^2 = E_1^2. \tag{3.23}$$

This is the equation of a circle (Fig3.8) .Hence, when the two linearly polarized component waves are in time phase quadrature and also are equal in amplitude, the resultant wave is circularly polarized.

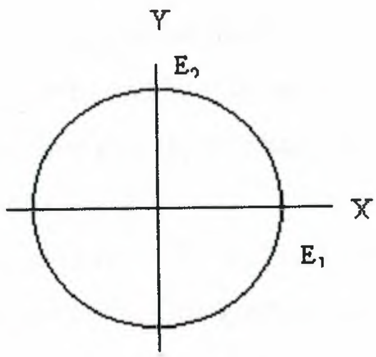


Figure 3.8 Circularly polarized waves

3.6 Clockwise and Counterclockwise Circular Polarization

According to Eq (3-23) the locus of the tip of the vector E is a circle. That is , at a fixed position on the z axis the resultant electric field vector E is constant in magnitude and rotates uniformly with time in the xy plane completing on revolution each cycle. However , gives no information as to the direction in which E rotates, that is , clockwise or counterclockwise, to determine the rotation direction, let us rewrite (3-9) and (3-10) for the special case we are considering, namely,

$$\delta = \frac{1+2k}{2}\pi. \quad \text{and } E_1=E_2 \tag{3.24}$$

where k = 0,1,2,...

Then, when k is even

$$E_x = E_1 \sin wt. \tag{3.25}$$

$$E_y = E_1 \cos wt. \tag{3.26}$$

and when k is odd Ex is the same but

$$E_y = -E_1 \cos wt. \tag{3.27}$$

Consider first the same where k is even ($\delta=\pi/2$, $5\pi/2$, etc). When t=0, Ex =0, and $E_y=+E_1$ so that E is in the positive y direction.

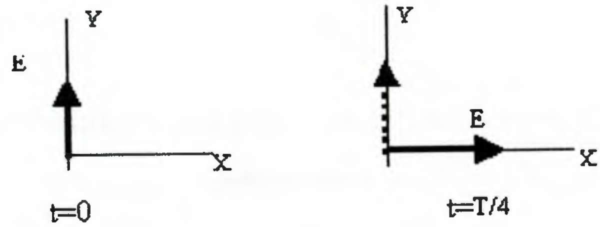


Figure 3.9 Examples of clockwise rotation of E.

One quarter of a cycle later $E_x = +E_1$ and $E_y = 0$ so that E is in the positive x direction. Hence, at a fixed position on the z axis the resultant electric field vector E rotates in a clockwise direction as illustrated in Fig 3.9

Next consider the case for k odd ($\delta = 3\pi/2$, $7\pi/2$, etc). When t=0, $E_x = 0$, and $E_y = -E_1$ so that E is in the negative y direction. One quarter cycle later $E_x = +E_1$ and $E_y = 0$ so that E is in the positive x direction. Hence, at a fixed position on the z axis the resultant electric field vector E rotates in a counterclockwise direction as illustrated in Fig3.9 . the wave is traveling in the positive z direction(out of page) in both this case and the one illustrated by Fig 3.9 . To avoid any uncertainly as to the wave direction, we can call the first case Fig 3.10.

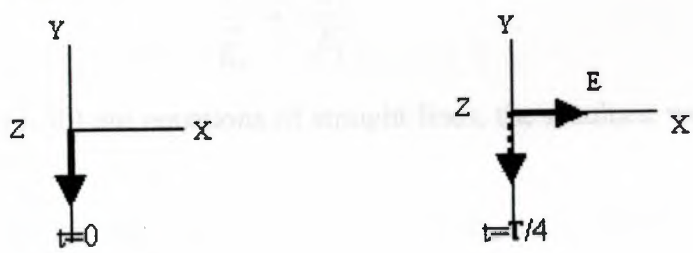


Figure 3.10 Example of counterclockwise rotation.

"clockwise circular polarization wave approaching" and the second case (Fig 3.10) "counterclockwise circular polarization wave approaching"

if the electric vector appears to rotate clockwise with the wave approaching, the electric vector of the same wave appears to rotate counterclockwise when the wave is viewed from the opposit direction, that is , with the wave receding from the observer. Hence, we may say that " clocksise circular polarization wave approaching " is the same as " counterclockwise circular polarization wave receding. "

according to the usage of classical physics, " clockwise circular polarization wave approaching" is called " right circular polarization" however, according to the IRE standards " clockwise circular polarization wave receding " is called " right circular polarization".

3.7 Clockwise and counterclockwise Elliptical Polarization.

In the general situation where the resultant wave is elliptically polarized, it is also of interest to know the direction of rotation of E. This can be determined by plotting E for several instants of time as calculated from Ez and Ey in (3-9) and (3-10). Or we can proceed in the following manner. Divide (3-6) by (3-5) obtaining

$$\frac{\dot{E}_y}{\dot{E}_x} = \frac{E_2}{E_1} e^{j\delta}. \qquad (3.28)$$

Equation (3-28) will now be applied to several special cases as illustrations.

Case 1. when E_y and E_x are in phase, δ=0. then (3-28) becomes

$$\frac{\dot{E}_y}{\dot{E}_x} = +\frac{E_2}{E_1}. \qquad (3.29)$$

when E_y and E_x are 180 degree out of phase, $\delta = \pi$ then(3-28) becomes

$$\frac{\dot{E}_y}{\dot{E}_x} = -\frac{E_2}{E_1} \qquad (3.30)$$

both (3-29) and (3-30) are equations of straight lines, the resultant wave being linearly polarized.

Case 2. Next consider the situation where E_y leads Ex by 90 or $\delta=\pi/2$. Then 3.30 reduces to

$$\frac{\dot{E}_y}{\dot{E}_x} = +j\frac{E_2}{E_1}. \qquad (3.31)$$

This is the case of clockwise elliptical polarization (wave approaching). The axial ratio of the polarization ellipse is in this instance E_2/E_1. if the axial ratio is unity ($E_2=E_1$), then

$$\frac{\dot{E}_y}{\dot{E}_x} = +j. \tag{3.32}$$

This is the case of clockwise circular polarization (wave approaching). It should be noted that the ratio E_2/E_1 equals the axial ratio only when δ =+ or – $\pi/2$.

Case 3. Finally consider the situation where E_y lags E_z by 90 or δ=-$\pi/2$. Then 15.49 becomes

$$\frac{\dot{E}_y}{\dot{E}_x} = -j\frac{E_2}{E_1}. \tag{3.33}$$

this is the case of counter clockwise elliptical polarization. When E_2= E_1 Eq 3.33 reduces to

$$\frac{\dot{E}_y}{\dot{E}_x} = -j. \tag{3.34}$$

This is the case of counterclockwise circular polarization(wave approaching). Thus , from Cases 2 and 3 we can conclude that a+j indicates clockwise rotation while a-j indicates counterclockwise rotation of E.

3.8 Polarization as a function of E_2/E_1 and δ

In the previous sections we have seen that the ratio E_2/E_1 and the phase angle determine the type of polarization of the resultant wave produced by two linearly polarized component waves (with their planes of polarization at right angles). The polarization ellipses for E of the resultant wave as a function of E_2/E_1 and δ are presented in Fig. 3.11 for E_2/E_1 values of ∞, 2, 1, 0.5, and 0 and δ values of 0, ±45, ±90, ±135 and ±180 degrees. The direction of rotation of E is indicated. It is clockwise for positive values of δ and counterclockwise for negative values of δ.

Referring to Fig.3.11, the resultant wave is linearly polarized and vertical for all values of δ when E_2/E_1 =∞, that is, when E1 =0. When E_2/E_1 =0,that is, when E2=0,the wave is linearly polarized and horizontal for all values of δ. The wave is also linearly polarized when δ=0 or ±180, the plane of polarization (horizontal, slant, or vertical)

Depending on the ratio E_2/E_1. Circular polarization occurs only for the case where E_2/E_1 =1 and δ = ±90. When δ =+90, the direction is clockwise (wave approaching, and when δ=-90, the rotation direction is counterclockwise (wave approaching). All these situations are special limiting cases of the general situation in which the wave is elliptically polarized. In Fig3.11 there are 16 cases of elliptical polarization.

In Fig3.11 we note that for a given value of E_2/E_1 all polarization ellipses are contained within a rectangle of height to width ratio equal to E_2/E_1. For $E_2/E_1 = 0$ or ∞ the rectangle degenerates to a line.

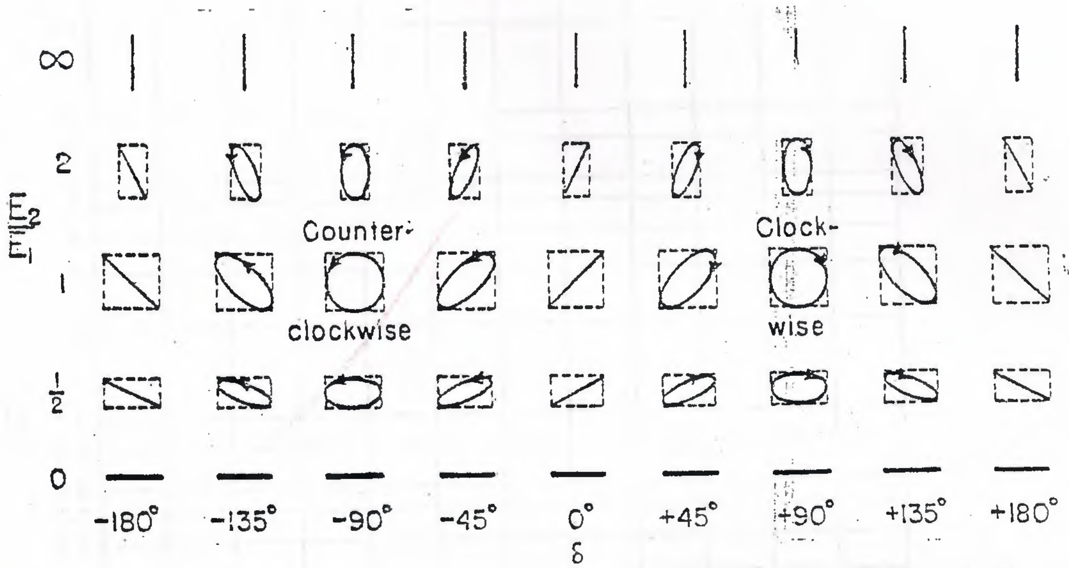


Figure 3.11 Chart of polarization ellipses as a function of the ratio E_2/E_1 and phase angle δ (wave approaching).

Two linearly polarized antennas oriented at right angles and energized with equal voltages in phase quadrature are sometimes employed to produce circular polarization. If the voltages are unequal or the phase relation is not 90, the polarization becomes elliptical. By means of polarization measurements of the radiated wave, it is possible to determine what adjustments should be made on the antenna to obtain circular polarization. For example, suppose that one of the linearly polarized antennas is vertical and the other is horizontal. Then if the polarization is elliptical, with the major axis of the polarization ellipse either vertical or horizontal, the phasing is ±90 but the two antennas are radiating unequal powers. If the major axis of the polarization ellipse is at , it indicated that tee two antennas are radiating the same power but the phase is not . for other ellipses, the power division and phasing can be estimated with the aid of Fig 3.11.

To present wave polarization data, a chart with coordinates similar to those in Fig.3.11 is useful. A chart of this type is presented in Fig 3.12. the ordinate is the ratio E_2/E_1, and the abscissa is the phase angle .

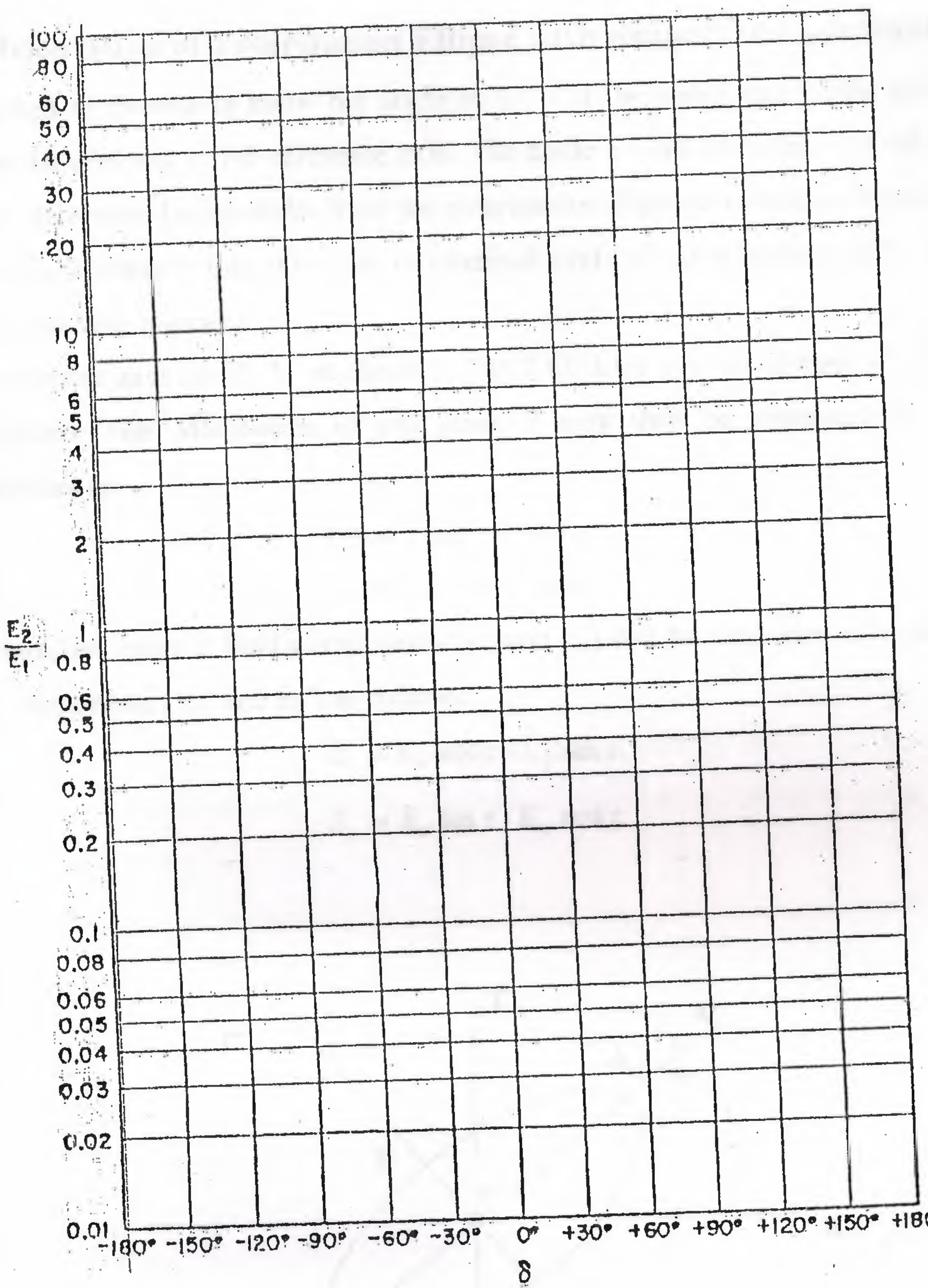


Figure 3.12 wave polarization chart.

A point on the chart defines the polarization uniquely. Thus, the point $E_2/E_1 = 1$ and corresponds to clockwise circular polarization. If the polarization of an antenna is observed to change as a function of frequency, this variation can be plotted as a line on the chart of Fig 3.12. the values of $E_2/E_1 = 1$ and $\delta = +90$ corresponds to clockwise circular polarization. If the polarization of an antenna is observed to change as a function of frequency, this variation can be plotted as a line on the chart of Fig 3.12. The values of E_2/E_1 and δ can also be conveniently presented on the charts of Fig 3.21.

3.9 Orientation of Polarization Ellipse with Respect to Coordinates

It is often of interest to know the angle of tilt τ of the major axis of the polarization ellipse with respect to the reference axis. The angle τ will be called the tilt angle. It may be determined graphically from the polarization ellipse as evaluated from (3-5) and (3-6) as a function of time. Or τ can be obtained explicitly as a function of E_1 , E_2 and δ in the following manner.

The reference axes are X, Y as shown in Fig.3.13. let a new set of axes X', Y' also be constructed. The coordinates of any point P may then be expressed in the new coordinates as

$$x = x' \cos\tau - y' \sin\tau \tag{3.35}$$

$$y = x' \sin\tau + y' \cos\tau \tag{3.36}$$

therefore, the electric field components (E_x and E_y) can be expressed in terms of new field components (E_x' and E_y') as follows,

$$E_x = E_{x'} \cos\tau - E_y \sin\tau. \tag{3.37}$$

$$E_y = E_{x'} \sin\tau - E_y \cos\tau. \tag{3.38}$$

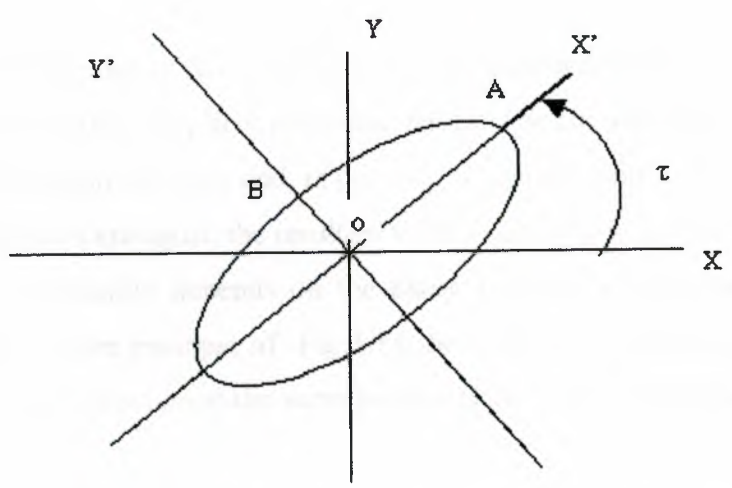


Figure 3.13 Construction for finding the angle τ between the x axis and the major or minor axis of the polarization ellipse.

Now substituting (3.37) and (3.38) into (3-14) and solving for τ yields to

$$\tau = \frac{1}{2}\arctan\frac{2E_1E_2\cos\delta}{E_1^2 - E_2^2}. \tag{3.39}$$

The tilt angle τ between the major or minor axis of the polarization ellipse and the positive x axis can be calculated from a knowledge of the phase angle δ and the amplitudes E1 and E2 in the x and y directions. The angle τ as given by 15.63 is the angle between the x axis and either the major or the minor axis of the ellipse.

For equal in_phase component fields $\delta = 0$ and $E_2=E_1$, we find from (3-39) that $\tau = 45$. This is the case of linear polarization at a slant (45) angle. From (3-39) it is also apparent that $\tau = \pm 45$ when $E_2=E_1$ for all values of δ.

3.10 Cross and Co polarization

When describing the polarizations over the radiation sphere, or portion of it, reference lines shall be specified over the sphere, in order to measure the tilt angles of the polarization ellipses and the direction of polarization for linear polarizations. At each point on the radiation sphere the polarization is usually resolved into a pair of orthogonal polarizations, the co polarization and the cross polarization. To accomplish this, the co-polarization must be specified at each point on the radiation sphere.

3.11 Elliptical Polarization as Produced by Two Circularly Polarized Waves.

In this section an elliptically polarized wave will be regarded from the point of view that it is the resultant of two circularly polarized waves. The circularly polarized waves are of the opposite rotation direction and, in general, of unequal amplitude.

When the amplitudes are equal, the resultant wave is linearly polarized (Fig3.14).

The plane of polarization depends on the phase relation between the two circularly polarized waves. In the example of Fig 3.14 the E vectors for both component waves are in the positive y direction at the same instant. Hence, the resultant wave is vertically polarized.

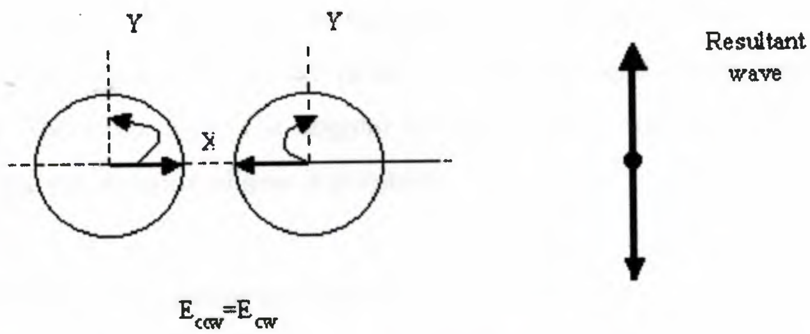


Figure 3.14 Circularly polarized components of linearly polarized waves.

When the amplitudes are unequal, the resultant wave is elliptically polarized. If, for example, the counterclockwise rotating wave has twice the amplitude of the clockwise rotating wave ($E_{ccw}=2E_{w}$). The resultant wave is elliptically polarized as illustrated in Fig 3.15. Since the E vector of both waves is in the positive y direction at the same instant, the major axis of the polarization ellipse is vertical. The rotation direction is counterclockwise, the same as for the larger component wave.

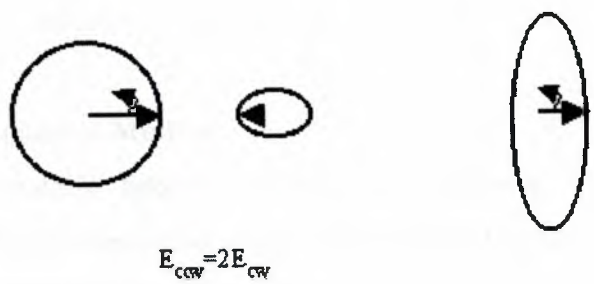


Figure 3.15 Circularly polarized components of elliptically polarized waves.

If one of the component waves becomes zero , we have a limiting case and the resultant wave is equal to the other component wave(Fig 3.16).

The resultant of two circularly polarized waves (of opposite rotation direction) is , in general, an elliptically polarized wave.

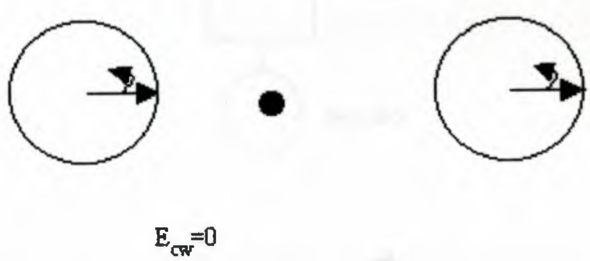


Figure 3.16 Circularly polarized components of circularly polarized waves.

The angular velocity of E for an elliptically polarized wave is smaller when E is in the direction of the major axis of the polarization ellipse and larger when it is in the direction of the minor axis. The angular velocity is such that the rate of sweeping out the area of the polarization ellipse is constant.

3.12 Polarization measurements:

Three methods by which the polarization characteristics of a wave can be measured are:

1. By measuring the polarization pattern with a linear antenna and also observing the direction of rotation of E. This will be called the polarization pattern method
2. By measuring the amplitudes (E_1 and E_2) of two perpendiculars linearly polarized components of the wave and the phase angle δ between them. This will be called the linear component method.
3. By measuring the amplitudes (E_3,E_4) of the two circularly polarized components (of opposite rotation direction) of the wave and the phase angle δ' between them. This will be called the circular component method.

3.11.1 Polarization pattern Method.

In this method a rotatable linearly polarized antenna, such as the half wavelength antenna in Figure 17, is connected to a receiver calibrated to read relative field intensity. Let the wave be approaching (out of page).

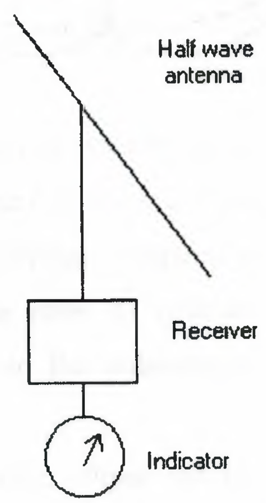


Figure 3.17 Shematic arrangement of a rotatable linearly polarized antenna for measuring polarization pattern.

Then as the antenna is rotated in the plane of the page, the field intensity observed at each position is proportional to the maximum component of E in the direction of the antenna. Such measurements of the incident wave with a rotatable linearly polarized antenna do not yield the polarization ellipse of the wave but rather its polarization pattern. Thus if the tip of the electric vector E describes the polarization ellipse shown in Fig 3.18 .(dashed curve), the variation measured with a linearly polarized receiving antenna is given by the polarization pattern in figure. For a given orientation OP of the linearly polarized antenna, the response is proportional to the greatest ellipse dimension measured normally to OP. as shown in fig. This is the length OP'. If the linearly polarized antenna orientation is OQ, the response is proportional to the length OQ'. For the case of linear polarization, the polarization ellipse degenerates to a straight line and the corresponding polarization pattern is a figure of eight as indicated in gig. By graphical construction as in fig, the polarization ellipse ca be constructed if the polarization pattern is known or vice versa. To determine the direction of rotation of E an auxiliary measurement is necessary.

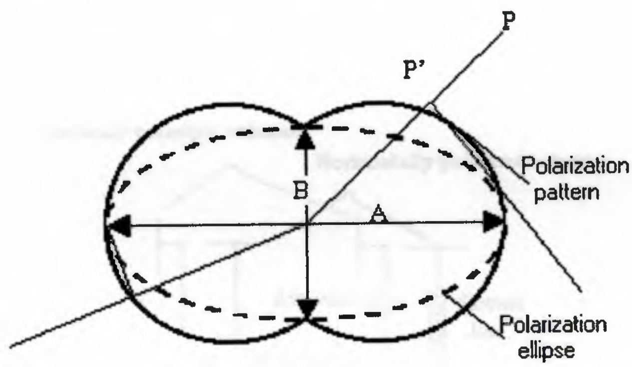


Figure 3.18 Schematic arrangement of linearly polarized antennas for measuring ratio E_2/E_1 and for measuring phase angle δ in linear component method.

For example, the output of two circular polarized antennas could be compared, one responsible to clockwise and the other to counterclockwise rotation. The rotation direction of E then corresponds to the polarization of the antenna with the larger response.

Thus, by this method the polarization ellipse can be drawn and the rotation direction indicated (F2ig 3.18). Although such a diagram completely describes the polarization characteristics of a wave, it is simpler to measure merely the maximum amplitude A/2 and the minimum amplitude B/2 and take the ration of the two amplitudes which, is called the axial ratio of the polarization ellipse or simply the axial ratio(AR). The axial ratio is usually expressed so that it is equal to or greater than unity. The axial ratio of the polarization ellipse of Fig 3.17

$$AR = A/B \tag{3.40}$$

Thus by specifying AR, τ and the rotation direction of E the polarization characteristics are completely described.

3.11.2 Linear component method.

In this method two fixed linearly polarized antennas can be mounted at right angles, like the two half wavelength antennas in Fig 3.17. The wave is approaching normally out of the page. By connecting the receiver first to the terminals of one antenna and the other, as in Fig 3.19a, The ratio E_2/E_1 can be measured. Then, by connecting both antennas to a phase comparator, the angle δ can be measured. This may be done as in Fig 3.19b , using a matched slotted line. From knowledge of E_2, E_1 and δ the polarization ellipse can be calculated from Eq 3.14 , Eq 3.9 and 3.10 and the direction of rotation of E determined from 3.9 and 3.10 and the chart of E_1/E_2 and δ can be plotted on the charts of Fig 3.12 or 3.21.

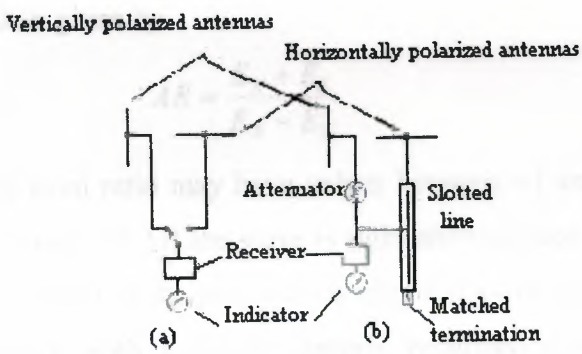


Figure 3.19 Shematic arrangement of linearly polarized antennas for measuring ratio E_2/E_1 (a) and for measuring phase angle δ in linear-component method (b).

3.11.3 Circular component method

In this method two circularly polarized antennas of opposite rotation direction are connected successively to the receiver and the amplitudes E3 and E4 of the circularly polarized component waves measured. The antennas can very conveniently consist of two long helical beam antennas one wound left handed and the other wound right handed as in Fig 3.20. The left handed helix responds to left circular polarization and the right handed helix to right circular polarization. The left circular component E_l of the wave is measured with the switch to the left as in Fig 3.20 so that the receiver is connected to the left handed helix.

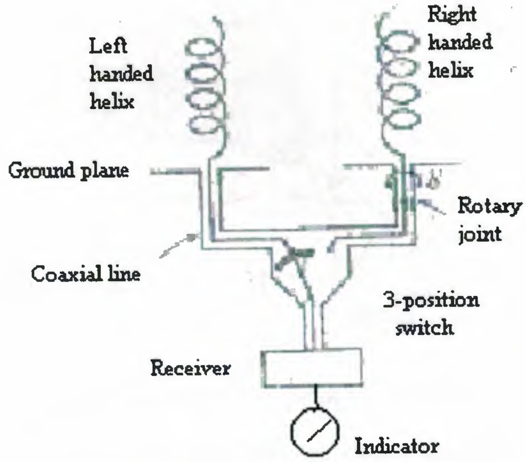


Figure 3.20 Arrangement for measuring left and right circular components of wave and phase angle δ' between them in circular component method.

The right circular component E_r of the wave is measured with the switch thrown to the right so that the receiver is connected to the right handed helix, the axial ratio (AR) of the receiver wave is then given by

$$AR = \frac{E_R + E_L}{E_R - E_L} \tag{3.41}$$

According to 3-41 the axial ratio may have values between +1 and +∞ and between -1 and – ∞. For positive values of AR the wave is right elliptical and for negative values is left elliptical. The tilt angle τ of the polarization ellipse may be measured by finding the direction of maximum E with a rotable linearly polarized antenna. Or τ may be determined with the helical antennas of Figure 3.20 by rotating one helix o its axis with both helices connected in parallel to the receiver. Assuming that the axes of the helices are in a horizontal plane, let the helix rotate angle be δ'and let its reference point (δ'=0) be taken when the receiver output is a minimum for a horizontally polarized incident wave. Then for any type of polarization with the polarization ellipse at a tilt angle τ to the horizontal, τ = δ'/2. Thus, three measurements E_l, Eb and δ' with the helical antennas determine the polarization characteristics of the received wave completely.

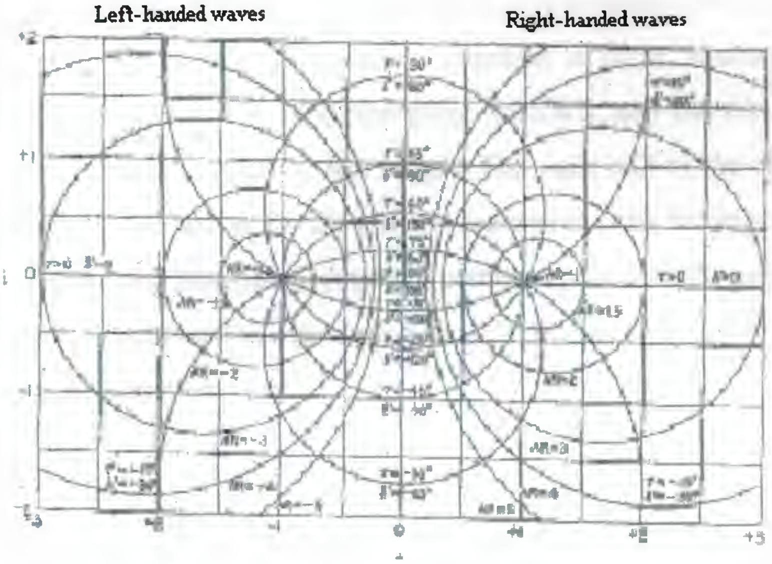


Figure 3.21 Rumsey and Tice type of wave polarization chart.

The circular component method using helical beam antennas is probably the most practical of the three methods, especially for measurements over a considerable frequency range. The accuracy depends on the circularity of polarization of the helices. This is improved (AR nearer unity) by making the helices long since by

$$AR = \frac{2n+1}{2n} \tag{3.42}$$

Where n = the number of turns of the helix.

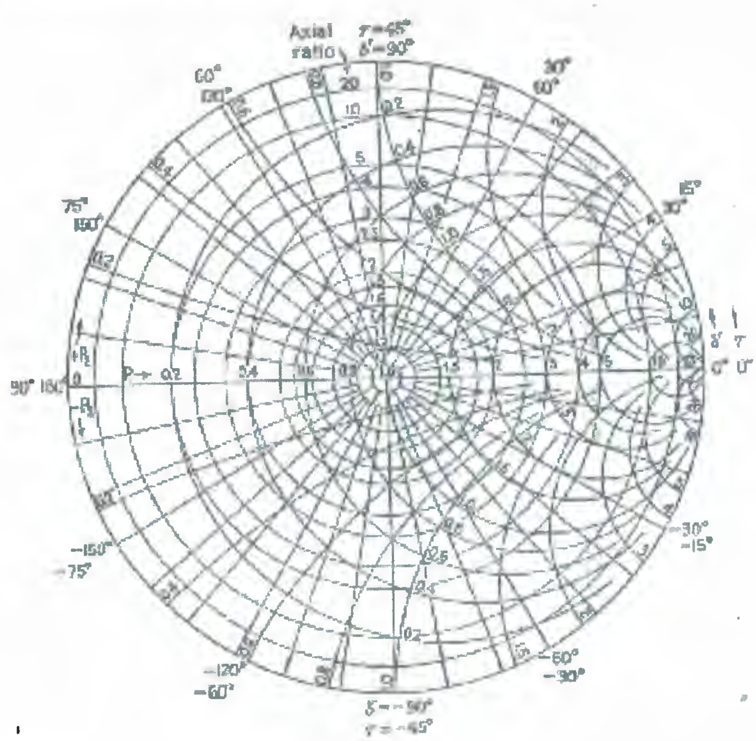


Figure 3.22 Rumsey and Tice type of wave polarization chart.

Rumsey and Tice have devised the very convenient presentation for wave polarization data shown in Fig 3.21. This presentation employs a chart similar to a bipolar impedance chart except that AR takes the place of SWR and the tilt angle τ of the polarization ellipse takes the place of line length. The right half of the chart is for right handed waves and the left half is for the left handed waves. The rectangular coordinates P_1 and P_2 are the real and imaginary parts of a complex polarization parameter P that is related to the linearly polarized components E_1 and E_2 of the wave and the phase angle δ between them by the equation

$$P = P_1 + jP_2 = j\frac{E_1}{E_2}\angle\delta. \tag{3.43}$$

A circle diagram similar to a Smith chart can also be used for this type of presentation as shown in Fig 3.22. Here the chart is limited to either left or to right handed waves unless some convention is adoped as, for example, that measurements of left handed waves be plotted as circles and right handed as crosses. Either of the charts in Fig 3.22 is especially convenient for plotting polarization data measured by the circular component method.

CHAPTER FOUR

MEASUREMENT APPLICATION

4.1 Yagi Antenna

A yagi antenna is essentially a dipole with directors and reflectors added before and behind the dipole. These reflectors and directors are simply metal rods of similar size to the dipole. They concentrate the energies into a "beam" in order to increase the gain. There is usually only one reflector and one or more directors. A yagi antenna is rated by its total number of elements. A yagi antenna may be considered as a directional form of a dipole antenna and the comments on the dipole antenna are applicable

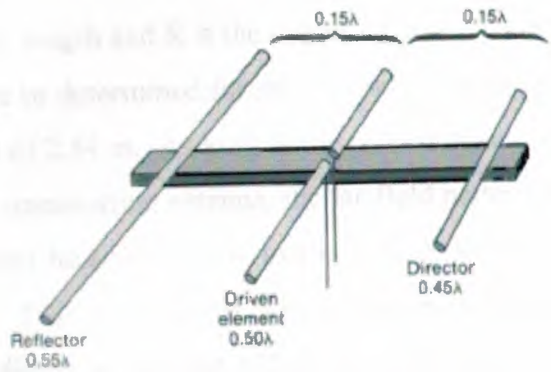


Figure 4.1 Yagi Uda Antenna

If the antenna is mounted for vertical polarization, the condensation drain hole (if present) on one end of the dipole should be at the bottom. If mounted for horizontal polarization, the antenna feeder cable entry to the dipole element should be at the bottom to allow condensation to drain. . Failure to do so may result in eventual damage to the antenna.

Large yagis are best center mounted rather than end mounted. Mounting a vertically polarized antenna centrally is not recommended as the mast is also vertical and will affect the antenna performance. If end mounting a large antenna it is best to use a bracket recommended by the manufacturer to avoid possible degradation of the antenna's characteristics.

4.2 Yagi-Uda Performance

After the theoretical design was completed and an antenna fabricated, testing needed to be done. We tested this in an elevated far-field range. This was done on a golf course to help minimize reflections off of surrounding objects. To minimize reflections off the ground, we lifted the antennas up over our heads with poles.

The first step in this process was to determine when we would be in the far field of the antenna. This was done with equation 4.1.

$$R \geq 2\frac{D^2}{\lambda} \qquad (4.1)$$

where D is the aperture length and R is the distance between antennas.
From equation 6 it can be determined for an antenna operating at 315 MHz, a 1.1 meter long antenna has an R of 2.54 m. As long as the receiving antenna stays this distance or farther away from the transmitting antenna, the far-field pattern is recorded.
The Yagi-Uda was then hooked directly to the radar's VCO and this was used as the transmitting antenna. This provided us with a transmitting power of 9 dBm or 7.94 mW. We set up a dipole at 50 feet (15.24 m) away and used it as the receiving antenna. We were able to read the power received at the dipole with a spectrum analyzer. By rotating the Yagi-Uda by 22.5 degrees and making note of the power received at the spectrum analyzer, we were able to calculate the relative power radiated by our antenna in all directions. With the help of a computer program, the information we obtained was translated into Figures 4.2 and 4.3.

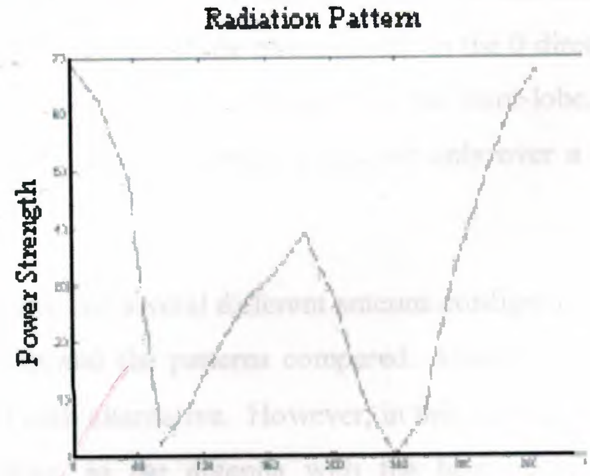


Figure 4.2 Linear Plot of the Normalized Radiation Pattern.

Figure 4.2 shows the radiation pattern of our Yagi-Uda with power received at the spectrum analyzer expressed in decibels on the vertical axis and the angle from the desired direction along the horizontal axis. From this plot we can obtain two pieces of information. First, there are two areas of high radiation, in front of the antenna, at 0 degrees and 360 degrees, and behind the antenna at 180 degrees. This plot also reveals to us that the back-lobe is approximately 30 dB below the front-lobe. This translates into a power level that is 1/1000th of the front-lobe.

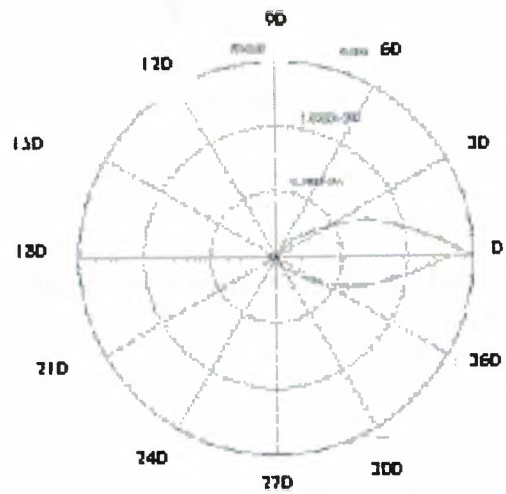


Figure 4.3 Polar Plot of the Normalized Radiation Pattern.

Figure 4.3 gives more insight into the pattern when coupled with Figure 4.2. This plot is a polar plot of the power received at the spectrum analyzer in Watts vs. the angle from the desired direction. Here, it is easier to see how the antenna radiates if you envision the antenna at the center of the plot pointing in the 0 direction. It is also more easily seen how the back-lobe is much smaller then the front-lobe. From this plot, one should notice how the radiation pattern is displayed only over a specific area and no side-lobes are detected.

This process was repeated for several different antenna configurations. Plots were made from each configuration and the patterns compared. Usually, this would be analyzed along with the gain of each alternative. However, in this case, all of the alternatives had similar received powers, so the antenna with the best pattern was selected here. The calculation of the gain from our test setup was much harder, however, because the 70 ohm, dipole antenna was connected to a 300 ohm line. This line lead to a balun, which transformed the 300 ohm line to 70 ohm coax. This 70 ohm coax runs to a 50 ohm input on the spectrum analyzer. This gave us a value for the received power, which is inaccurate for gain but good for relative power. We took measurements of this power received at the spectrum analyzer as we rotated the Yagi-Uda by 22.5 degrees.

To find the actual received power, the following equation was employed:

$$P_r = \frac{P_{read}}{q} \tag{4.2}$$

where q $=q_1 * q_2$ =impedance mismatch of the system. and P_{read} is the power read on the spectral analyzer (W).

$$q_1 = 1 - \left|\frac{Z_1 - Z_2}{Z_1 + Z_2}\right|^2 = \text{impedance mismatch of antenna to line.} \tag{4.3}$$

$$q_1 = 1 - \left|\frac{70 - 300}{70 + 300}\right|^2 = 0.613$$

where Z_1 is the impedance of dipole antenna (70 ohms) and Z_2 is the impedance of line(300 ohms).

$$q_2 = 1 - \left| \frac{Z_3 - Z_4}{Z_3 + Z_4} \right|^2 = \text{impedance mismatch of line to spectral analyzer.} \tag{4.4}$$

$$q_2 = 1 - \left| \frac{70 - 50}{70 + 50} \right|^2 = 0.972.$$

where Z_3 is the characteristic impedance of the coax (70 ohms) and Z_4 is the input impedance of spectral analyzer (50 ohms).

The highest power read for the design chosen on the spectrum analyzer was –27.2 dBm. This correlates to a Pread=1.91x10-6 W. From equation 4.2, we calculate Pr=3.194x10-6 W. Combining this information with the dipole gain of 1.64 and the antenna separation of 15.24 m, the Friis equation, equation 4.5, can be used to calculate gain.

$$P_r = P_t \frac{G_t G_r}{(4\pi \mathrm{R})^2} \lambda^2 \tag{4.5}$$

which can be rewritten as:

$$G_t = \frac{P_r (4\pi \mathrm{R})^2}{P_t G_r \lambda^2} \tag{4.6}$$

This gives us a gain of 9.92 or 9.95 dB. This is very close to the goal antenna gain of 10 dB.

After the operation of the antenna at the operating frequency, the characteristics it displays at the other areas of interest should be investigated. Analyzing the antenna's impedance match over these frequencies does this investigation.

An antenna's impedance match is a critical factor in the overall efficiency of a transmitted signal. An antenna's impedance match can be analyzed by finding the Voltage Standing-Wave Ratio (VSWR) of an antenna system. High VSWR degrades transmitter performance, and can permanently damage the transmitter, so accurate measurement of VSWR is a necessity. We used the network analyzer in the microwave lab to measure the VSWR pattern. Figure 4.4 is the VSWR pattern of our Yagi-Uda using the network analyzer for measuring and plotting. The network analyzer does not measure VSWR directly. Instead, it measures forward and reflected power on the antenna feed-line. This data is used in equation 7 to find the antenna system VSWR.

$$VSWR = \frac{1 + \sqrt{\frac{P_{reflected}}{P_{forward}}}}{1 - \sqrt{\frac{P_{reflected}}{P_{forward}}}} \tag{4.7}$$

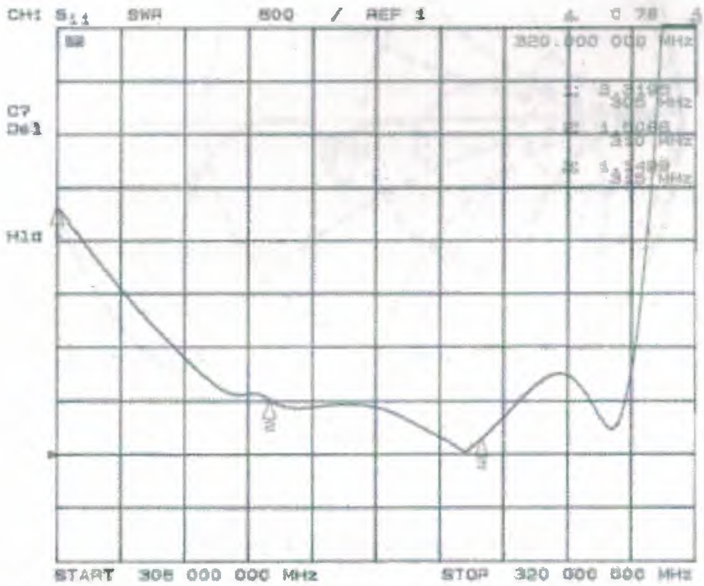


Figure 4.4 VSWR Plot

Figure 4.4 shows the VSWR for the radar antenna. It is taken over the frequency range of 305 MHz to 320 MHz. Notice the VSWR stays low until the frequency reaches 318 MHz. This suggests the antenna does not radiate as efficiently at frequencies above 318 MHz. Essentially, the radar antenna has a bandwidth of 18 MHz centered at 309 MHz. Unfortunately, this does not quite meet design specifications.

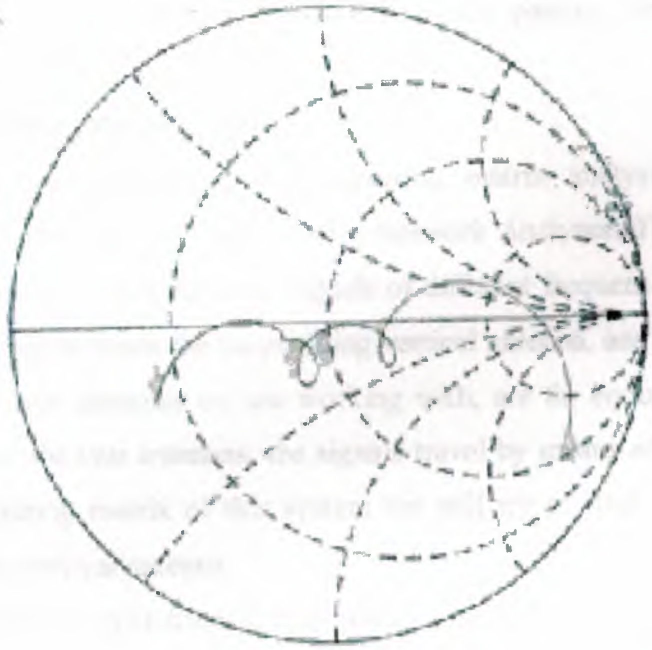


Figure 4.5 *Impedance Plot*

Figure 4.5 is a Smith Chart representing the impedance of the antenna and balun system over the frequency range of 305 to 320 MHz. This allows us to study the impedance mismatch closer and shows the input impedance of the balun and antenna system is very close to the transmission line impedance of 50ohms. It stays close to this 50 ohm line over most of this range. However, once again we begin getting poor performance around 318 MHz. This explains the poor VSWR of the antenna at these higher frequencies.

Two options may improve the input impedance of our antenna. First, we could try to find a balun that will provide a better match over the desired range. Second, the antenna elements might be trimmed. This would move the entire operation spectrum to higher frequencies.

4.3 Laboratory Experiment.

In this experiment we are going to measure radiation pattern, front to back ratio, directivity, gain and SWR.

4.3.1 Radiation Pattern Measurement

For measuring the radiation pattern, a scattering matrix analysis is used. The calculation of the S-parameters is done with a Network Analyzer. The system that we will use functions in the following way: signals of different frequencies will be sent by the Network Analyser between the transmitting vertical antenna, and the receiving coil. Assuming that the two antennas we are working with, are far enough apart to ignore near fields, between the two antennas, the signals travel by means of EM radiation. By analyzing the scattering matrix of this system we will try to find the basic radiation Patterns of a simple vertical antenna.

Vertical antenna have a cylindrically symmetric (omni-directional) radiation pattern, which is very useful in situations when we want to transmit and receive the same no matter which direction we are facing. When driving, for example, many of us use a vertical antenna for receiving broadcast FM stations, and the reception is more or less the same no matter which way our car is facing.

However, for other operations we need antennas with directional gain. This means that some antennas have a front and a back, a typical application for a "gain" antenna is for receiving TV broadcasts. Directivity here helps zero in on the signal, while attenuating signal reflections that might arrive a bit later from other directions and show up on the TV screen as "ghosts". Another reason for directional antenna is that, in their optimum direction, they will receive a stronger signal than a comparable omni-directional antenna.

For many applications the choice of the best directional gain antenna usually falls on Yagi-Uda arrays. To measure the transmission between two ports consisting of a 5/8 λ, loaded, ground plane cell phone "transmitting" antenna and a small "receiving" magnetic loop we use the concept of scattering .

By measuring the two-port scattering matrix elements S11, S12, S21, and S22 using a Network Analyser one can determine the kind of load the antenna presents to the transmitter over a range of frequencies, the kind of load the loop presents to the transmitter, and the type of signal attenuation between the antenna and the receiving loop. If one plots S21 (or S12, since this is a "reciprocal" network) as a function of the position of the loop, one can then plot the radiation pattern of the antenna.

4.3.2 Calibration Procedure:

Power up the HP Network Analyzer. Press the Cal Button. Choose "Calibrate Menu", "full2-port", and then "Reflection" from soft keys menu. Remove everything from the ports so that the brass connectors are showing. Choose "S11 open" and "S22 Open" from the soft key menu. Acquire a short terminator from the HP Calibration Kit, attach it to port 1 and press "S11short". Repeat with the short for port 2. Place a 50 Ohm load on port 1 and press "load". Repeat with the load for port 2. Now we are finished with the reflection part of the calibration—press "Reflection Done". Choose "transmission" from the soft keys. Short port 1 to port 2 using the coax from the receiving magnetic loop. To calibrate press all soft keys starting from the top in the sequence 1,2,4,3, so that the final line on the display is a horizontal. Choose "transmission Done" now select "isolation", then press "Omit Isolation", "Done" and "Done2-port" soft keys in the above sequence. Then press "SaveReg1". We are now finished with the calibration.

4.3.3 Measurements:

To start measurement, the antenna coax must be attach to port 1 and the receiving loop coax to port 2. Select the frequency span to the one desired (we took it 146 MHz in this experiment). By selecting SWR display from the format menu for S11 we can find the resonant frequency and standing wave ratio of the antenna.

Now switch to dual display by pressing "Display", "On".Select Ch.2—now we are operating on the bottom display. Press "Meas" button and choose "Trans:FWD S21, [B/R]". Select the format of the bottom display to be logarithmic, and auto-scale the display from the "Scale Ref" menu.

Slip the receiving coil through a loop in the string attached to the base of the antenna. Measure the strength of signal in dB in the range of (-10,90dB)degrees the ground plane in 10 degree increments. Keep the plane of the receiving loop perpendicular to the floor at all times—this will restrict measurement to the component of the magnetic field vector in the azimuthal (ϕ) direction.

4.3.4 Measuring the Gain of a Directional Antenna.

Attach the 3 element Yagi antenna to the output of the signal generator. Enter the frequency desired and amplitude of 0dB, with zero modulation. Attach the RF probe to *the HP Spectrum Analyser. Set the frequency scan from 140 to 150 Mhz.* set the sweep time to 300ms. We should see a spike at 146 Mhz on the display. Position the marker on top of the spike. In the top right hand corner you should see the signal strength of the antenna in dBm (decibels referenced to 1 milliwatt).

Rotate the antenna and mark the position that shows the strongest signal. The antenna front-to-back ratio is the difference between its best gain and the gain 180 degrees away from its best gain. Measure the antenna front-back ratio.

Front-to-back ratio = max signal (dB)-signal at 180 off (dB)

4.3.5 Modified Yagi Antenna

Measure the antenna gain in dBd (referenced to a dipole) by removing the reflector (the longer of the three elements) and the director (the shorter of the three elements) and noting the signal strength of the remaining dipole.

Note that the dipole after the director and reflector are removed is not resonant, i.e, it does not present the transmission line with a 50ohn termination any more. Use the Network Analyser to measure the SWR of the yagi antenna at 146 MHz with and without the reflector and director elements.

Because the antenna is no longer matched, the coaxial cable losses are about 1 dB greater than when the load was matched.

4.4 Design of Patch Triangular Microstrip Antenna

4.4.1 Antenna design procedure

Figure shows the configuration of the stacked triangular patch antenna that was considered in this session. The design procedure starts with the determination of the sidelength of the driven patch using the resonant frequency formula

$$f_{m,n,l} = \frac{2c}{3a\sqrt{\epsilon_r}}\sqrt{m^2 + mn + n^2}. \tag{4.8}$$

where m, n, and l are the mode integers due to the electric and magnatic boundary conditions, c is the speed of light in free space, ε is the dielectric constant and a is the sidelength of the equilateral triangle.

The simplest correction formula for the sidelength due to the effect of the fringing fields was given by Dahele and Lee:

$$a_{eff} = a + \frac{h}{\sqrt{\epsilon_r}}. \tag{4.9}$$

where a_{eff} is the effective sidelength of the equilateral triangle.

4.4.2 Measurements

Initial measurements carried out on the isolated driven patch, using a Wiltron Network Analyser System, showed that the return loss was -2.88dB at a resonant frequency of 2.523 GHz. The discrepancy between the design and measured frequencies is less than 1 percent. Figure illustrates the dependence of the return loss on the frequency for various values of height separation between the two patches. It is evident from Figure that, for the isolated driven lower patch, the value of the return loss obtained is a measure of the existing strong mismatch between the 50 feeding cable and the input impedance of the patch. An optimum impedance bandwidth of 330 MHz was measured for S=6.4mm with the two ends of the frequency band being at 2.43 GHz and 2.76 GHz respectively.

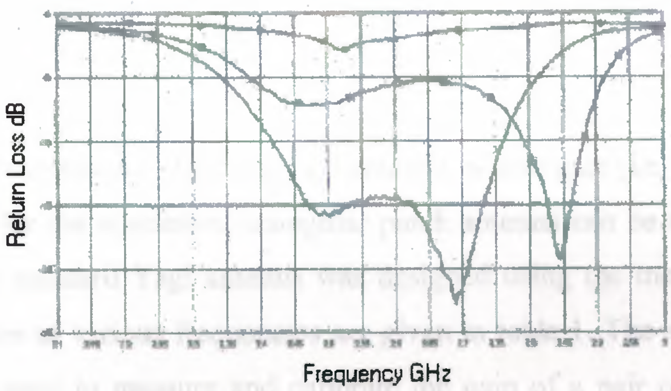


Figure 4.6 Return loss versus frequency.

The antenna gain and polar patterns were measured at these frequencies using the set-up shown in figure.

In the antenna gain measurement, the Friis transmission formula and a standard Yagi antenna were used in the following experiment procedure. First, the total power radiated from the transmitting antenna P_t must be established and kept constant. Then, readings for the power received P_r by the second antenna are taken, at various separations R between the two antennas as shown in Figure 3. The Friis transmission formula has the form y=m(x) when R^2 and the ratio P_t/P_r are made the variables. The relationship is as follows

$$R^2 = \left[\left(\frac{\lambda}{4\pi}\right)^2 G_t G_r\right]\frac{P_t}{P_r}.$$

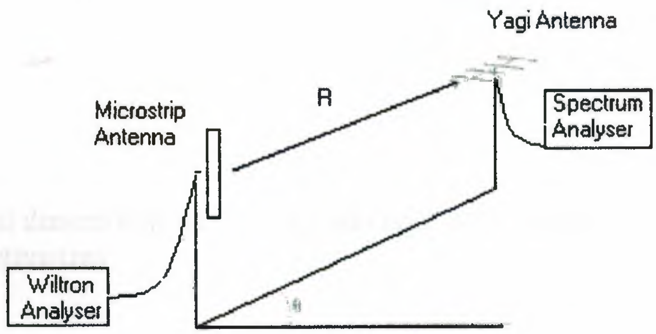


Figure 4.7 Antenna gain and polar pattern measurement set up.

The plot of R^2 on the y-axis against P_t/P_r on the x-axis gives a straight line with slope equal to

$$\left[\left(\frac{\lambda}{4\pi}\right)^2 G_t G_r\right]$$

Since one of the antennas is a standard Yagi antenna, whose gain (i.e, either G_t or G_r) is known, the gain for the equilateral triangular patch antenna can be readily calculated. The five element standard Yagi antenna was designed using the method of moments, and the dimensions at various frequencies are given in table 1. The same experimental procedures were used to measure and calibrate the gain of a pair of similar standard Yagi antennas.

In the present investigation, it was found that the antenna gain at 2.43 Ghz and 2.76 Ghz was 8.7 and 8.8 dB respectively. The E and H plane radiation patterns were also measured these frequencies (i.e. at the two ends of the frequency band). These are illustrated in Figure 4. The corresponding half –power beam widths are:

2.43 GHz : E-plane = 66 degrees

H-plane = 83 degrees

2.76 GHz: E-plane = 66 degrees

H-plane = 83 degrees

It is evident that there is no significant change in the polar pattern at these frequencies.

Table 4.1 Physical dimensions of the Yagi antenna, with frequency given in GHz and dimensions in centimetres

Frequency	Wavelength	Spacing	Reflector Length	Driver Length	Director length
1	30.00	7.50	14.31	13.53	13.26
1.1	27.27	6.82	13.01	12.30	12.05
1.2	25.00	6.25	11.93	11.28	11.05
1.3	23.08	5.77	11.01	10.41	10.20
1.4	21.43	5.36	10.22	9.66	9.47
1.5	20.00	5.00	9.54	9.02	8.84
1.6	18.75	4.69	8.94	8.46	8.29
1.7	17.65	4.41	8.42	7.96	7.80
1.8	16.67	4.17	7.95	7.52	7.37
1.9	15.79	3.95	7.53	7.12	6.98
2	15.00	3.75	7.16	6.77	6.63
2.1	14.29	3.57	6.81	6.44	6.31
2.2	13.64	3.41	6.50	6.15	6.03
2.3	13.04	3.26	6.22	5.88	5.77
2.4	12.50	3.13	5.96	5.64	5.33
2.5	12.00	3.00	5.72	5.41	5.30
2.6	11.54	2.88	5.50	5.20	5.10
2.7	11.11	2.78	5.30	5.01	4.91
2.8	10.71	2.68	5.11	4.83	4.74
2.9	10.34	2.59	4.93	4.67	4.57
3	10.00	2.50	4.77	4.51	4.42
3.1	9.68	2.42	4.62	4.36	4.28
3.2	9.38	2.34	4.47	4.23	4.14
3.3	9.09	2.27	4.34	4.10	4.02
3.4	8.82	2.21	4.21	3.98	3.90
3.5	8.57	2.14	4.09	3.87	3.79
3.6	8.33	2.08	3.98	3.76	3.68
3.7	8.11	2.03	3.87	3.66	3.58
3.8	7.89	1.97	3.77	3.56	3.49
3.9	7.69	1.92	3.67	3.47	3.40
4	7.50	1.88	3.58	3.38	3.32
4.1	7.32	1.83	3.49	3.30	3.23
4.2	7.14	1.79	3.41	3.22	3.16
4.3	6.98	1.74	3.33	3.15	3.08
4.4	6.82	1.70	3.25	3.08	3.01
4.5	6.67	1.67	3.18	3.01	2.95

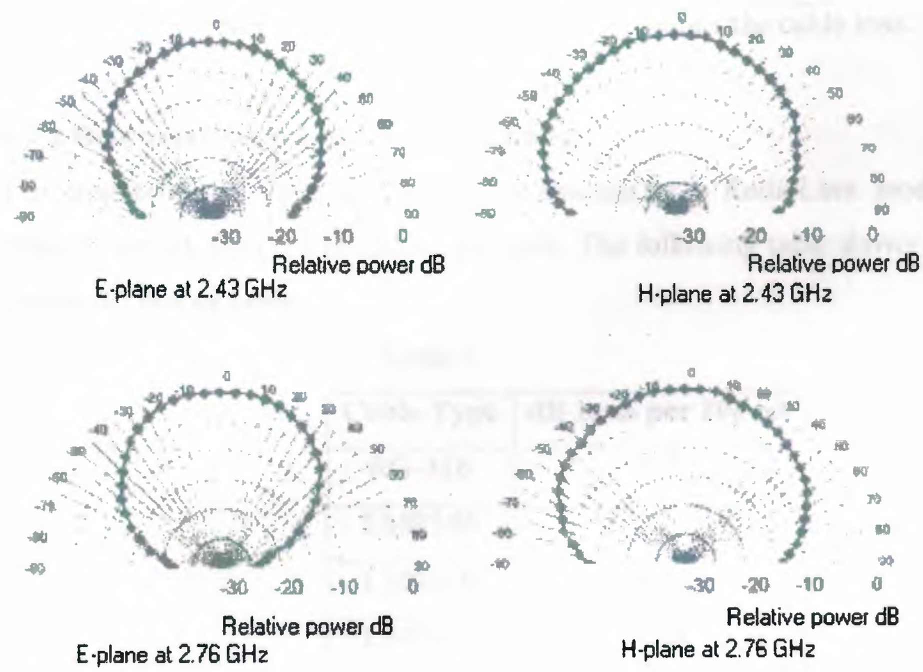


Figure 4.8 Radiation patterns at the two ends of the frequency band.

4.5 Gain Calculation

To choose the best antenna the following procedure must be followed:

4.5.1 Determine Required Gain

1- Identify the unobstructed distance between Radio Linx antennas

2- Find the row that corresponds to the distance between antennas.

3- Find the Link Gain in the appropriate *row.*

Table 2

Unobstructed Distance (miles)	Net dB (dB)
1	4
2	7
3	10
5	15
7	20
10	25

12	30
15	35

Net dB is the gain of the antenna minus the cable loss.

4.5.2 Determine Gain Loss Due to Cabling

The lenght of the cable connecting an antenna to a RadioLinx modem or wireless Tthernet switch affects overall antenna gain. The following table shows approximate dB loss per10 feet of cable.

Table 3

Cable Type	dB Loss per 10Feet
RG-316	4.2
LMR195	1.8
LMR400	0.67
LMR600	0.44
LMR1200	0.23
LDF4	0.35
LDF5	0.20
LDF7	0.13

Determine how much cable you will need by measuring the approcimate distance beteween where your RadioLinx modem or wireless Ethernet switch will be placed and the proposed location of the antenna in feet. Do the same for the remote RadioLinx modem or wireless Ethenet switch.

Now use the chart above to determine how much gain loss you will have to calculate when choosing your equipment. This will give your cable gain loss.

4.5.3 Calculate total gain

The "link Gain: is a composite of the gains of each of the antennas (the Master's antenna and the Remote's antenna)as well as any cable loss. For example, if you want to communicate over a 5 mile unobstructed distance, you should include at least 15db of Link Gain . if you are using 4 feet of cable on the master side and 10 feet of cable at the Remote side, you need to determine what the cable loss is , supposing that the unobstructed distance beteween RadioLinx devices is 5 miles.

Using the tables above we can get :

Cable Loss at theMaster (4 feet)	2 dB.
Cable Loss at the Remote (10 feet)	5 dB.
Equals Cable Gain Loss	7 dB.

To make sure that we have antennas that will cover cable loss, we need to add the Required Gain to the Cable Gain Loss.

Unobstructed Distance: (5 miles)	Net dB needed is	15 dB.
Cable Gain Loss		7 dB.
Required Gain when Selecting Antenna:		22 dB

So we can select antennas based on the calculated gain vlaue of 22dBi. The gain of each of our antennas must be equal to or greater than this value.

More gain will give you more distance. İf doesnt make any difference whether the gain is on the Master or the Remote antenna. The available gain is the sum of both antennas.

Based on this , we can select an antenna at the Master site with a gain of 8dBi and an antenna at the Remote site with a gain of 15dBi totaling 23dBi.

The calculation now looks like this:

Gain Required:15dBi (5miles distance)

Master Antenna Gain:	8dBi
Remote Antenna Gain:	15dBi
Total Antenna Gain:	23dBi
Total Cable Loss:	7dBi

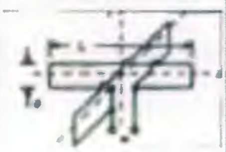
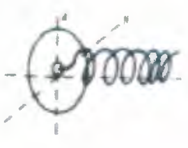
Total Antenna Gain minus Total Cable Loss = Antenna Gain

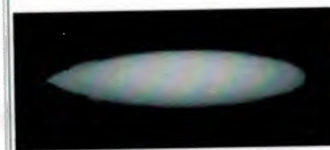
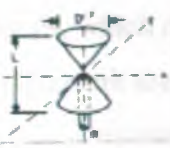
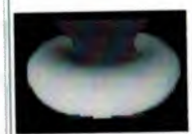
Or

23dBi – 7 dBi = 16 dBi

we can now choose our equipment based on this information.

4.6 Antenna Chart

Name	Shape	Gain (over isotropic)	Beamwidth - 3 dB	Radiation Pattern
Isotropic		0 dB	360	
Dipole $\lambda/2$		2.14 dB	55	
Folded Dipole $\lambda/2$ Cylindrical		5.64 dB	45	
Turnstile $\lambda/2$		-0.86 dB	50 due to cusps	
Full wave loop $D = \lambda/\pi$		3.14 dB	200	
Yagi $\lambda/2$		7.14 dB	25	
Helical $L = 6\lambda$		10.1 dB	30	

Parabolic Dipole $D = 5\lambda/2$		14.7 dB	20	
Horn 3λ		15 dB	15	
Biconical Horn $H = 9\lambda/2$ $D = 14\lambda$		14 dB	360x200	

Conclusion

The ideal antenna is the one that will radiate all the power delivered to it by a transmitter in the desired polarization. Practical antennas can never fully achieve this ideal performance, but their merit is conveniently described in terms of the degree to which they do so. For this purpose, certain parameters of antenna performance are defined.

Antenna measurements are needed often to validate theoretical data, and sometimes to determine some values, which are very difficult to have by calculations.

The two basic measurement are : gain measurement and polarization measurement.

The first one (gain measurement) deal with the measurement of the ability to concentrate in a particular direction the net power accepted by the antenna from the connected transmitter.

The second one (polarization measurement) deal with the measurement of polarization's nature.

The most problems we are facing while we are measuring are : the unwanted reflections from the ground and the surrounding and in many cases, it is impractical to move antenna from the operating environment to the measuring site.

The antenna measurements are very expensive and need gigantic instruments to pursue this work; so that, we decided to search about this subject to make these measurements cheaper and easier ways for finding these results.

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