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Antenna Basics

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1.1 Introduction

Antennas radiate and receive electromagnetic waves by converting guided waves supported by a guiding structure into radiating waves propagating in free space and vice versa. This function has to be accomplished by fulfilling specific requirements which affect the antenna design in different ways. In general, a number of antennas are installed in a satellite and their requirements vary depending on the application and on the mission. They can be roughly classified into three types: antennas for telemetry, tracking and control (TT&C), high-capacity antennas, and antennas for space instruments or for other specific applications. Several examples of the latter class are reported in the third section of this book.

This chapter provides an overview of the basic antenna parameters and antenna types, and it presents other basic concepts related to the space environment which will introduce the reader to the development of antennas for space applications. Although many basic definitions are presented, the chapter is not intended to provide a comprehensive background to antennas. For this reason, the reader should refer to the extensive literature available on the subject, some of which we list as references.

The chapter is organized as follows. In the first part, the main antenna parameters will be given in accordance with the IEEE Standard Definition of Terms for Antennas [1] and with the IEEE Standard Test Procedures for Antennas [2] which will be adopted throughout the book. In the second part of the chapter, basic antenna types commonly employed in spaceborne applications will be presented. In the third part of the chapter, antenna development will be related to the space environment by introducing fundamental concepts such as multipaction and outgassing.

1.2 Antenna Performance Parameters

Numerous parameters exist for characterizing the performance of antennas and in the following subsections the most significant of these are reviewed. The relevance of these antenna parameters will be seen in Chapter 3 where they are combined into the Friis transmission formula which links the available power of the transmitter to the received power of the receiver in a radio communication system.

1.2.1 Reflection Coefficient and Voltage Standing Wave Ratio

For a multi-port antenna as shown in Figure 1.1, the scattering parameters, S_{ij} , relate the equivalent voltage of the outgoing wave at port i , V_i^- , to the equivalent voltage of the incoming wave at port j , V_j^+ , that is, $V_i^- = S_{ij}V_j^+$ [3]. The reflection coefficient at the i ' port is

$$\Gamma_i \equiv V_i^- / V_i^+ = S_{ii} + \sum_{j \neq i} S_{ij} V_j^+ / V_i^+ \quad (1.1)$$

For a single-port antenna, or for a multi-port antenna with all other ports matched (thus $V_j^+ = 0$ for $j \neq i$), the reflection coefficient Γ_i equals the scattering coefficient S_{ii} and, if the antenna is passive, the magnitude of the reflection coefficient is then less than or equal to 1. Note that the reflection coefficient is defined in terms of equivalent voltage which requires the existence of a well-defined mode in the port of the antenna. Furthermore, the voltage is defined at a specific position – the reference plane – in the antenna port, and the reflection coefficient is thus referenced to that position.

The voltage standing wave ratio (VSWR) is the ratio of the maximum and minimum voltages on the transmission line connected to the antenna, and it follows directly from the reflection coefficient Γ as

$$\text{VSWR} = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad (1.2)$$

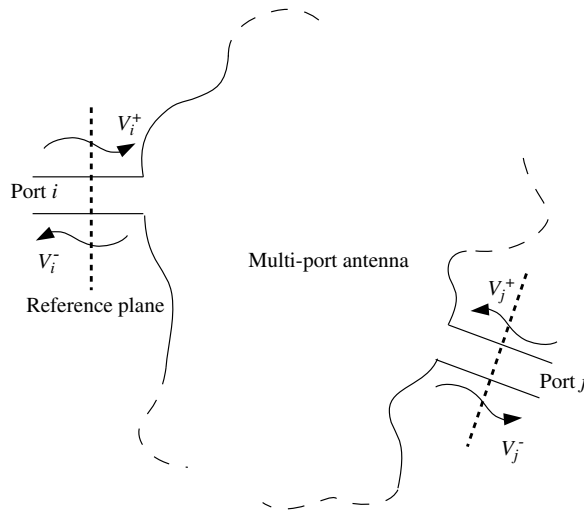


Figure 1.1 Arbitrary multi-port antenna.

The scattering parameters are the main representation of antenna behavior with respect to the circuit to which the antenna is connected. This is particularly true for passive antennas while more complex parameters are required for active antennas.

1.2.2 Antenna Impedance

The input impedance of an antenna Z_A is the ratio of the voltage V and current I at the port of the antenna when the antenna is isolated in free space; that is, without the presence of other antennas or scattering structures. Thus, this is sometimes referred to as the isolated input impedance. Since voltage and current are not practical quantities at radio frequencies (RFs), the input impedance is usually determined from the reflection coefficient Γ and the characteristic impedance Z_C of the transmission line connected to the port of the antenna; that is,

$$Z_A \equiv \frac{V}{I} = Z_C \frac{1 + \Gamma}{1 - \Gamma} \quad (1.3)$$

For a linear multi-port antenna the voltage at the i th port can be related to the currents at all ports as

$$V_i = Z_{ii}I_i + \sum_{j \neq i} Z_{ij}I_j \quad (1.4)$$

where Z_{ii} is the self-impedance of the i th port and Z_{ij} is the mutual impedance between the i th and j th ports. The input impedance of the i th port is then

$$Z_{A,i} \equiv V_i/I_i = Z_{ii} + \sum_{j \neq i} Z_{ij}I_j/I_i \quad (1.5)$$

which is seen to depend on the excitations (currents) of the other ports and therefore differs from the isolated input impedance. Thus, the input impedance of a port in a multi-port system is sometimes referred to as the active input impedance. Even the self-impedance, which is seen from above to equal the active input impedance when all other ports are open-circuited (zero current), is generally different from the isolated input impedance since the open-circuited ports may still act as scattering structures. For an antenna array, see Section 1.4, with identical antenna elements and thus identical isolated input impedances, the active input impedances may differ due to the mutual coupling. Furthermore, if the excitation of the ports is changed, for example, to scan the main beam in a phased array, the active input impedance of an individual port can vary drastically and become very poorly matched to the transmission line characteristic impedance.

If the scattering parameters are arranged in a scattering matrix $\bar{\bar{S}}$ and the self- and mutual impedances in an impedance matrix $\bar{\bar{Z}}$, the relationship between these, for a multi-port antenna with the common characteristic impedance of the transmission lines on the ports Z_C , can be expressed as ($\bar{\bar{U}}$ is the unit matrix)

$$\bar{\bar{Z}} = Z_C(\bar{\bar{U}} + \bar{\bar{S}}) \cdot (\bar{\bar{U}} - \bar{\bar{S}})^{-1} \quad (1.6)$$

$$\bar{\bar{S}} = (\bar{\bar{Z}} + Z_C\bar{\bar{U}})^{-1} \cdot (\bar{\bar{Z}} - Z_C\bar{\bar{U}}) \quad (1.7)$$

1.2.3 Radiation Pattern and Coverage

The radiation pattern is a ‘mathematical function or graphical representation of the radiation properties of the antenna as a function of space coordinates’ [1]. In the most common case, antenna radiation patterns are determined in the far-field region [4]. This region is ‘where the angular field distribution is essentially independent of the distance from a specified point in the antenna region’ [1]. Typically, the far-field region is identified by those distances greater than $2D^2/\lambda$, D being the maximum overall dimension of the antenna and λ the free-space wavelength. In the far-field region of any antenna the radiated field takes a particularly simple form. For time-harmonic fields, and using phasor notation with the suppressed time factor $\exp(j\omega t)$ with ω the angular frequency and t time, the far-field can be expressed as

$$\lim_{r \rightarrow \infty} \mathbf{E}(\mathbf{r}) = \mathbf{P}(\mathbf{a}_r) \frac{e^{-jkr}}{r} \quad (1.8)$$

Thus, the radiated electric field $\mathbf{E}$ at the position of the position vector $\mathbf{r}$ can be expressed as the product of a pattern function $\mathbf{P}$ that depends only on the direction $\mathbf{a}_r$ of the position vector and the term $\exp(-jkr)/r$ that depends only on the length r of the position vector. Furthermore, the pattern function $\mathbf{P}$ has only transverse components w.r.t. $\mathbf{a}_r$; that is, $\mathbf{P} \cdot \mathbf{a}_r = 0$. The position vector $\mathbf{r}$ is referenced to the origin of the antenna coordinate system. Note that the pattern function $\mathbf{P}$ defines all radiation properties that are particular for the antenna.

The parameter represented by the radiation pattern is typically a normalized magnitude of the pattern function or one of its components, the directivity or partial directivity, or the gain or partial gain – but it may be the phase of a polarization-phase vector component, the axial ratio, or the tilt angle as well; these parameters are reviewed in the following subsections. The graphical representation may be two or three dimensional with the transmission/reception direction typically expressed by the polar θ and azimuthal ϕ coordinates of the antenna coordinate system for a full-sphere pattern or the projected coordinates $u = \sin\theta \cos\phi$ and $v = \sin\theta \sin\phi$ for a hemispherical pattern.

An antenna can be defined as directional when it can ‘radiate or receive electromagnetic waves more effectively in some directions than in others’ [1]. In order to discriminate between directional and non-directional antennas, the half-wave dipole is normally taken as reference while the antenna directivity is generally compared to the ideal isotropic radiator [5]. Normally, the portion of the radiation pattern of a directive antenna where the radiation intensity is maximum is defined as the main lobe. Side, minor, back and grating lobes can also be identified. The first three types are related to the direction and to the intensity of radiation while the last one can be present only in an antenna array environment.

1.2.3.1 Half-Power Beamwidth

The half-power beamwidth (HPBW) is identified in a cut of a radiation pattern as the angle between the two directions in which the radiation intensity is half of its maximum value (see Figure 1.2a). HPBW characterizes the behavior of the antenna in its main lobe but it does not take into account the amount of power radiated out of the main beam. For this reason, parameters are normally used to more accurately evaluate the antenna’s directional performance.

1.2.3.2 Coverage

The coverage C of an antenna is the range of transmission/reception directions over which one or more antenna parameters meet certain specifications. In most cases, the coverage C refers to the directivity or gain, or the

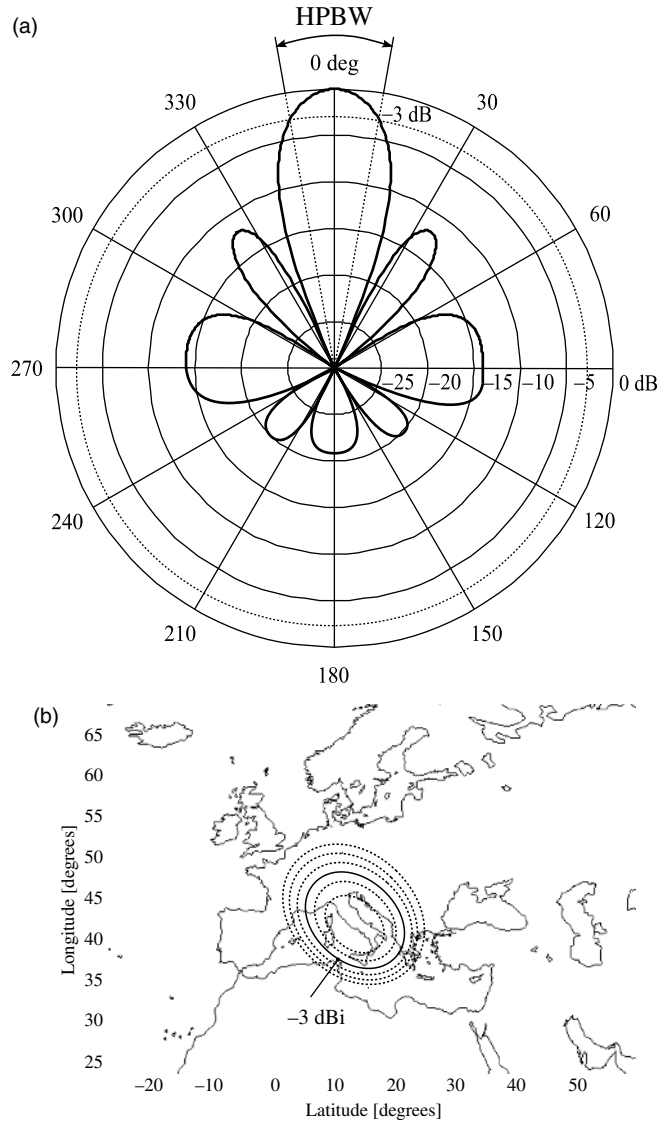


Figure 1.2 Radiation pattern: (a) half-power beam width (HPBW); (b) footprint example (1 dBi per circle).

co-polarized partial directivity or gain, and is thus the range over which the relevant parameter is larger than a specified minimum value; this could be 3 dB below the maximum value. When the antenna points towards the Earth, it is convenient to express the coverage in terms of Earth footprint, which is the projection of the satellite antenna pattern onto the Earth's surface (see Figure 1.2b). The footprint is that portion of the Earth's surface where the antenna points with a given gain. For some applications the footprint corresponds simply to a circle in a (θ, ϕ) coordinate system, calling for a pencil-beam antenna, while for other applications the

coverage is the projected shape of a country, calling for a shaped-beam antenna. Clearly, both the footprint and the coverage C can be determined from the radiation pattern and thus the pattern function $\mathbf{P}$.

1.2.4 Polarization

The polarization of an antenna in a given direction is the polarization of the plane wave transmitted (or received) by the antenna in the far field. Polarization is classified as linear when the electric field in a given direction is always directed along a line. Pure linear polarization is an ideal case as all antennas generate both a co-polarization field, that is, the polarization the antenna is intended to radiate, and a cross-polarization field, that is, in the case of linearly polarized fields, the component of the electric field orthogonal to the desired polarization. For this reason, the electric field vector normally describes an ellipse and the polarization is classified as elliptical. If the axes of the ellipse are equal, then the polarization is referred to as circular. It is worth noticing that the polarization of an antenna is normally defined by taking into account the radiating wave. Satellite–Earth communication links typically adopt circularly polarized (CP) signals. Indeed, the use of linear polarization would lead to high polarization mismatches arising from alignment issues or from the Faraday rotation effect of the ionosphere [6–8].

Antenna polarization can be described in terms of the polarization-phase vector, $\mathbf{p}$, that is, a unit vector that represents the polarization as well as the phase of the radiated field of an antenna. The simple distance dependence of the phase due to the term $\exp(-jkr)/r$ is not included in the polarization-phase vector, and from the pattern function $\mathbf{P}$ the polarization-phase vector is thus defined as

$$\mathbf{p} \equiv \mathbf{P}/|\mathbf{P}| \quad \text{with} \quad |\mathbf{P}| = \sqrt{\mathbf{P} \cdot \mathbf{P}^*} \quad (1.9)$$

The pattern function $\mathbf{P}$ can be decomposed w.r.t. two orthogonal polarization unit vectors referred to as the co- and cross-polarization unit vectors; that is,

$$\mathbf{P} = \mathbf{P} \cdot \hat{\mathbf{a}}_{co}^* \hat{\mathbf{a}}_{co} + \mathbf{P} \cdot \hat{\mathbf{a}}_{cross}^* \hat{\mathbf{a}}_{cross} \equiv P_{co} \hat{\mathbf{a}}_{co} + P_{cross} \hat{\mathbf{a}}_{cross} \quad (1.10)$$

where the polarization unit vectors, $\hat{\mathbf{a}}_{co}$ and $\hat{\mathbf{a}}_{cross}$, are typically the linearly polarized θ and ϕ unit vectors of the standard spherical coordinate system, the linearly polarized unit vectors according to Ludwig's third definition [9], or the circularly polarized unit vectors defined from either of these linearly polarized unit vectors. Obviously, the polarization-phase vector can be decomposed in the same manner; that is, $\mathbf{p} \equiv p_{co} \hat{\mathbf{a}}_{co} + p_{cross} \hat{\mathbf{a}}_{cross}$.

The polarization-phase vector can also be represented in terms of the polarization ellipse with its axial ratio, tilt angle, and sense of rotation. In each direction of observation $\hat{\mathbf{a}}_r$, a local right-hand orthogonal, rectangular xyz coordinate system is defined with the unit vectors $\hat{\mathbf{a}}_x$ and $\hat{\mathbf{a}}_y$ transverse to, and the unit vector $\hat{\mathbf{a}}_z$ parallel to, the direction of observation $\hat{\mathbf{a}}_r$. The polarization-phase vector is now decomposed as $\mathbf{p} \equiv p_x \hat{\mathbf{a}}_x + p_y \hat{\mathbf{a}}_y$. The axial ratio (AR) can then be expressed as

$$\text{AR} = \sqrt{\frac{1 + |\mathbf{p} \cdot \mathbf{p}|}{1 - |\mathbf{p} \cdot \mathbf{p}|}} \quad (1.11)$$

while the tilt angle (TA) w.r.t. the direction $\hat{\mathbf{a}}_x$ is

$$\text{TA} = \arctan \left(\text{Re} \left(p_y \exp \left(-\frac{j}{2} \arg(\mathbf{p} \cdot \mathbf{p}) \right) \right) / \text{Re} \left(p_x \exp \left(-\frac{j}{2} \arg(\mathbf{p} \cdot \mathbf{p}) \right) \right) \right) \quad (1.12)$$

and the sense of rotation (SOR) is

$$\text{SOR} = \begin{cases} \text{right-hand if } 0 < \arg(p_x) - \arg(p_y) < \pi \\ \text{left-hand if } \pi < \arg(p_x) - \arg(p_y) < 2\pi \end{cases} \quad (1.13)$$

Alternatively, the AR can be determined from the magnitude of the right- and left-hand circularly polarized components of the electric field, $\mathbf{E}_{RHCP}$ and $\mathbf{E}_{LHCP}$ respectively. The expression for AR can thus be cast in the form

$$\text{AR} = \frac{|\mathbf{P}_{RHCP}| + |\mathbf{P}_{LHCP}|}{|\mathbf{P}_{RHCP}| - |\mathbf{P}_{LHCP}|} \quad (1.14)$$

Differences between the polarization of the incident wave and the receiving antenna are normally referred to as polarization mismatch. In general, they can cause strong link losses which can be taken into account by using different figures of merit. One possibility is to use co-polarization and cross-polarization field patterns. Alternatively, the polarization efficiency e_p could be defined as [1]

$$e_p = |\hat{\mathbf{p}}_i \cdot \hat{\mathbf{p}}_a|^2 \quad (1.15)$$

where $\hat{\mathbf{p}}_i$ and $\hat{\mathbf{p}}_a$ are the polarization vectors of the incident wave and of the receiving antenna respectively. If the polarization of the incident wave and that of the receiving antenna are the same, the inner product defined in Equation 1.15 is equal to 1.

1.2.5 Directivity

The directivity D of an antenna is the ratio of the far-field radiation intensity in a given direction to the average radiation intensity over the radiation sphere. The radiation intensity in a given direction, U , is the radiated power per solid angle Ω and thus $U = |\mathbf{P}|^2 / 2\eta_0$ with η_0 being the free-space intrinsic impedance and $|\mathbf{P}|^2$ the power radiated in a given direction. Thus the directivity D in a given direction can be expressed as

$$D \equiv \frac{4\pi U}{P_{rad}} = \frac{2\pi |\mathbf{P}|^2}{\eta_0 P_{rad}} = \frac{4\pi |\mathbf{P}|^2}{\int_{4\pi} |\mathbf{P}|^2 d\Omega} \quad (1.16)$$

P_{rad} is the total radiated power, which can be calculated by integrating the power radiated in a given direction over the entire radiating sphere. When the direction is not specified, the maximum directivity is usually taken.

Antenna directivity can be discriminated in terms of polarization by defining partial directivities. The partial directivities, D_{co} and D_{cross} , in a given direction for the co- and cross-polarized components can then be expressed as

$$D_{co} = \frac{4\pi |P_{co}|^2}{\int_{4\pi} |\mathbf{P}|^2 d\Omega} \quad \text{and} \quad D_{cross} = \frac{4\pi |P_{cross}|^2}{\int_{4\pi} |\mathbf{P}|^2 d\Omega} \quad (1.17)$$

1.2.6 Gain and Realized Gain

The gain G of an antenna in a given direction is the ratio of the radiation intensity to the average radiation intensity over the radiation sphere if all accepted power is radiated isotropically. In mathematical form, this can be written as

$$G \equiv \frac{U}{P_{acc}/4\pi} = \frac{2\pi|P|^2}{\eta_0 P_{acc}} \quad (1.18)$$

where P_{acc} is the power accepted in input by the antenna. The antenna gain can be related to the directivity by taking into account the antenna radiation efficiency, e_{cd} which can be defined as the ratio of the radiated power to the accepted power. Using the definitions of directivity and gain, it can thus be expressed as

$$e_{cd} \equiv \frac{G}{D} \quad (1.19)$$

For a lossless antenna where all accepted power is also radiated, $P_{rad} = P_{acc}$ and $e_{cd} = 1$, the gain G equals the directivity D . However, for most practical antennas the loss is non-negligible, $P_{rad} \neq P_{acc}$ and $e_{cd} < 1$, and it is important to distinguish between gain and directivity. When the direction of radiation is not stated, the direction of maximum radiation is normally presumed. In analogy to partial directivity, partial gain can be defined to discriminate the antenna gain w.r.t. the polarization of the radiated fields.

According to IEEE standards, the definition of antenna gain does not include reflection losses and polarization mismatches. The realized gain $G_{realized}$ of an antenna is the ratio of the radiation intensity to the average radiation intensity over the radiation sphere if all incident power is radiated; it thus includes the effect of the impedance mismatch at the antenna terminals and can be expressed as

$$G_{realized} \equiv \frac{U}{P_{in}/4\pi} = G(1-|\Gamma|^2) = e_0 D \quad (1.20)$$

where $e_0 = e_{cd}(1-|\Gamma|^2)$ is the overall antenna efficiency. The relevance of realized gain is clear from the Friis transmission formula that comprises the product of the gain and the impedance mismatch factor for both the transmitter and the receiver (see Chapter 3 for further details). However, since the loss and the mismatch are two completely different mechanisms, it is still important to use gain and reflection coefficients separately and to distinguish between gain and realized gain.

1.2.7 Equivalent Isotropically Radiated Power

Equivalent isotropically radiated power (EIRP) in a given direction is defined as ‘the gain of a transmitting antenna multiplied by the net power accepted by the antenna from the connected transmitter’ [1]. EIRP can be written as

$$\text{EIRP} = P_T G_T \quad (1.21)$$

where P_T is the net power accepted by the antenna from the transmitter and G_T is the gain of the transmitting antenna. In order to include transmitter output power, P_{Tx} , and interconnection losses between transmitter and antenna, L_c , Equation 1.21 can be changed to

$$\text{EIRP} = \frac{P_{Tx} G_T}{L_c} \quad (1.22)$$

The EIRP definition is important because it allows calculation of absolute power and field strength values and it makes possible comparisons between different emitters regardless of the type of configuration.

1.2.8 Effective Area

The effective area A_{eff} of a receiving antenna is the ratio of the available power at the terminals of the antenna to the power density of a polarization-matched incident plane wave. The effective area A_{eff} can be measured itself, but in most situations it is found from its reciprocity-based relation to the gain G as

$$A_{eff} = \frac{\lambda^2}{4\pi} G \quad (1.23)$$

1.2.9 Phase Center

In IEEE standards, the phase center is defined as ‘the location of a point associated with an antenna such that, if it is taken as the center of a sphere whose radius extends into the far-field, the phase of a given field component over the surface of the radiation sphere is essentially constant’. As the size of real antennas is not null, the phase center depends on the observation direction. In general, the phase center is calculated experimentally by measuring the phase pattern at different cut planes [10].

For some applications, knowing the location of the phase center is very important. For example, in a reflector antenna the phase center of the feed needs to be located at the focal point of the paraboloid. Another example where the phase center location is critical is the global navigation satellite system (GNSS) [11]. Indeed, one of the parameters which defines the accuracy of high-precision GNSSs is the invariance of the phase center which should be highly stable in order to minimize positioning errors.

1.2.10 Bandwidth

The bandwidth (BW) of an antenna is the range of frequencies over which one or more antenna parameters meet certain specifications. In most cases, BW refers to the reflection coefficient Γ and is thus the range over which Γ is less than a specified maximum allowable value, Γ_{max} , with the tacit assumption that other parameters remain within their specifications too. BW depends strongly on the value of Γ_{max} and it is important that this be stated explicitly. With f_u and f_l denoting the upper and lower limits of the frequency range, respectively, the fractional bandwidth (FBW) is

$$FBW = \frac{f_u - f_l}{f_c} \quad \text{with} \quad f_c = \frac{f_u + f_l}{2} \quad (1.24)$$

with the condition that the center frequency f_c coincides with the frequency of operation.

When multiple antenna parameters have to be considered, BW is given by the minimum range of frequencies over which specifications are satisfied. Typically, link budget calculations pose stringent requirements on the antenna gain and coverage, on the polarization efficiency, and on the reflection coefficient over the system bandwidth.

1.2.11 Antenna Noise Temperature

The antenna noise temperature T_A of a receiving antenna is the temperature (in kelvin) that, through the formula $P_n = KT_A BW$, gives the noise power P_n at the terminals of the antenna; K is Boltzmann’s constant

and BW the bandwidth [12]. In terms of the background noise temperature $T_B(\Omega)$ over the radiation sphere of the antenna, expressing the noise from the sky, the satellite structure and the Earth, and the antenna physical temperature T , the antenna noise temperature T_A in the radio frequency range can be expressed as

$$T_A = \frac{\eta_{rad}}{4\pi} \int_{4\pi} T_B(\Omega) D(\Omega) d\Omega + (1 - \eta_{rad}) T \quad (1.25)$$

While all previous antenna parameters relate only to the antenna itself, and any influence of the surroundings on these is considered parasitical, the antenna noise temperature relates equally to the antenna and its surroundings and it is particular also for the latter. Equation 1.25 shows that the antenna noise temperature can be calculated from the directivity D , the radiation efficiency η_{rad} and the background temperature T_B .

1.3 Basic Antenna Elements

Depending on the electrical and physical requirements, spacecraft antenna design can be based on different classes of radiators. In this section a basic overview of some of the most common antenna types is provided and includes references to relevant chapters of this book.

1.3.1 Wire Antennas

The dipole antenna is the most representative type of wire radiator. In the most common case, it consists of a linear wire with a feed point at the center as shown in Figure 1.3a. The radiation properties depend on the current distribution along its main axis, this current being mainly related to the dipole length. Some radiation occurs in all directions with the exception of the dipole axis itself. Due to the rotational symmetry of the dipole around its main axis (z -axis in Figure 1.3a), the radiation pattern is symmetrical over the azimuthal ϕ coordinate. In Figure 1.3b, c the radiation pattern of an ideal half-wavelength dipole is shown. Its maximum directivity is 2.15 dB and its HPBW is equal to 78° . The behavior of a dipole antenna changes when the dipole interacts with the spacecraft. For this reason, the ideal pattern shown on Figure 1.3b, c is valid only for isolated dipoles and it does not take into account the interactions with the spacecraft as shown in the examples of Chapter 2.

The monopole antenna is formed by replacing one-half of a dipole with an infinite ground plane perpendicular to the dipole axis as shown in Figure 1.3d. Using image theory [13], the fields above the ground plane can be found by substituting the ground plane with image currents forming the missing half of the dipole. The radiating behaviour of these two wire antennas is similar, but the monopole radiation below the ground plane is ideally null. For this reason, the directivity of a monopole antenna of length l is twice that of the equivalent dipole antenna of double length, $2l$.

Since the early spacecraft missions (see Chapter 7), wire antennas have been widely used in space exploration. Due to their omnidirectional radiation properties, dipole and monopole antennas are generally used to send or receive telemetry and command signals during launch, when the spacecraft attitude is out of control or in other circumstances when the high-directivity antennas cannot be employed.

1.3.2 Horn Antennas

Another important type of antenna which has found wide application in space missions is the horn antenna. In general, horn antennas are employed in satellite missions to produce wide-beam coverage such as Earth

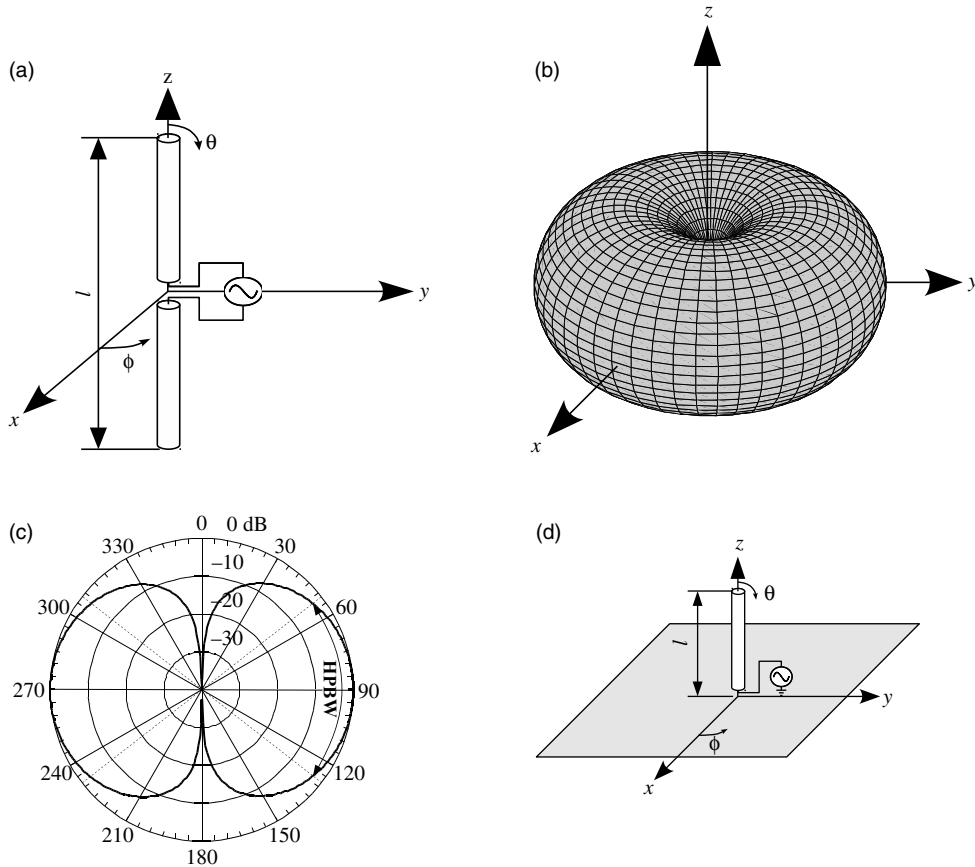


Figure 1.3 Dipole and monopole antennas: (a) dipole antenna geometry; (b) 3D normalized amplitude radiation pattern (dB) of a dipole antenna; (c) elevation plane normalized amplitude pattern (dB); (d) monopole antenna geometry.

coverage or to feed reflector antennas. Horn antennas are designed to provide a smooth transition between the feeding waveguide and a wider aperture which serves to focus the main lobe. Horn antennas belong to the category of aperture antennas, their radiation characteristics being determined by the field distribution across the aperture. The most common type of horn antenna is the pyramidal horn shown in Figure 1.4a. The horn provides a transition of length d between a feeding section of rectangular waveguide of height a and width b and a radiating aperture of height A and width B . In the most common case, the waveguide is excited by a single TE_{10} mode. In this case the dominant horn polarization would be linear with the main electric field component directed along the z -axis. Horn polarization can also be circular or dual linear depending on the modes excited in the waveguide section. Knowing the waveguide dimensions and the gain specifications, the pyramidal horn geometry can be defined through simple analytical formulas derived from the hypothesis of an aperture terminating in an infinite flange [14]. In general, the finiteness of the terminating flange can lead to inaccuracies which can be overcome through full-wave analysis. As a general rule, for a given horn length, d , as the aperture width, B , increases, the gain increases until it reaches a maximum after which it starts to decrease.

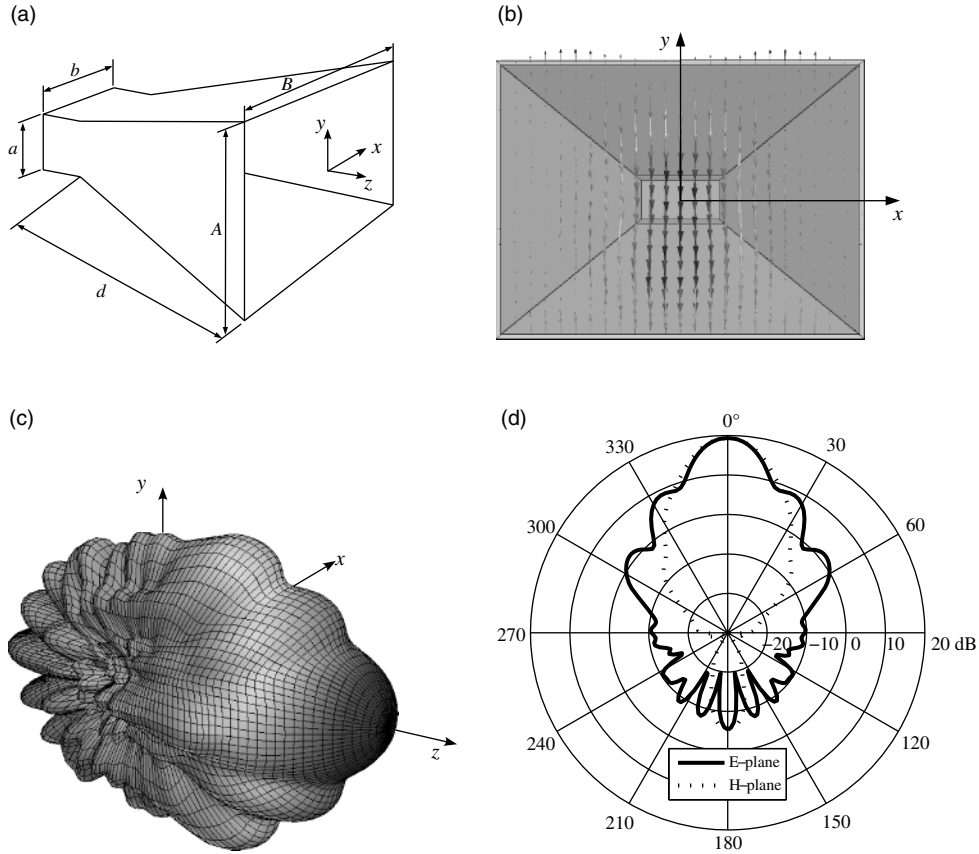


Figure 1.4 Pyramidal horn antenna: (a) geometry; (b) electric field distribution on the antenna aperture; (c) typical 3D normalized amplitude radiation pattern (dB) of a horn antenna; (d) typical gain pattern on the E- and H-plane of a vertically polarized pyramidal horn.

Figure 1.4 shows the field patterns of a pyramidal horn simulated through full-wave software [15]. Results were obtained from a rectangular horn antenna with $A = 120$ mm, $B = 90$ mm and $d = 120$ mm at a frequency of 10 GHz. The horn is fed through a standard section of WR102 waveguide excited in its fundamental mode. As can be observed in Figure 1.4b, the electric field vector on the antenna aperture is polarized along the y-axis. The y-z plane is thus referred to as the *E-plane* as it contains the *E*-field vector and the maximum direction of maximum radiation. Similarly, the x-z plane is referred to as the *H-plane*. The dominant polarization is linear (vertical) polarization. For the proposed example, the gain is around 19 dB at 10 GHz while the HPBW is equal to 19° and 20° in the H- and E-plane, respectively. The asymmetry of the beam amplitude in the two main planes is a common problem of pyramidal horn antennas. Another limitation is related to the diffraction arising from the horn flanges and, in particular, from those that are perpendicular to the electric field vector. In general, such diffraction produces back radiation and sidelobes which are indeed more evident in the E-plane.

Another important type of horn antenna is the conical horn whose geometry is shown in Figure 1.5. The conical horn aperture is circular and, in the most typical configuration, is fed by a section of circular waveguide which is typically excited by a TE_{11} mode. The behavior of a conical horn is similar to that of a pyramidal horn. The directivity can be expressed as [16]

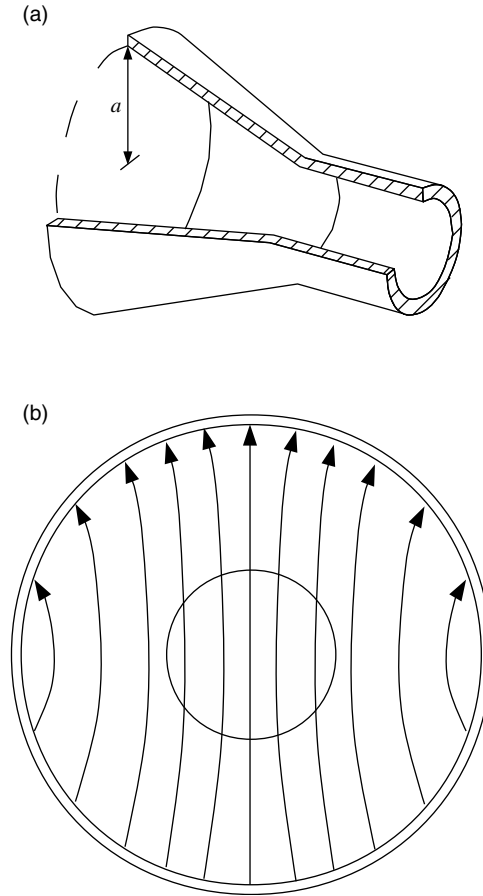


Figure 1.5 Conical horn antenna: (a) geometry; (b) electric field distribution of a smooth-walled conical horn aperture.

$$D_c(\text{dB}) = 10 \log_{10} \left[\epsilon_{ap} \frac{4\pi}{\lambda^2} (\pi a^2) \right] \quad (1.26)$$

where a is the aperture radius and ϵ_{ap} is the aperture efficiency. Although the conical horn is geometrically symmetric, its pattern is asymmetric and it suffers from similar limitations to that of pyramidal horns. In particular, the conical horn can present high cross-polar levels, which can be easily explained by looking at the transverse electric field distribution on the antenna aperture as shown in Figure 1.5b for vertical polarization. As can be observed, components of the electric field are also present along the y -axis. In the far field, such components would give rise to an electric field horizontally polarized with peaks of intensity at $\pm 45^\circ$. Poor polarization performance can be a severe limitation both in radio astronomy applications and in satellite communication systems as reported in Section 12.4.

The lack of symmetry in pyramidal and conical horn antennas can cause severe limitations in terms of efficiency, increasing losses when global coverage is required and generating spillover losses when horns are used as reflector feeds. A common way to improve the field distribution across the horn aperture is to

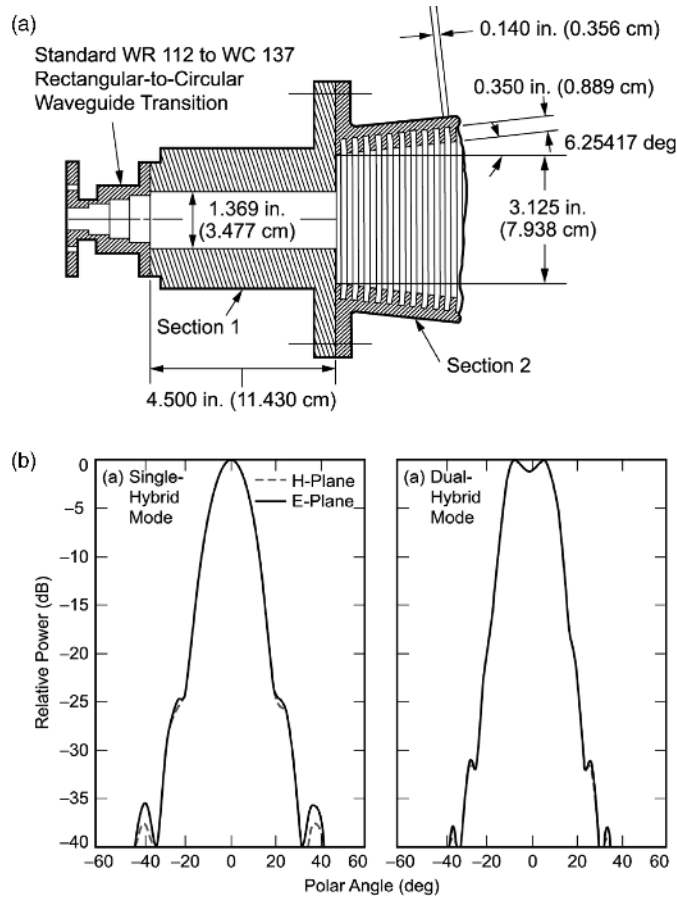


Figure 1.6 Dual-hybrid-mode feed horn: (a) geometry; (b) normalized radiation patterns [22].

employ grooved walls [17]. Corrugations perpendicular to the walls are designed to provide a capacitive reactance which inhibits surface wave propagation thus avoiding spurious diffraction from the edges. For pyramidal horns, the corrugations are usually placed only on the E-plane walls as edge currents on the H-plane walls are negligible. However, most corrugated horns are conical horns, this type of antenna being easier to fabricate. An example of a conical corrugated horn is shown in Figure 1.6a. As the groove response is polarization independent, the fundamental mode of a corrugated horn is the hybrid mode HE_{11} that can be associated to a combination of a TE_{11} and TM_{11} modes in a smooth circular waveguide. In general, the two modes are optimally phased to yield a highly symmetric field distribution across the aperture which, in turn, generates a symmetric radiation pattern ideally with very low sidelobes [18]. The performance of this type of radiator can be further optimized by using a Gaussian profiled conical horn [19]. In this case, the radius increases longitudinally following the expansion law of a Gaussian beam. As a result, the field distribution at the horn mouth is almost perfectly Gaussian, thus generating a far-field pattern ideally without sidelobes.

Another technique which can be employed to improve the horn pattern is to use a multimode approach. In this case, higher order modes can be deliberately excited with a specific phase and amplitude relationship, improving the horn radiation performance [20]. When even more demanding performance is required,

multi-hybrid-mode corrugated horn antennas can be designed as shown in [21] and in [22] for Deep Space Network antennas.

1.3.3 Reflectors

Reflector antennas are by far the most common antenna element for applications requiring high gain and directivity. This class of antennas has been widely employed in space missions since the early days of space exploration (see Section 7.2). Over the years, their concepts evolved both mechanically and functionally to meet technical requirements of increasing complexity. In this section only a basic review of this type of antenna is provided, the interested reader being directed to following chapters and to the referenced literature [16,22–25] for further study.

1.3.3.1 Main Reflector Parameters

Although reflector antennas can be made in different types, shapes and configurations, they all essentially consist of a passive reflecting surface illuminated by a smaller primary feed. Reflector antenna performance is influenced by several parameters, as follows.

Spillover and Aperture Illumination Efficiency Reflector efficiency is highly influenced by the feed radiation characteristics. In particular, an ideal reflector should be uniformly illuminated and all power should be focused on the reflecting surface. The portion of the feed power that does not reach the reflector is referred to as spillover loss while the ability to uniformly feed the parabola is referred to as illumination efficiency. Since primary feeds have a tapered radiation pattern, a compromise between spillover losses and illumination efficiency must be considered to maximize the aperture gain.

Aperture Blockage Feed and mechanical support structures located in front of the aperture, partially block field radiation in the far field. This phenomenon is referred to as aperture blockage and its main effect is to reduce the on-axis gain and to increase the sidelobe amplitude level. The reduction of efficiency due to aperture blockage varies depending on the feed configuration and aperture size.

Axial and Lateral Defocusing Axial and lateral defocusing are the errors generated by displaced feed positions along the reflector axis and orthogonally to the reflector axis respectively. Axial displacements generate a broader beamwidth while lateral defocusing causes beam squints [26,27].

Reflector Surface Deviation Deviations from the curvature surface cause a distortion of the reflector antenna radiation pattern [28]. The effect of surface deviation can be significantly high in deployable reflector antennas as outlined in Chapter 5.

Feed Feed selection and design have a major role in the correct and efficient operation of a reflector system. In general, the feed type depends on the system requirements in terms of frequency band, radiation characteristics and efficiency. Although simpler antenna types can be used, the best performance is usually achieved through horn antennas with Gaussian beam characteristics [29].

1.3.3.2 Basic Reflector Types

Some of the most common reflector systems are shown in Figure 1.7. The simplest form of reflector antenna is the parabolic reflector shown in Figure 1.7a. This configuration benefits from the geometrical properties of the

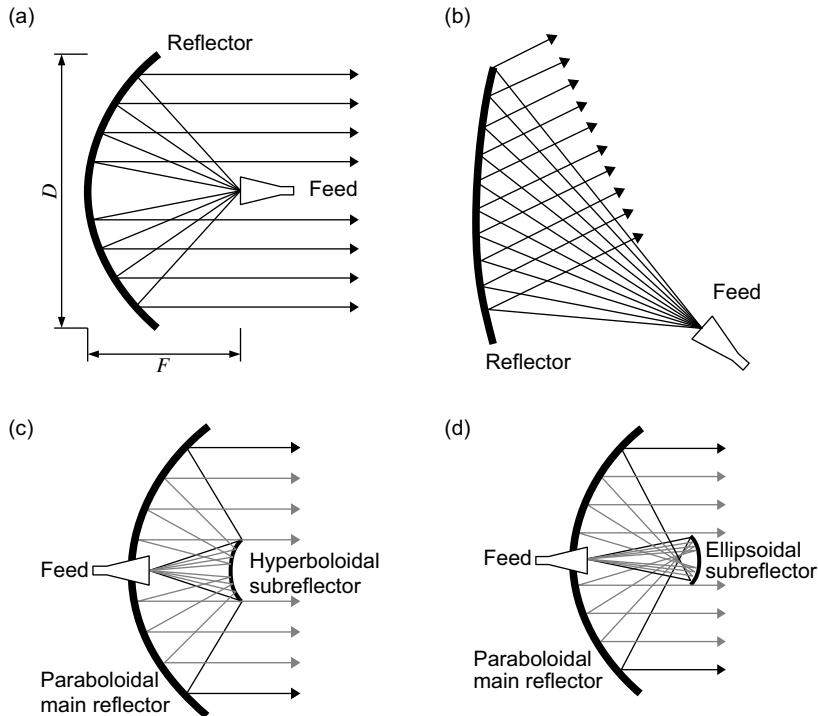


Figure 1.7 Reflector antenna configurations: (a) on-focus parabolic reflector; (b) off-axis reflector; (c) Cassegrain reflector; (d) Gregorian reflector.

parabola since spherical waves radiated by a source placed at the focal point are transformed into plane waves directed along the aperture rotation axis. This type of reflector generates a pencil beam whose characteristics are mainly controlled by the aperture diameter, D , the focal length, F , the reflecting surface curvature, F/D , and the pattern and size of the feed antenna. The electrical performance of this elementary reflector system is limited by the effect of aperture blockage [30]. As a possible solution to this problem, configurations employing an offset feed and a sectioned parabolic reflector [31] can be considered as illustrated in Figure 1.7b. In this case, the blockage effect of the feed is negligible and the direction of maximum radiation can be controlled by optimally shaping the reflector surface. The absence of feed blockage can be particularly important for those applications where multiple-feed systems are needed. Compared to the axisymmetric configuration, the main drawbacks of this type of reflector system are related to the large cross-polar fields for linear polarization [32]. Depolarization effects are due to reflector curvature and they can be reduced by selecting a relatively large F/D ratio [33]. However, when it is not possible to increase the reflector curvature, polarization rotation can be cancelled by using a polarization grid [34] or by optimally designing the primary feed [35]. When offset reflector antennas are illuminated by a circularly polarized primary feed, high cross-polar fields generate angular displacements of the main beam [32,36]. Beam squinting can be counteracted by using reflectors with large curvatures or by employing compensation techniques at feed level [37,38].

For larger apertures, a more compact feed arrangement can be realized by employing smaller subreflectors. Classical axisymmetric geometries for the Cassegrain and Gregorian reflector types are shown in Figure 1.7c and d respectively. In both systems, the primary feed is located on the rear of the main paraboloidal reflector. In the Cassegrain arrangement the subreflector is a section of a hyperboloid located within the focus of the main

reflector, while in the Gregorian configuration the subreflector is an ellipsoid located outside the focus of the main reflector. Both systems have similar electrical features but Cassegrain designs are more commonly used in satellite applications.

Shaped Reflectors Dual reflectors have higher efficiency and reduced sidelobes with respect to the on-focus fed parabolic reflector [39]. In particular, it has been demonstrated [40] that aperture efficiency can be improved by controlling the shape of the main and sub-reflector surfaces to improve aperture energy distribution. Varying the shape of the reflector surface has a direct impact on the illumination function which can be controlled in both amplitude and phase, thus reducing both spillover losses and illumination efficiency.

Cross-polarization Reduction Offset dual-reflector antennas can be designed to have very limited cross-polar components. In particular, the optogeometrical condition for eliminating cross-polarization [41] depends on whether the subreflector surface is concave or convex, on the eccentricity and on the angles of the axes of the main reflector surface and subreflector surface, and on the axis of the primary radiation.

Contoured- or Multiple-Beam Reflectors Contoured- or multiple-beam configurations can be obtained through specific offset dual-reflector arrangements. In the most conventional approach, contoured-beam patterns can be achieved using a multi-feed dual-reflector system [42]. In this case, the desired coverage contour is achieved by superposing overlapping spots generated by different feeds whose fields are then combined through a beam-forming network. This approach is also used when multiple beams have to be generated from a single antenna. In this latter case, individual beam-forming networks for each beam have to be implemented. Digital beam forming can also be employed for implementing beam scanning capabilities [43].

Alternatively, it is also possible to generate contoured beams by using a single feed and by shaping the reflector surface [44]. Shaped reflectors are the most common design approach for single-beam applications in satellite applications due to lower weight and lower spillover losses w.r.t. a single-feed design [45].

Deployable Reflector Antennas Reflector antennas have evolved significantly over the years, boosted by space-related research. In particular, significant improvements have been achieved in terms of aperture size through the employment of deployable structures which can be larger than 20 m, as described in Chapter 8. Space-related research continues to lead the technological development of reflector antennas as is evident from the list of future configurations reported in Section 18.4.

1.3.4 Helical Antennas

Helical antennas are widely used in satellite communication systems mainly because of their circular polarization and wide-band features. In its simplest form, a helical antenna consists of a conducting wire wound in the form of a helix as shown in Figure 1.8a. Generally, this type of antenna is fed through a coaxial transition and includes a ground plane. The radiation characteristics of this antenna and its input impedance depend on the helix diameter, d , on the wire diameter, t , on the pitch, p , and on the number of turns, N .

The helix antenna has different modes of radiation. In normal mode (or broadside mode) the helix length is short compared to the wavelength and its behavior is similar to a short dipole [16]. This type of antenna radiates in directions normal to its axis (Figure 1.8b) and can be designed to operate in linear polarization or circular polarization. In this configuration, the helix behavior is highly sensitive to the antenna dimensions.

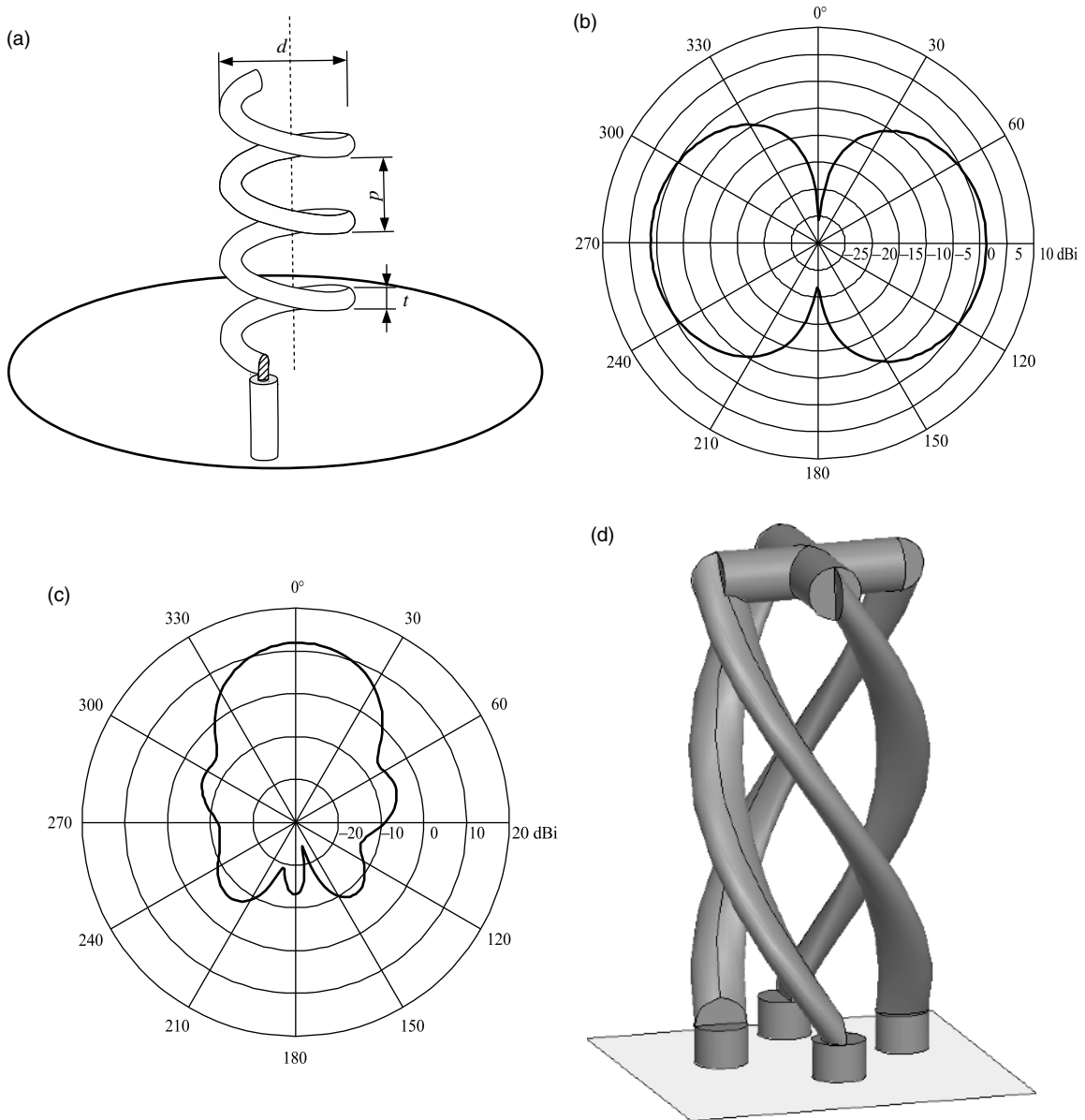


Figure 1.8 Helical antenna: (a) monofilar helical antenna geometry; (b) typical radiation pattern of a helical antenna operating in normal mode; (c) typical co-polar gain pattern of a helical antenna operating in axial mode; (d) short-circuited quadrifilar helical antenna (QHA).

In axial mode (or end-fire mode) the helical antenna has a main lobe directed along its axis, as shown in Figure 1.8c. This operating mode is achieved when both the helix diameter, d , and the pitch, p , are large fractions of the wavelength [46]. Helical antennas operating in axial mode are circularly polarized and they are normally installed on a ground plane. However, when the diameter of the ground plane of a conventional

helical antenna is less than the diameter of the helix, the helix radiates with its main beam in the backfire direction when the pitch angle is small [47].

The helix radiation characteristics can be controlled by changing the geometrical parameters of the antenna or by varying the number of wires [48–50]. For example, quadrifilar helical antennas (QHAs) (Figure 1.8d) are widely used for TT&C [51]. QHAs consist of four helical wires equally spaced and circumferentially located 90° apart from each other and sequentially fed with 90° of phase shift.

1.3.5 Printed Antennas

In the past few decades, microstrip antennas [52] have been one of the most commonly used antennas for space applications and, in all likelihood, will play a key role also in the coming years. In its most classical configuration, a microstrip radiator consists of a metallic patch element printed on a thin insulating dielectric layer placed above a ground plane. Figure 1.9 shows the two most popular microstrip antenna configurations: the rectangular patch antenna and the circular patch antenna. Since their first introduction [53,54], printed antennas have become a very popular research topic gaining the attention of both the industrial and the academic communities. Thousands of papers have been published on this subject, introducing many improvements and contributing to a rapid evolution of the early concept and widespread diffusion in many applications.

1.3.5.1 Features and Limitations

The diffusion of microstrip radiators is mainly due to their unique features, which are outlined below. Microstrip antennas are very low profile, of light weight and can be conformal to the mounting surface. These characteristics can be extremely important in several military, commercial or space applications where

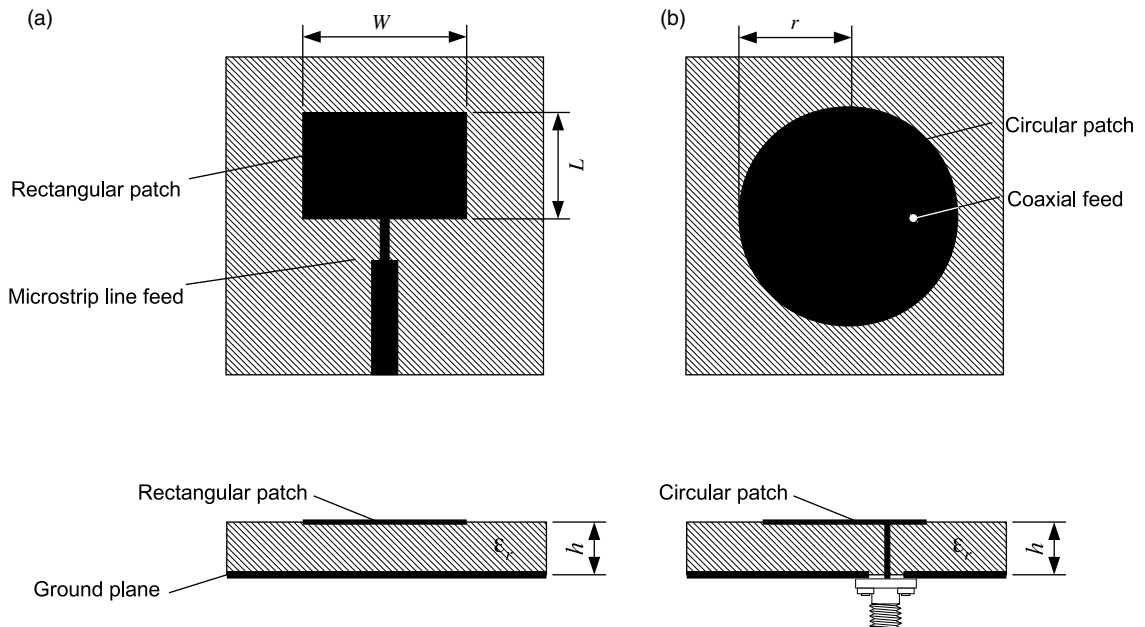


Figure 1.9 Basic types of microstrip antennas: (a) rectangular patch; (b) circular patch.

physical constraints are of prime concern. Depending on the type of materials, on the configuration and on the required fabrication process, microstrip antennas can also be low cost when compared to other types of antenna elements. Microstrip technology is naturally flexible, making possible the design of antennas of different shapes and configurations using single or multilayer arrangements and covering multiple bands. Furthermore, integration of printed antennas in microwave integrated circuits (MICs) is straightforward and high degrees of integration levels can be reached.

The main operational limitations of microstrip antennas are due to their narrow bandwidth. Indeed, a classical microstrip antenna would normally have a bandwidth of a few percent. Moreover, when compared to other radiators (e.g., horns, reflectors), the efficiency of microstrip antennas is much lower and the gain of a single patch is usually around 5–7 dBi. Another major disadvantage of printed radiators is related to their low power handling capability. This limitation is due to the small distance between the radiating patch and the ground plane. Depending on the substrate material characteristics and thickness, and on the thickness of metal layers, a microstrip radiator can be designed to handle hundreds of watts [55]. However, due to the multipacting breakdown effect [56], microstrip power handling in space is significantly reduced with respect to the expected value of the Earth's atmosphere. This aspect will be reconsidered in Section 1.5.1.

1.3.5.2 Basic Characteristics

In this subsection a rectangular patch antenna is taken as reference to discuss the basic radiation characteristics of microstrip antennas. A rectangular patch antenna consists of a rectangular patch of width W and length L printed on a substrate having relative dielectric permittivity ϵ_r of thickness h as shown in Figure 1.9. Generally, dielectric thickness is a fraction of wavelength ($0.003 \leq \lambda_0 \leq 0.05$ where λ_0 is the free-space wavelength) [16] while metal layers are tens of microns thick. The relative dielectric constant depends on the type of dielectric material. It mainly influences the resonant patch length L , the bandwidth and the patch efficiency.

A microstrip antenna designed to operate in its fundamental mode can be related to a half-wavelength resonator with two radiating edges. As can be observed in Figure 1.10, the electric field distribution at the patch radiating borders can be associated to that of two slots. This equivalence is the basis of the so-called transmission line model [57–59] that is the most intuitive way to represent a rectangular patch antenna. Yet this model does not capture many important physical phenomena which take place on a rectangular patch antenna. One of the effects which is not included in the transmission line model is the far-field radiation of the so-called non-radiating edges that are the patch borders orthogonal to the feed line axis (Figure 1.9a). The electric field associated to these borders for the fundamental mode is shown in Figure 1.10. It can be demonstrated that their

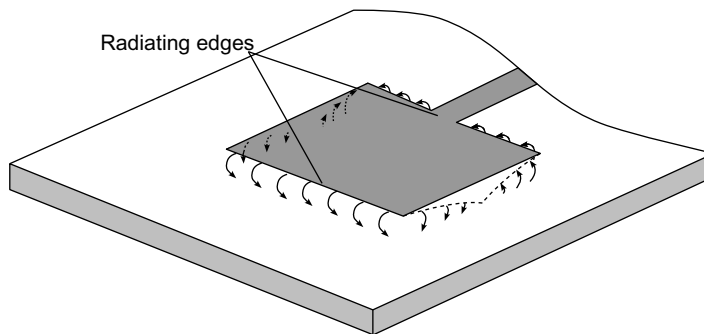


Figure 1.10 Electric field distribution at the edges of a rectangular patch antenna excited in its fundamental mode.

contribution to the radiation pattern on the H- and E-plane is virtually null [16]. A more accurate analytical representation can be obtained by treating the antenna region as a cavity bounded by electric conductors (patch and ground plane) and by magnetic walls along the perimeter of the patch. Although the cavity model provides a more realistic depiction of microstrip antenna behavior for different radiator shapes, it is normally used only for a first rough approximation of the antenna geometry or to understand design principles and physical insights. Indeed, the most common design approach is based on one of the commercially available simulators which make use of the full-wave techniques discussed in Chapter 2 of this handbook.

The patch configurations shown in Figures 1.9 and 1.10 radiate a linearly polarized field. In general, the polarization purity of a microstrip radiator is poor, as discussed in detail in Chapter 14. Patch configurations with improved linear or circular polarization performance will be presented later on in this subsection.

Typical radiation patterns of a rectangular patch are shown in Figure 1.11. In general terms, microstrip radiators are wide beam antennas. Their radiation performance is directly related to the equivalent magnetic current densities at the patch borders. For a given resonant frequency and dielectric material, the patch length, L , cannot be modified. Directivity can be indeed slightly changed by controlling the patch width, W . Typical gain values for a standard single-patch radiator are usually in the range from 5 to 7 dBi. The antenna gain and, consequently, the efficiency are strongly influenced by the characteristics of the dielectric material and by metal losses. Another type of loss in a microstrip antenna is related to surface wave excitation. Surface waves are generated at the discontinuity between the substrate and the dielectric above the antenna (e.g., air or free space). Surface wave power propagates at the dielectric interface causing efficiency reduction, spurious radiation and diffraction from the ground plane border, and mutual coupling in array scenarios [60].

1.3.5.3 Feeding Techniques

The electromagnetic behavior of a microstrip antenna is strongly influenced by the feed techniques. Illustrations of the most common feeding methods are shown in Figure 1.12. Feeding techniques based on coaxial probes are implemented by soldering the outer connector of a coaxial cable to the ground plane and elongating the inner conductor to fit flush against the patch. This technique is usually implemented when the antenna has to be attached to a standard 50Ω coaxial probe. However, it is possible to use the same coaxial configuration also in multilayer microstrip circuits. The connector should be located on the patch E-plane axis and the position has to be selected to match the coaxial feed characteristic impedance. When the height of the dielectric is too high, the metal pin penetrating inside the substrate provides an inductive reactance which shrinks the bandwidth and makes this configuration unsuitable for thick structures. In general, pin inductance can be compensated by adding a capacitive load [61]. The vertical currents excited by the coaxial probe generate spurious radiation which is indeed evident by looking at the asymmetries present in the E-plane co-polar pattern of Figure 1.11a.

Another common technique for feeding microstrip antennas is to use a simple microstrip transmission line feed as shown in Figures 1.9a and 1.10. In this case, a microstrip transmission line is connected to the radiating border of a patch. In order to match the characteristic impedance of the microstrip line with the patch input impedance two approaches can be adopted: using an impedance transformer (e.g., quarter-wavelength transformer) or inserting the feed line inside the patch. Both radiating element and feed line are printed on the same layer. Although this configuration is simple to fabricate, the leakage radiation from the feed line can significantly deteriorate the radiation pattern. A similar phenomenon takes place also when proximity feed arrangements are used. In this configuration (Figure 1.12c), the feeding microstrip line is printed on an additional metal layer underneath the patch radiator. Another common feeding scheme is the aperture-coupled technique (Figure 1.12d) proposed in [62]. A microstrip line printed back to back with the patch radiator is coupled to the antenna by means of a slot on the ground plane. Slot coupling provides better bandwidth,

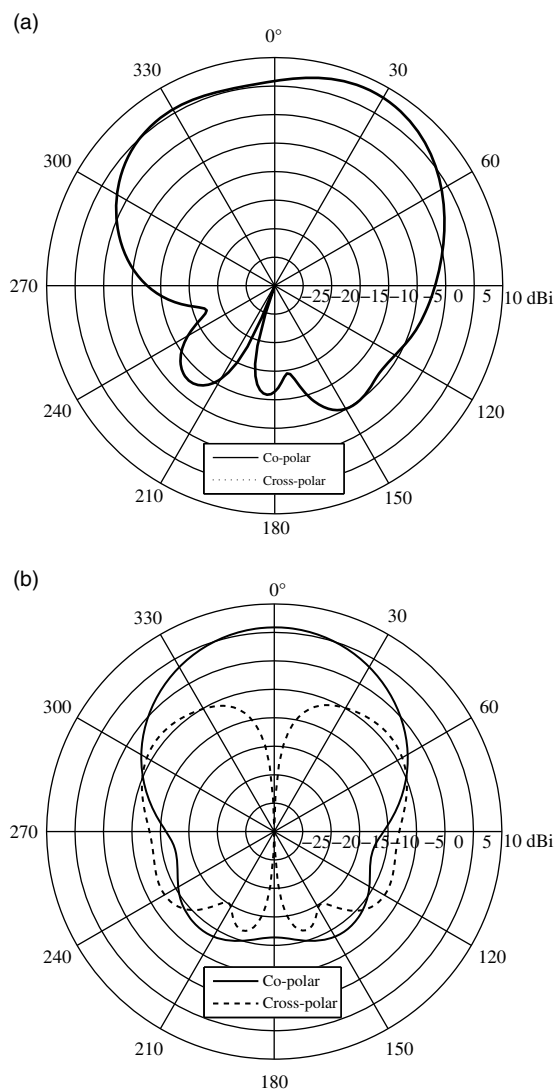


Figure 1.11 Typical radiation pattern of a rectangular patch antenna with coaxial feed: (a) E-plane and (b) H-plane co-polar and cross-polar gain.

minimizes spurious radiation from the microstrip lines and avoids vertical elements and soldering. The main limitation of this solution is related to possible unwanted radiation from the slot. Proximity coupling, aperture coupling and all other non-contact feeding techniques provide better performance in terms of passive inter-modulation distortion [4] (see Section 1.5.2 for details).

1.3.5.4 Materials and Fabrication Processes

The selection of the dielectric material is of key importance in the design of microstrip antennas, affecting mechanical, thermal and electrical performance. The dielectric substrate in a microstrip antenna mainly serves

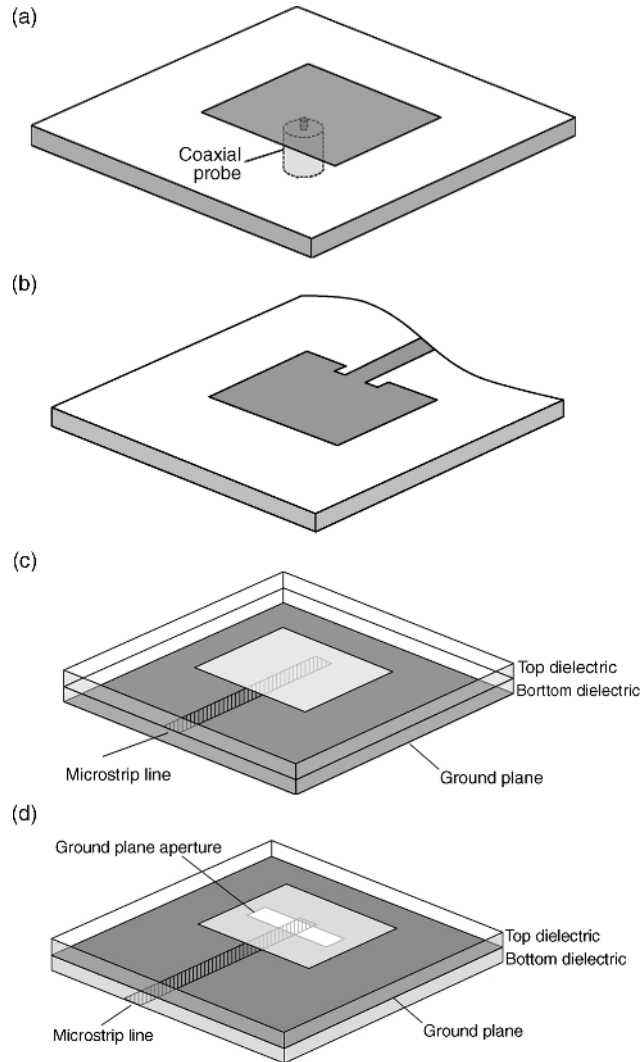


Figure 1.12 Common microstrip antenna feeding techniques: (a) coaxial probe; (b) microstrip line; (c) proximity coupling; (d) aperture coupling.

as mechanical support for the patch providing uniform spacing and mechanical stability. Lower values of relative dielectric constant (normally between 1 and 2) can be obtained by using polystyrene foam or honeycomb structures. Dielectrics based on fiberglass reinforced Teflon, also known as PTFE (polytetrafluoroethylene), typically provide a relative dielectric constant between 2 and 4. Higher values can be obtained through materials based on ceramic, quartz or alumina. However, these materials should be carefully employed as they provide a reduction in the patch size at the expense of radiation efficiency and mechanical stability. Another important selection driver for microstrip antenna substrates are dielectric losses. Acceptable dielectric losses are usually related to the application requirements and to the antenna architecture. In general, low tangent loss results in higher dielectric cost.

For a satellite microstrip antenna, the thermal behavior of the substrate and the temperature dependence of its main parameters are of primary importance. Indeed, a microstrip antenna mounted in a spacecraft operates under large ranges of thermal variations. At Earth-like distances from the Sun, temperatures of 273 ± 100 K can be expected, whereas larger ranges can be expected for interplanetary missions. For example, variations in the relative dielectric constant have a direct effect on the antenna operating frequency. For this reason, antenna bandwidth is usually evaluated through a sensitivity study including temperature effects. Material behavior in a space environment is discussed in Chapter 4 while basic effects such as multipaction and outgassing are described in Section 1.5. Aspects related to thermal conductivity, heat dissipation and mechanical stability are discussed in detail in Chapter 5.

In the last few years, the complexity of microstrip antennas has constantly increased. In particular, multilayer configurations with a high integration level and many vertical transitions are becoming evermore popular. In general, the development of multilayer assemblies can be particularly difficult as layers with different materials are usually employed for the circuit elements and for the antennas. Indeed, dielectrics of low dielectric constant are preferred for radiating structures whereas materials with high permittivity are usually employed for microwave circuits. This difference usually results in different coefficients of thermal expansion which can generate mechanical deformations of the multilayer structure. For this reason, new types of materials and fabrication processes are attracting the attention of many researchers. In particular, interesting results [63] have been obtained with liquid crystal polymers (LCPs). LCPs are low dielectric-constant, low loss-tangent [64] materials with very good package hermeticity and a low cost [65]. LCP has gained attention especially as a potential high-performance microwave substrate and packaging material for multilayer arrays and for highly integrated circuits [66].

Another interesting solution for multilayer antennas is the usage of low-temperature cofired ceramic (LTCC). This technology allows the implementation of flexible multilayer configurations with a high integration density and many vertical transitions. LTCC multilayer circuits are produced by firing in a single laminate multiple tape layers where conductive, dielectric and/or resistive pastes are selectively deposited to form transmission lines, resistors, inductors, and so on [67]. Although LTCC is mainly used for microwave integrated circuits, several interesting results have been obtained also for antenna elements [68–72]. LTCC antenna examples will be discussed in Chapter 11 along with examples of other emerging technologies for on-chip and in-package antenna integration.

1.3.5.5 Microstrip Antenna Configurations

It is very difficult to enumerate all the possible microstrip antenna configurations as new designs are proposed in every issue of specialized journals. In this subsection we discuss only a basic subset of possible configurations mainly related to dual or circular polarization operation and to multiband or wide-band applications.

Dual and Circular Polarization Operation In dual-polarized microstrip antennas two orthogonal modes have to be excited. Excitation can be actuated by two orthogonal feeds as shown in Figure 1.13a. Each feed is designed to excite a single mode and to be isolated as much as possible from the orthogonal mode. Ideally, orthogonal location of the feeds accomplishes high isolation between the two ports because each feed is located in the area where fields of the orthogonal modes are virtually null. However, in practice feed coupling is one of the major challenges in the design of dual-polarized antennas. Dual-feed arrangements provide narrow-band performance and are well suited only for thin substrates. An improved frequency response can be achieved by using two oppositely located feeds for each mode as shown in Figure 1.13b. The use of two feeds with 180° of phase difference reinforces the polarization mode, helps cancel unwanted feed radiation and

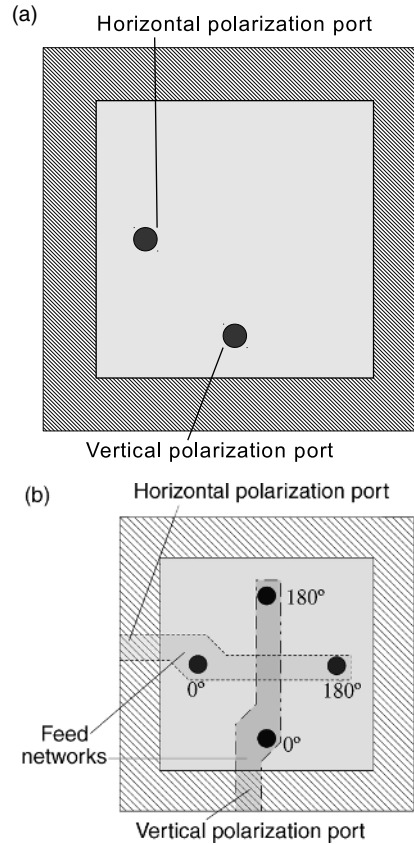


Figure 1.13 Dual-polarization generation in a square microstrip antenna: (a) dual feed; (b) four-feed configuration.

suppresses higher order modes of thick substrates [73]. Examples of dual-polarized antennas for SAR applications are reported in Chapter 13.

A microstrip antenna is circularly polarized when it radiates two orthogonally polarized electric fields with $\pm 90^\circ$ of phase difference between them. CP excitation is thus achieved by exciting two orthogonal modes in a patch radiator. All different circular polarization techniques for microstrip antennas can be grouped into two main classes: perturbative and multi-feed. Figure 1.14 shows examples of both techniques. To the first class of architectures belong single-feed microstrip antennas where perturbation of the patch shape is used to excite orthogonal modes with $\pm 90^\circ$ of phase difference. Typical single-feed CP configurations include square patches with truncated diagonal corners, circular patches with notches, elliptical patches or rectangular patches [16]. Although simple to fabricate, this configuration provides narrow-band CP performance.

In the second case, circular polarization is enforced by exciting a patch through multiple feeds orthogonally located and with an appropriate phase difference. Such a technique usually provides higher polarization purity, suppresses higher order modes and provides wider bandwidth. The main drawback is related to mutual coupling between multiple feeds and to the feeding network size and complexity. Examples of CP GNSS

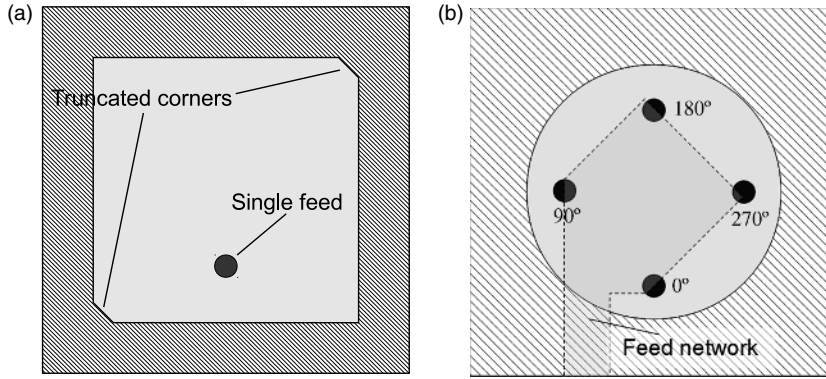


Figure 1.14 Circular polarization generation in microstrip antennas: (a) perturbative technique example, square patches with truncated corners; (b) multi-feed example: four-feed LHCP Rx antenna.

microstrip antennas are discussed in Chapter 14 while examples of polarization enhancement techniques at array level are described in Chapter 9.

Bandwidth Enhancement Techniques Bandwidth enhancement of microstrip antennas is usually required for many practical applications. Bandwidth can be increased by lowering the Q factor of the microstrip antenna. This can be achieved by using thick substrates or by using materials with low relative dielectric constant. In both cases, insurgence of higher order modes should be carefully considered. Another common bandwidth enhancement technique is to use radiators with multiple contiguous resonances. This method can be implemented by using parasitic stacked patches [74] or through reactive loading by means of shaped slots, notches, cuts, pins or posts. Wide-band can be achieved also at feed level by designing broadband matching networks [75] or by means of reactive feeds such as L-shaped probes [76–78].

1.4 Arrays

Antenna arrays are a set of antennas arranged to provide highly directive patterns. They yield an increment in the aperture area which can be controlled geometrically and electrically by optimally setting the location and the excitation of the array elements. The geometry of an arbitrary array of N elements is represented in Figure 1.15. In its most general form, the radiation pattern of an arbitrary array of N antennas can be written as

$$E_{total}(\theta, \phi) = \sum_{n=1}^N A_n F_n(\theta_n, \phi_n) e^{-j(k_0|r_n| + \beta_n)} \quad (1.27)$$

where:

- $E_{total}(\theta, \phi)$ is the total far field radiated by the array in the (θ, ϕ) direction
- A_n is the amplitude factor of the n th array element

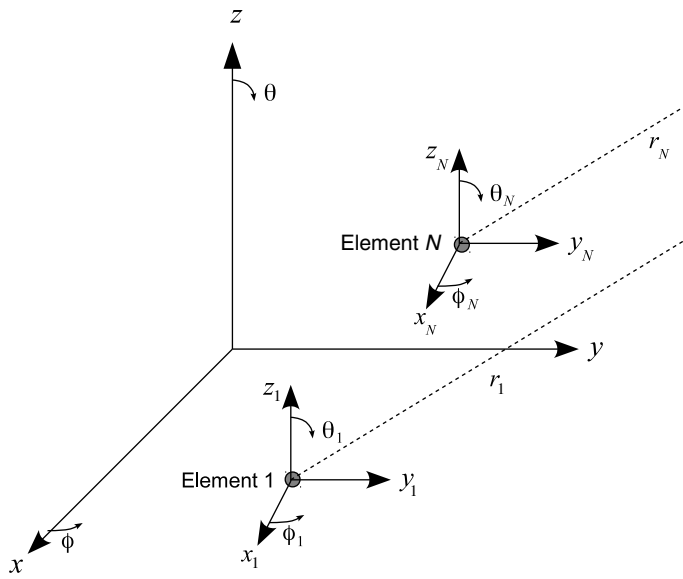


Figure 1.15 Far-field geometry of an arbitrary array of N elements.

- $F_n(\theta_n, \phi_n)$ is the pattern function of the n th array element defined w.r.t. the local coordinate system
- k_0 is the free-space propagation constant
- $|r_n|$ is the distance from the observation point of the n th array element
- β_n is the phase of the n th array element.

More simple and compact expressions of the array pattern can be obtained when identical antennas are employed and regular geometries are adopted. In these cases, the total pattern can be decomposed into two contributions: the element pattern, $E_{element}(\theta, \phi)$, and the so-called array factor, $AF(\theta, \phi)$:

$$E_{total}(\theta, \phi) = E_{element}(\theta, \phi)AF(\theta, \phi) \quad (1.28)$$

The array factor is a function of the array geometry, the inter-element distance, the element excitation in amplitude and phase, the number of elements and the frequency. The simplest case is that of the so-called uniform array: a ‘linear array of identically oriented and equally spaced radiating elements having equal current amplitudes and equal phase increments between excitation currents’ [1]. For a uniform linear array of N elements arranged along the z -axis, with inter-element distance d and progressive phase shift β , the array factor can be written as

$$AF(\theta, \phi) = \frac{\sin(N\psi/2)}{\sin(\psi/2)} \quad (1.29)$$

where $\psi = k_0 d \cos\theta + \beta$. When $\psi = 0$ the array factor has a maximum which corresponds to the main array lobe. Other array factor maxima are found when $\psi/2 = \pm r\pi$ with $r = 1, 2, \dots$. Lobes corresponding to these

maxima are referred to as *grating lobes* and they have the same array factor amplitude of the main lobe. As a consequence, directions of maximum amplitude of the array factor can be written as

$$\theta_{\max} = \arccos[(\lambda/2\pi d)(-\beta \pm 2r\pi)], \quad r = 0, 1, 2, \dots \quad (1.30)$$

Both the main lobe ($r=0$) and grating lobe directions depend on the array spacing, d , and on sequential phase shift, β . However, grating lobes appear only when the argument of the cosine in Equation 1.30 is less than in module 1. For example, for a uniform linear array with equiphase distribution ($\beta = 0$) the main beam direction is $\theta_{\max, r=0} = 90^\circ$, while grating lobes do not appear for inter-element spacings of less than a wavelength. For $d = \lambda$ grating lobes are in the visible region at $\theta_{\max, r=1} = 0^\circ$ and 180° . In general, grating lobes are an unwanted effect that has to be carefully avoided, especially in phased-array scenarios.

In general, the geometry of the array and its excitations should be defined through a synthesis procedure where the starting point is a set of given requirements specified on the array radiation pattern [79–81]. The target of the synthesis process is to find an array geometry and excitation distribution to suitably approximate the desired pattern.

1.4.1 Array Configurations

In this subsection a basic review of the most common array types is presented.

1.4.1.1 Direct Radiating Arrays (DRA)s

The simplest case is that of an array of regular geometry whose elements are excited through a beam-forming network (BFN). BFNs distribute and/or collect the power of individual elements to a single port. Power distribution is accomplished so that each element receives the signal with a desired amplitude and phase. In their most typical form, BFNs are implemented through binary power combiners/dividers as shown in Figure 1.16a. Usually, DRA efficiency is limited by BFN losses which become evident in arrays with a large number of elements. In some cases, it is possible to obtain good radiation performance by selectively removing some of the DRA elements. This configuration is referred to as a thinned array [82–85]. In other cases, a synthesis technique can be used to design a so-called sparse array whose elements are uniformly excited and located in a non-regular grid [82,83,86–88].

1.4.1.2 Phased Arrays

In many applications, the main antenna beam has to be moved dynamically to point in different directions. Although mechanical scanning is possible under certain conditions, a common solution is represented by phased arrays. In phased-array antennas beams are formed by shifting the phase of the signal emitted from each radiating element. An example of a corporate-fed linear phased array is shown in Figure 1.16b. In this case, the direction of maximum radiation, $\theta_{\max}$ in Equation 1.30, is controlled by changing the phase of each array element, φ_i , so that different progressive phase shifts, β , can be achieved. A common problem in large phased arrays with wide-band elements is referred to as scan blindness. This phenomenon is generated by inter-element mutual coupling. Indeed, the active impedance of a given radiator in the array changes as a function of the amplitude and phase distribution across the entire array. As a result, it can happen that when the array is scanned, at certain angles the module of the reflection coefficient of the array rapidly increases to 1 and the

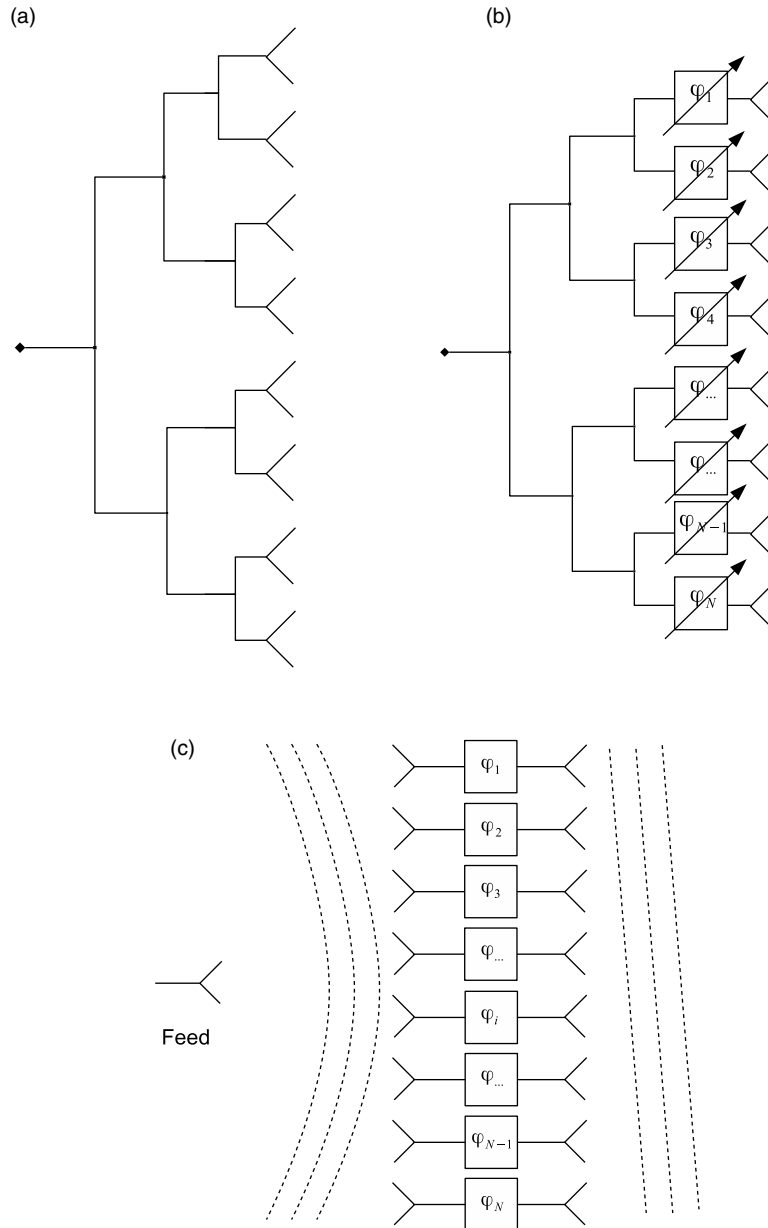


Figure 1.16 Array configurations: (a) direct radiating array (DRA); (b) phased array; (c) transmit array.

array pattern forms a null [89,90]. Therefore, scan blindness limits the scan range and lowers the antenna efficiency.

This book discusses several examples of phased arrays for space applications. Some of them are presented in Chapter 9 ‘Microstrip Array Technologies for Space Applications’, in Chapter 10 ‘Printed Reflectarray Antennas for Space Applications’ and in Chapter 13 ‘Antennas for Spaceborne Synthetic Aperture Radar’.

1.4.1.3 Reflectarrays

Reflectarrays [91] are flat reflectors illuminated by an external feed. The reflecting surface is realized through a spatially fed antenna array as shown in Figure 10.1 for the case of a printed reflectarray. In other terms, reflectarrays are discrete flat reflectors where the reflected field is controlled by each element of the array which should be designed to reradiate the incident field at a proper phase. A detailed review of reflectarray antennas for satellite applications is given in Chapter 10 of this handbook.

1.4.1.4 Transmit Arrays

Transmit array antennas (Figure 1.16c), also called lens antennas, are planar discrete lenses that operate a phase-front transformation by converting an incident spherical wavefront into an outgoing plane wave propagating in a specified direction. Transmit array antennas are an attractive solution for achieving high gain at millimeter wave frequencies where the free-space feeding improves the radiation efficiency by eliminating the losses that occur with corporate feed networks. With respect to reflectarrays, they are inherently less affected by surface errors if employed in deployable configurations.

1.5 Basic Effects of Antennas in the Space Environment

Satellites must survive for the duration of their mission with the required stability and survivability. For this reason, mechanical and thermal behavior as well as material characteristics are a major issue for all satellite components. From this point of view, since antennas are located in the external satellite body, they are highly exposed to radiation and to thermal variations. In this section we will provide a basic description of the most important phenomena involving antennas in space. An in-depth description of specific space environmental threats to space antenna constituent materials is presented in Chapter 4 while the mechanical and thermal behavior of satellite antennas is provided in Chapter 5. Moreover, Chapter 6 presents a detailed description of the most important tests required for space antenna assessment before flight.

1.5.1 Multipaction

Multipaction [92], or multiple impaction, is a resonance type of discharge that occurs between two electrodes with RF fields, usually in a vacuum or low-pressure condition. In a high-vacuum environment, an electron may have a free path that is larger than the gap between the electrodes. When the electron, accelerated by the electric field, collides with the electrode it might cause the emission of secondary electrons from the material surface. If the impact energy, the frequency and the distance between the two conductors are favorable, a resonant multiplication of the number of electrons takes place resulting in operational impairments and potential physical damage. The generation of multipaction depends upon several constraints [93]:

1. the vacuum condition (*usually lower than 10^{-3} torr*);
2. the applied RF voltage (*depends on the material and on the angle of incidence of the primary electrons*);
3. the electrode geometry and the operating frequency (*the gap size should correspond to a multiple half cycle of the applied RF voltage to satisfy the condition of electron resonance*);
4. the material surface (*material contaminations or impurities enhance the possible occurrence of multipaction*).

In RF space systems, multipaction limits the power handling capability of RF systems and can cause loss and distortion of the RF signal (increase in noise figure or bit error rate) as well as damage to RF components or subsystems due to excess RF power being reflected back or dissipated by them [94].

Multipactor prevention represents a very significant problem in the design and implementation of high-power antenna feed networks for communications satellites. In general, multipaction can be avoided by optimally modifying the device geometries, by laminating electrodes with particular materials or by reducing the RF power levels. For instance, in a microstrip antenna, the multipaction effect can be estimated by assuming that the two electrodes are the patch and its ground plane. High power handling cannot be achieved when thin dielectrics are used. Design rules or RF design tools such as the Multipactor Calculator [95] developed by the European Space Agency (ESA) can be used to achieve design margins that preclude the onset of multipaction.

1.5.2 Passive Inter-Modulation

Passive inter-modulation (PIM) is comparable to the phenomenon taking place in active devices due to their inherent nonlinearity when two or more RF carriers are mixed in a passive system and form unwanted signals. Nonlinearity in a passive device such as an antenna can be caused by several physical effects which can be grouped into two categories: contact nonlinearity and material nonlinearity. Causes of contact nonlinearity include the formation of a junction capacitance due to thin oxide layers between conductors, the presence of contaminating particles or mechanical imperfections on the surfaces, the tunnel/Schottky effect, contact resistance caused by two dissimilar metals or loose metal-to-metal contacts (some metals, like stainless steel, are more susceptible than others, like Al or Ti alloys) [93]. Material nonlinearities are generally due to hysteresis effect in ferromagnetic materials or to insufficient thickness of plated metal, causing RF heating. Contact sources can be excited by relatively low energy. For this reason, the accuracy of the manufacturing and assembly process as well as materials employed in a satellite RF system have to be accurately validated in an attempt to detect and eliminate any possible PIM source. A material selection guide for PIM reduction is proposed in Section 4.3.3.

The PIM interference caused by antenna nonlinearity has a serious impact on the performance of high-power multi-frequency communication systems, especially when the antenna is shared by the transmitter and the receiver at the same time [96]. Typical satellite communication systems are designed to avoid PIM orders below VII or XI. However, passive inter-modulation evaluation is not entirely amenable to theoretical design and experimental assessment is required in most cases.

1.5.3 Outgassing

Outgassing is generated by pockets of gas trapped or absorbed in materials or components during manufacture. In a vacuum environment, trapped gas evolves causing a material to lose volatile mass particles acting as contaminants to other surfaces and harming the satellite.

All materials used for space flight applications should satisfy the outgassing requirements [97–99] recognized by NASA, ESA and other space agencies. Contamination requirements are usually expressed in terms of total mass loss (TML), collected volatile condensable materials (CVCMS) and recovered mass loss (RML). Typical acceptance criteria are $TML < 1.0\%$, $CVCMS < 0.10\%$ and $RML < 1.0\%$. Screening test data of materials that have successfully passed the NASA outgassing standard is listed in the database ‘Outgassing Data for Selecting Spacecraft Materials’ [100]. Materials included in this database are approved for use in a space environment unless more stringent constraints are required by a specific application.

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