EENG 470
Satellite Communications

Lecture # 7
Chapter 3 : Antennas Fundamentals + Satellite Antennas

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Coordinate System

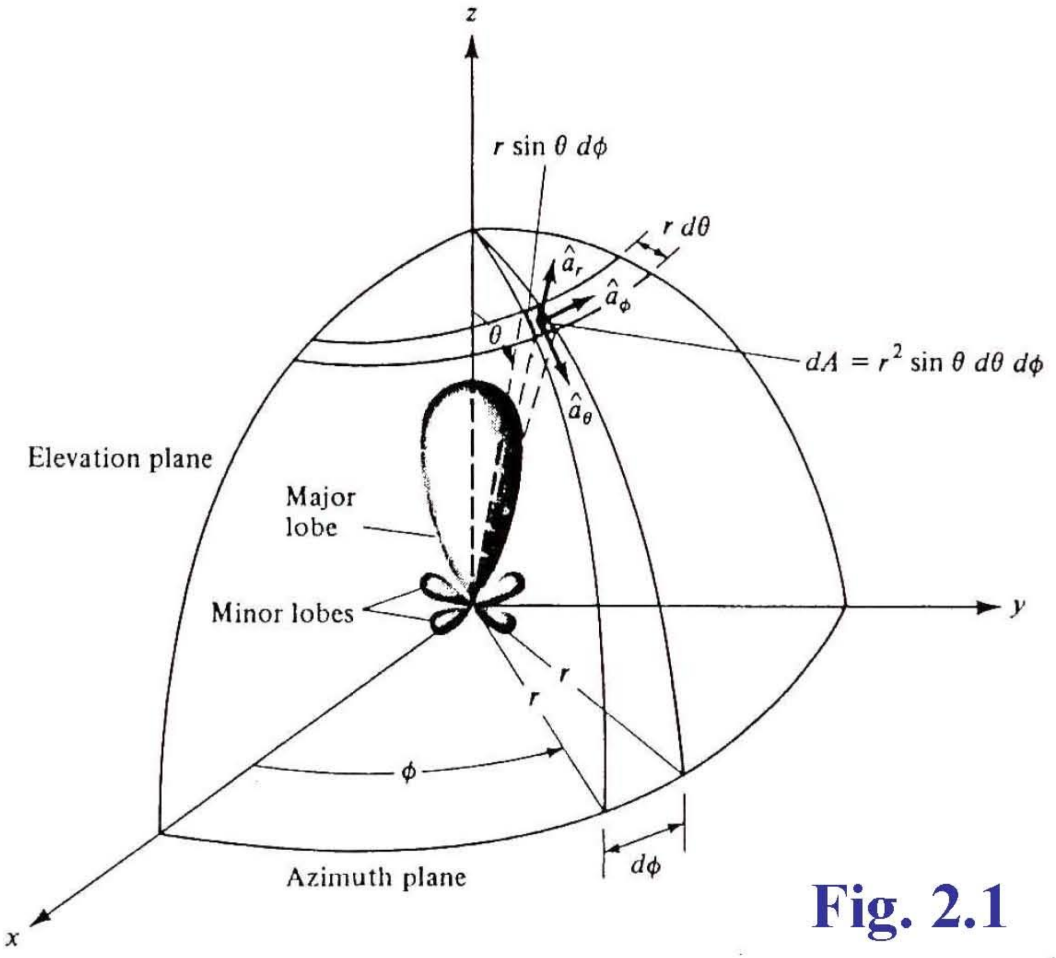


Fig. 2.1

$$\int_0^{2\pi}\left[\int_0^{\pi} r^2 \sin\theta \, d\theta\right] d\phi = 4\pi r^2$$

$$0 \le \theta \le \pi$$

$$0 \le \phi \le 2\pi$$

✓

Radiation Pattern

A mathematical and/or graphical representation of the radiation properties of an antenna, such as the:

- amplitude
- phase
- polarization, etc.

as a function of the angular space coordinates θ, ϕ.

Amplitude Radiation Pattern

- Field Pattern:

  A plot of the field (either electric $|\underline{E}|$ or magnetic $|\underline{H}|$) on a *linear* scale

- Power Pattern:

  A plot of the power (proportional to either the electric $|\underline{E}|^2$ or magnetic $|\underline{H}|^2$ fields) on a *linear* or *decibel (dB)* scale.

2-D Normalized *Field* $|\underline{E}_n|$ Pattern of a Linear Array

Linear Scale

N = 10 elements

$d = \lambda/4$ spacing

HPBW = 38.64°

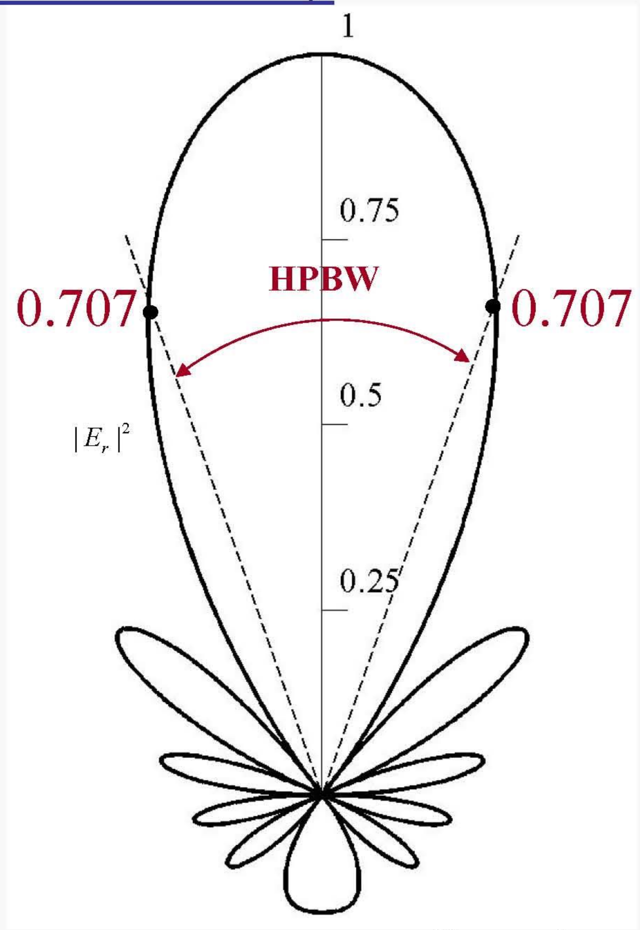


Fig. 2.2(a)

2-D Normalized *Power* $|\underline{E}_n|^2$ Pattern of a Linear Array

Linear Scale

N = 10 elements

$d = \lambda/4$ spacing

HPBW = 38.64°

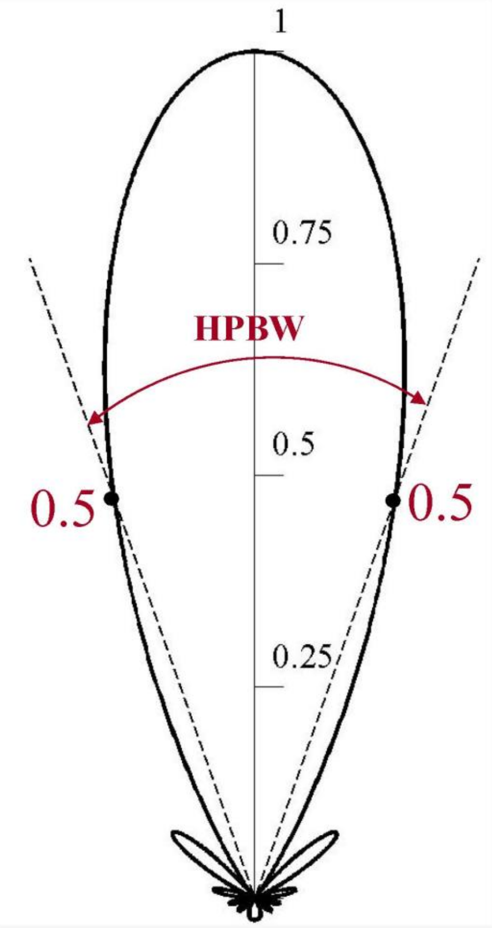
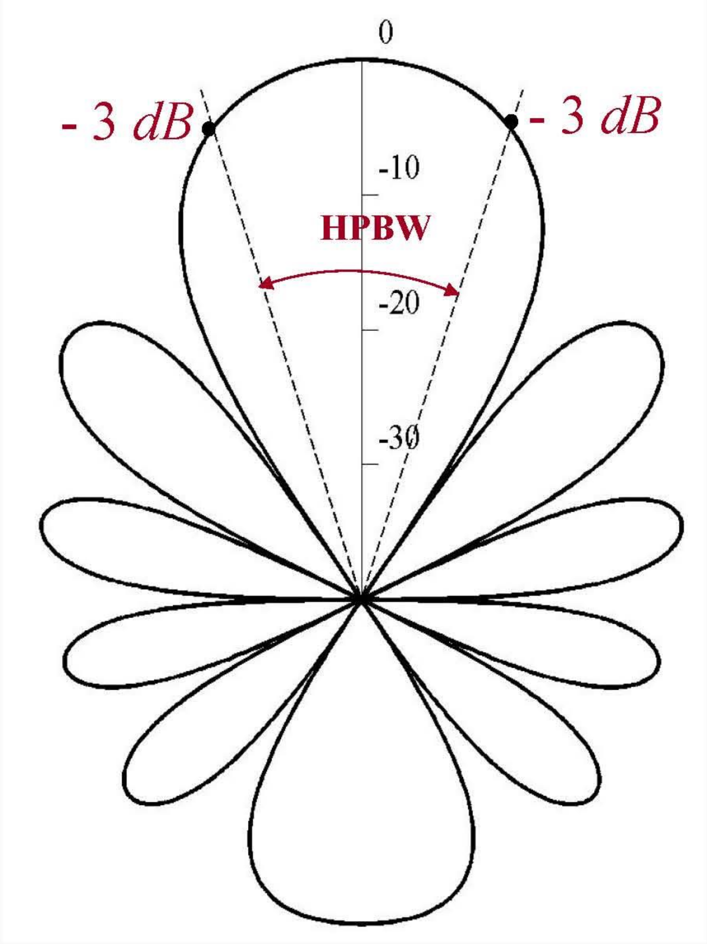


Fig. 2.2(b)

2-D Normalized *Power* $|\underline{E}_n|^2$ Pattern of a Linear Array



dB Scale

N = 10 element

$d = \lambda/4$ spacing

HPBW = 38.64°

Fig. 2.2(c)

Polar Pattern

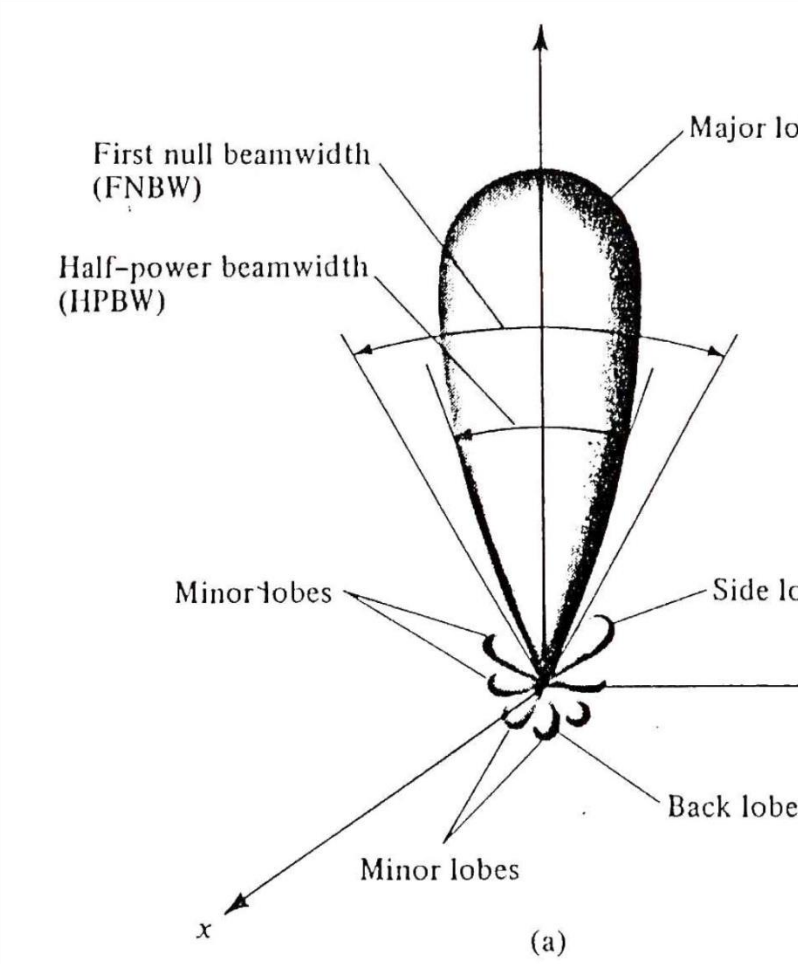


Fig. 2.3a

Linear Pattern

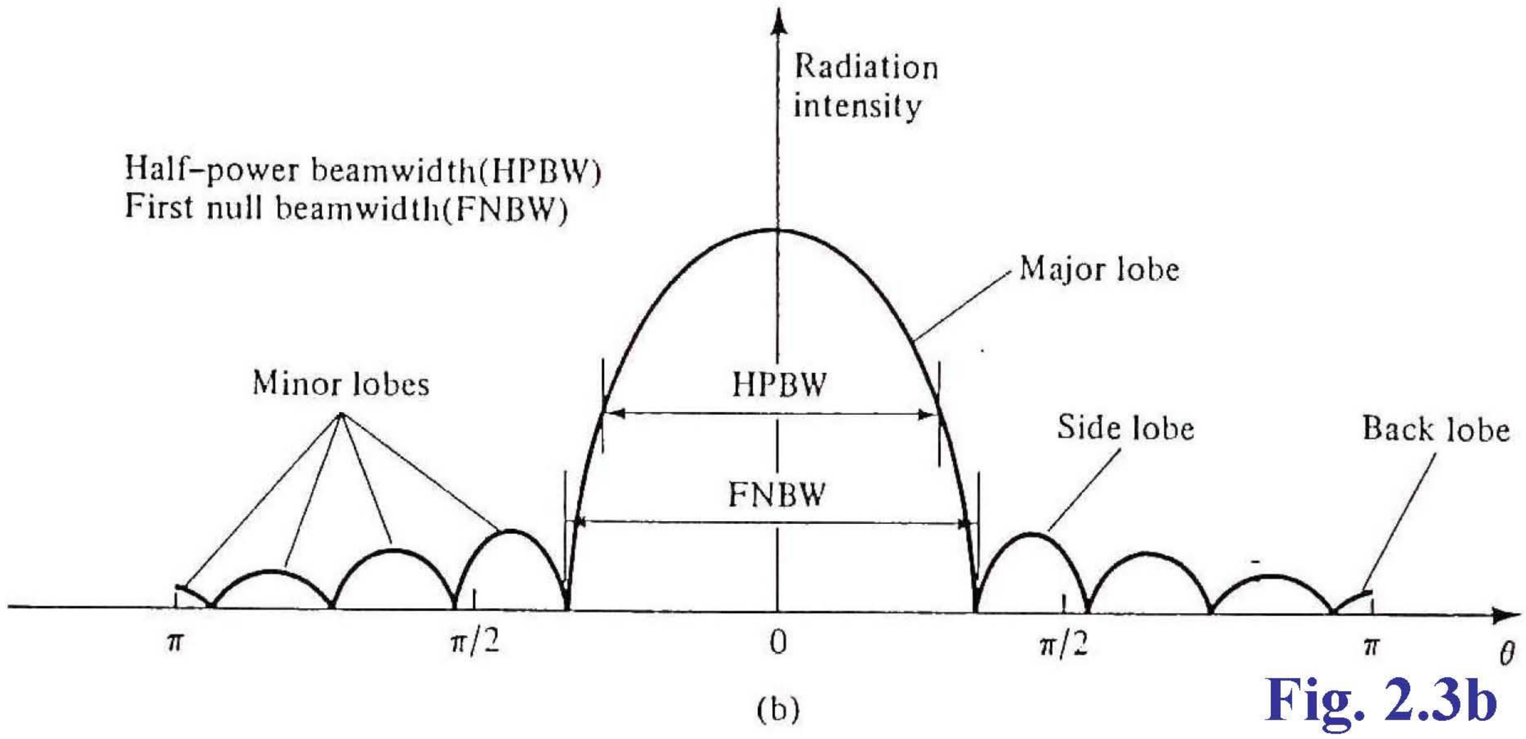
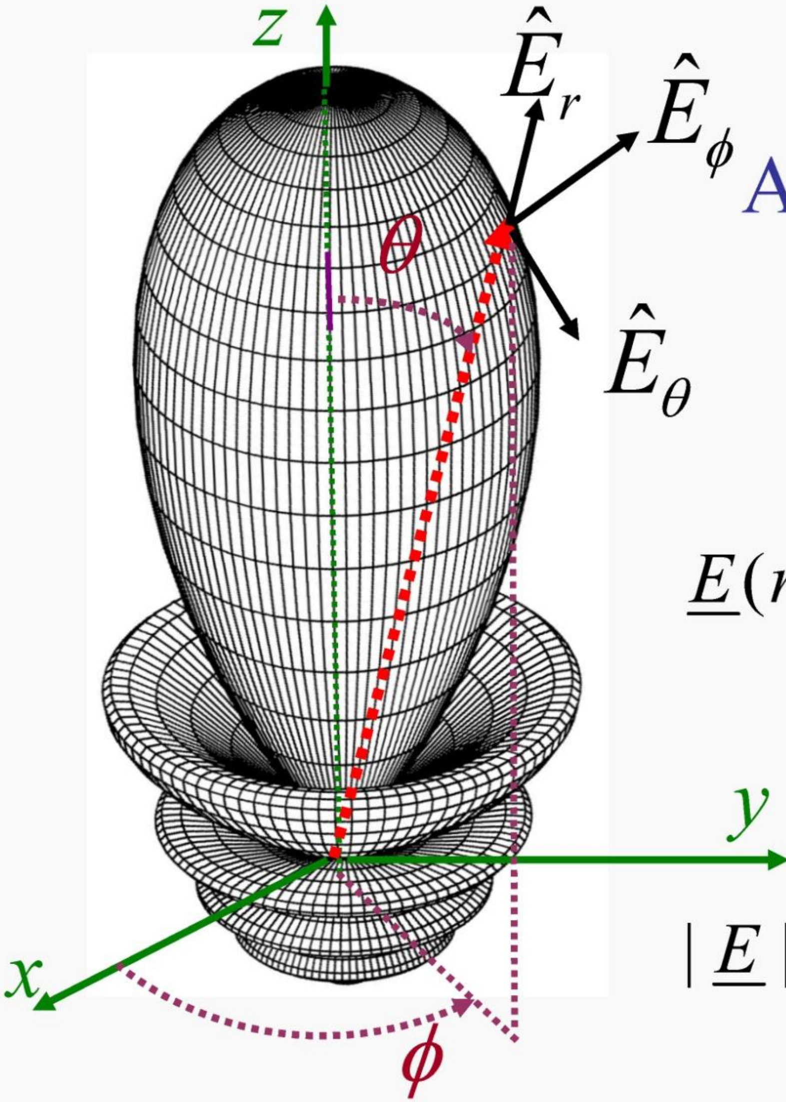


Fig. 2.3b

Normalized 3-D Amplitude *Field* Pattern of Linear Array



Linear Scale

$N = 10,\ d = \lambda/4$

$$\underline{E}(r,\theta,\phi)\,|_{r=r_c} = \hat{a}_r E_r(r_c,\theta,\phi) + \hat{a}_\theta E_\theta(r_c,\theta,\phi) + \hat{a}_\phi E_\phi(r_c,\theta,\phi)$$

$$|\,\underline{E}\,| = \sqrt{|\,E_r\,|^2 + |\,E_\theta\,|^2 + |\,E_\phi\,|^2}$$

Fig. 2.4

Amplitude Radiation Pattern

1. Isotropic,
   Directional,
   Omnidirectional
2. Principal patterns
3. Pattern lobes

Directional Pattern of a Horn

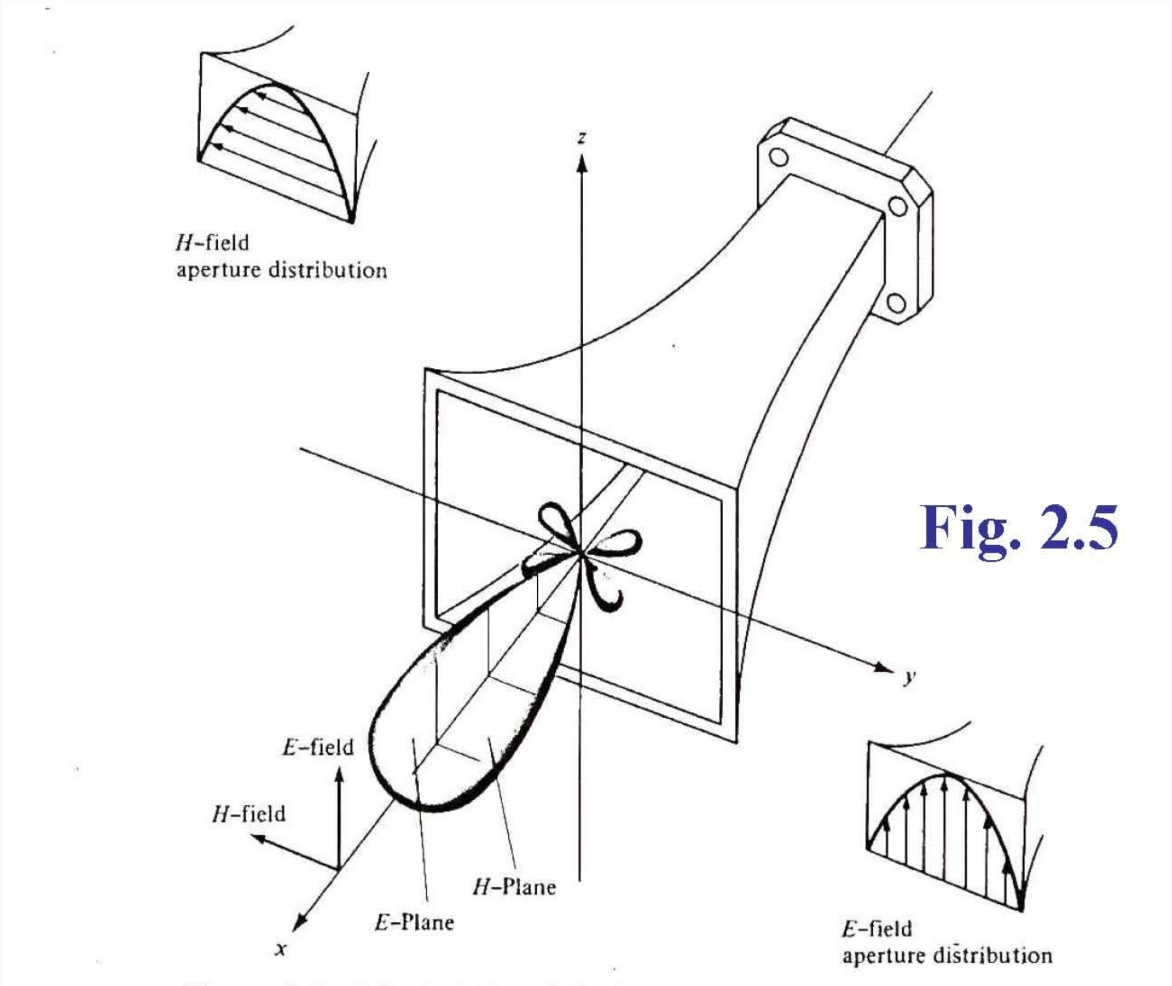
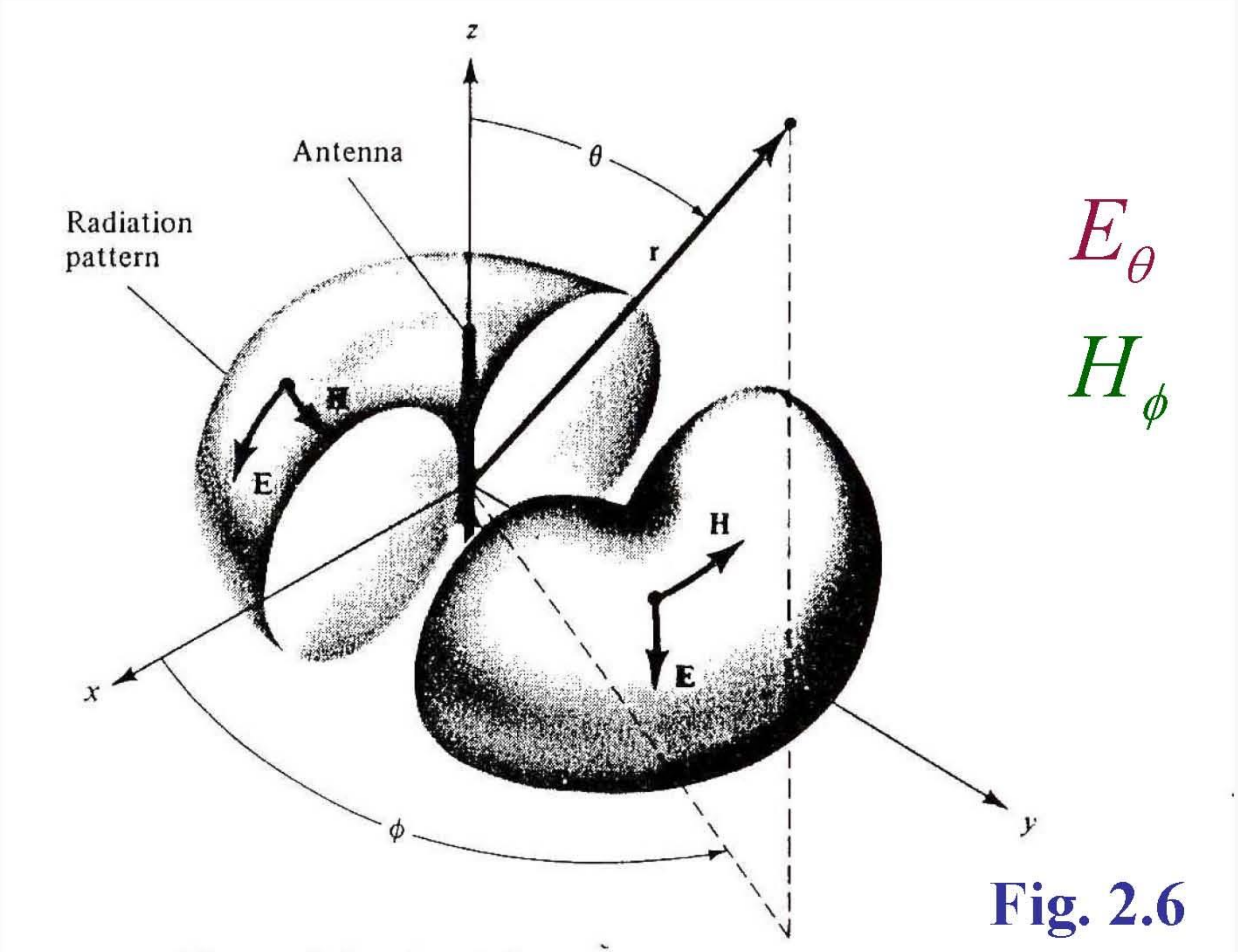


Fig. 2.5

Omnidirectional Pattern



E_θ

H_ϕ

Fig. 2.6

Radian

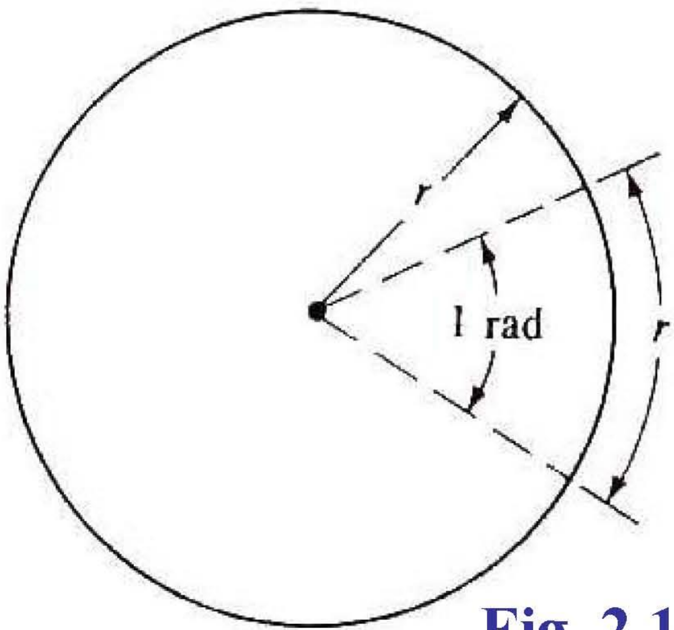


Fig. 2.10a

$$C = 2\pi r$$

$$\text{Rads} = \frac{C}{r} = \frac{2\pi r}{r}$$

$$\text{Rads} = 2\pi$$

Steradian

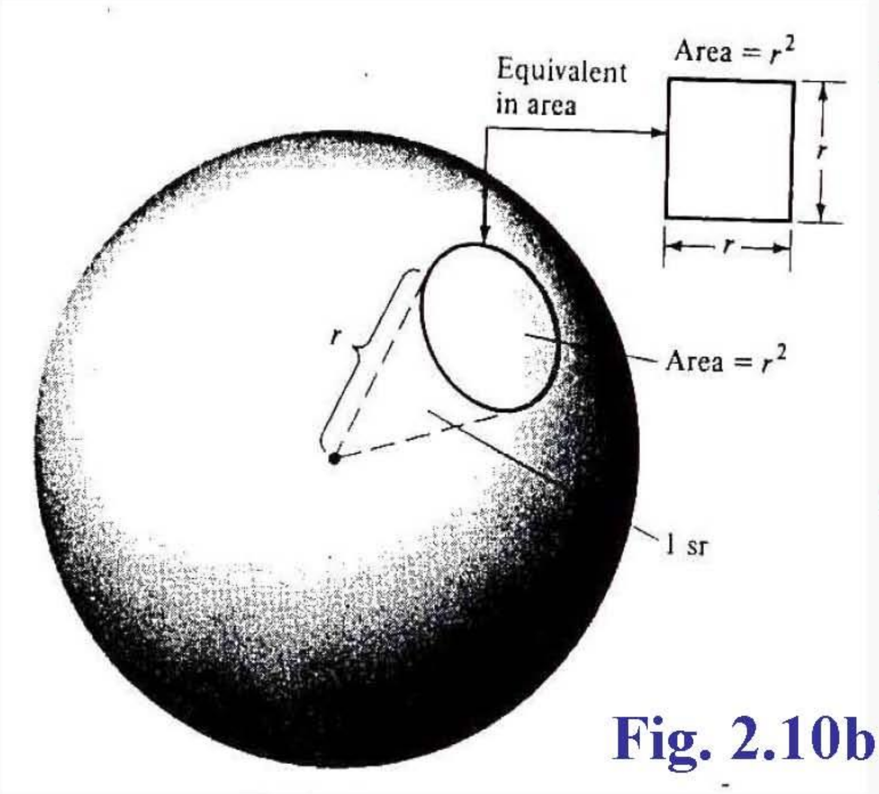


Fig. 2.10b

$$d\Omega = \frac{dA}{r^2} \qquad (2\text{-}1)$$

$$= \frac{r^2 \sin\theta d\theta d\phi}{r^2}$$

$$d\Omega = \sin\theta d\theta d\phi \qquad (2\text{-}2)$$

$$0 \le \theta \le \pi$$

$$0 \le \phi \le 2\pi$$

1. Radiation Power Density
2. Radiation Intensity

Spatial Variations in Power Density

$$\left[\underline{\mathcal{W}}(x,y,z;t)\right]_{ave} = \underline{W}(x,y,z)\Big|_{ave} = \frac{1}{2}\mathrm{Re}\left[\underline{E}\times\underline{H}^{*}\right]$$

$$\underline{W}\Big|_{ave} = \frac{1}{2}\mathrm{Re}\left[\underline{E}\times\underline{H}^{*}\right] \qquad (2\text{-}8)$$

It is assumed that both $\underline{E}$ and $\underline{H}$ represent peak amplitude values (not RMS: Root Mean Square). If $\underline{E}$ and $\underline{H}$ were to represent RMS values, then the one-half (1/2) must be omitted. Measuring instruments typically measure RMS values. Equation (2-8) is analogous to Ohm's Law in circuits

$$P = \frac{1}{2}VI^{*}$$

where V and I present peak (not RMS) values.

Intensity U (Far-field)

$$U = r^2\, W_{rad} = r^2 W_{ave} \qquad \text{(2-12)}$$

$$d\Omega = \frac{dA}{r^2} = \frac{r^2 \sin\theta\, d\theta\, d\phi}{r^2}$$

$$d\Omega = \sin\theta\, d\theta\, d\phi$$

$$W = \text{Power Density}$$

$$= P / A\left(\frac{W}{m^2}\right)$$

$$U = \text{Radiation Intensity}$$

$$= P / \Omega\left(\frac{W}{Sr}\right)$$

Isotropic Source

$$W_o = \frac{P}{4\pi r^2} \left(W/m^2 \right)$$

$$U_o = \frac{P}{4\pi} \left(W/Sr \right)$$

$$U = \frac{Power}{Unit\ Solid\ Angle} = \frac{Power}{Unit\ Area / r^2}$$

because

$$d\Omega = \frac{dA}{r^2} = \frac{r^2 \sin\theta \, d\theta \, d\phi}{r^2} = \sin\theta \, d\theta \, d\phi$$

$$U = r^2 \frac{Power}{Unit\ Area} = r^2 W_{ave} = r^2 W_{rad}$$

$$U = r^2 W_{rad} \Rightarrow W_{rad} = \frac{U}{r^2}$$

$$P_{rad} = \int_0^{2\pi}\int_0^{\pi} \hat{a}_r W_{rad} \cdot \hat{a}_r r^2 \sin\theta \, d\theta \, d\phi$$

$$= \int_0^{2\pi}\int_0^{\pi} W_{rad} r^2 \sin\theta \, d\theta \, d\phi$$

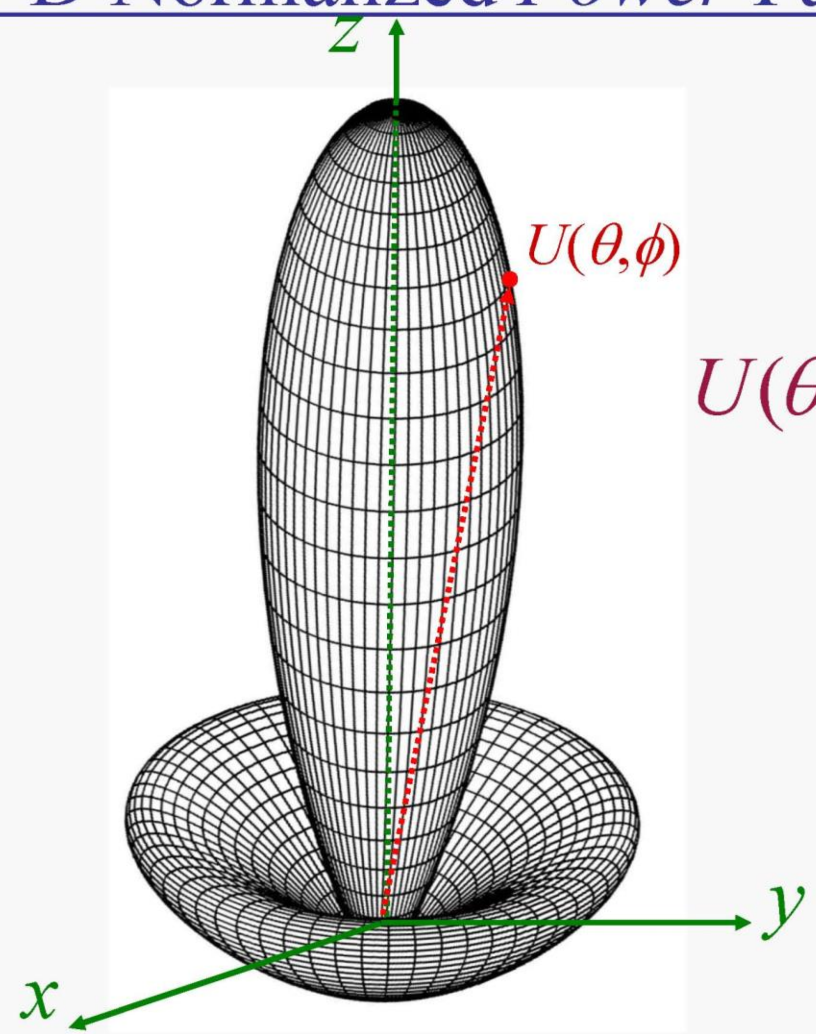
<u>Since</u> $W_{rad} = \dfrac{U}{r^2}$

$$P_{rad} = P_{ave} = \int_0^{2\pi}\int_0^{\pi} U \underbrace{\sin\theta \, d\theta \, d\phi}_{d\Omega}$$

Beamwidths

- Half-Power (*HPBW*)
- First Null (*FNBW*)

3-D Normalized *Power* Pattern of *U*

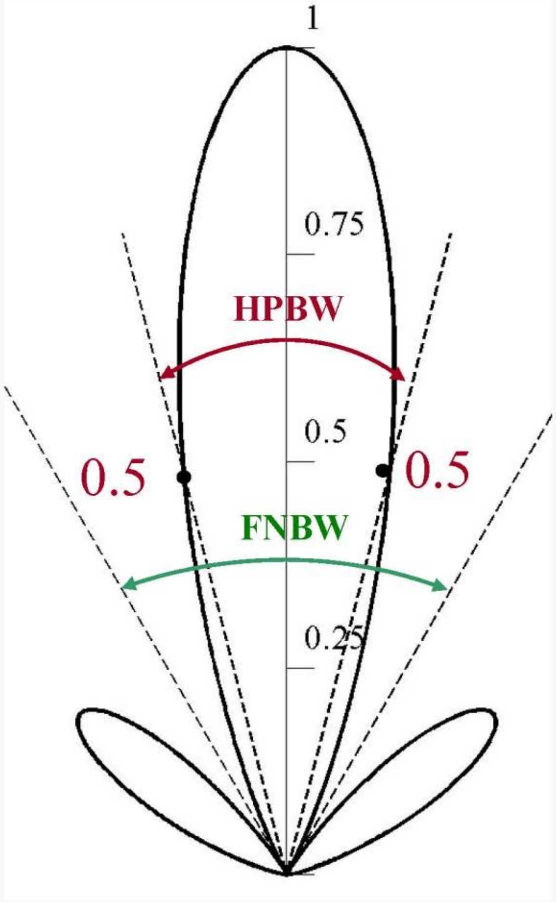


Linear Scale

$$U(\theta,\phi) = \cos^2\theta\cos^2 3\theta$$

Fig. 2.11(a)

HPBW and FNBW of Radiation Intensity U

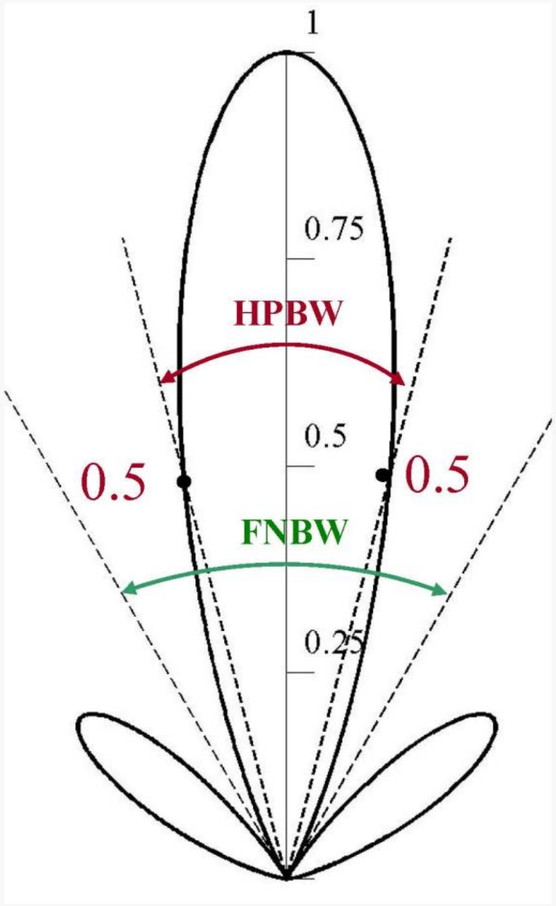


Linear Scale

$$U(\theta,\phi) = \cos^2\theta \cos^2 3\theta$$

Fig. 2.11(b)

HPBW and FNBW of Radiation Intensity U



Linear Scale

$$U(\theta,\phi) = \cos^2\theta \cos^2 3\theta$$

HPBW $= 28.65^o$

FNBW $= 60^o$

Fig. 2.11(b)

Directivity D

$$D = \frac{U(\theta,\phi)}{U_o} = \frac{4\pi U(\theta,\phi)}{P_{rad}} \quad (2\text{-}16)$$

$$D_{\max} = D_o = \frac{U_{\max}}{U_o} = \frac{4\pi U_{\max}}{P_{rad}} \quad (2\text{-}16a)$$

$$\boxed{D(dB) = 10\log_{10}[\,D(\text{dimensionless})\,]}$$

Directivity : The measure of how much power , power density or power Intensity is concentrated in a certain beam

$$D = U_{max} / U_o$$

Where:

U_o is the average power intensity and

U_{max} is maximum intensity

When $U_{max} = U_o$, the antenna is omindirectional &

D = 1 = 0 dB .

The directivity is usually inversely proportional with the half power beam width

D [illegible](1 / HPBW)

Antenna Efficiency

Antenna Reference Terminals

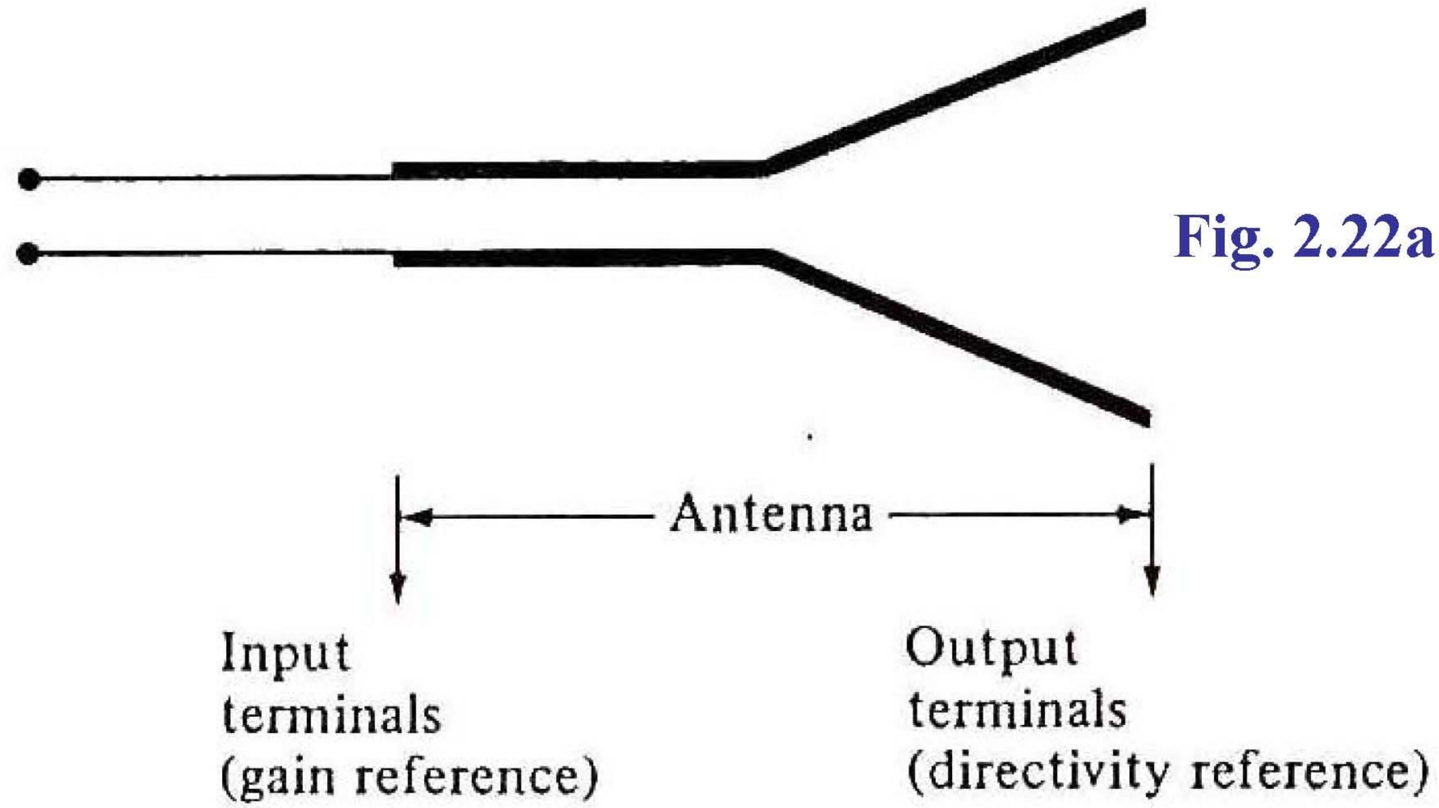


Fig. 2.22a

Reflection, Conduction, And Dielectric Losses

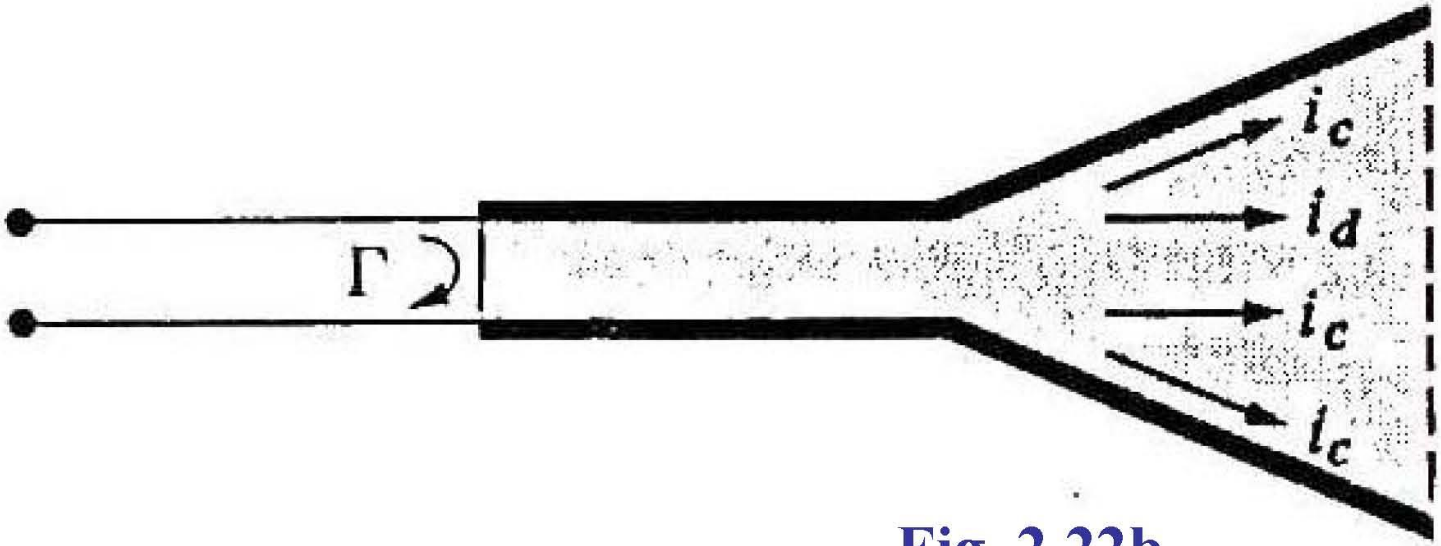


Fig. 2.22b

Antenna Efficiency e_o

$$e_o = e_r e_c e_d = e_r e_{cd} \qquad (2\text{-}44)$$

$$e_o = (1 - |\Gamma_{in}|^2) e_{cd} \qquad (2\text{-}45)$$

$$e_o = Total\ efficiency$$

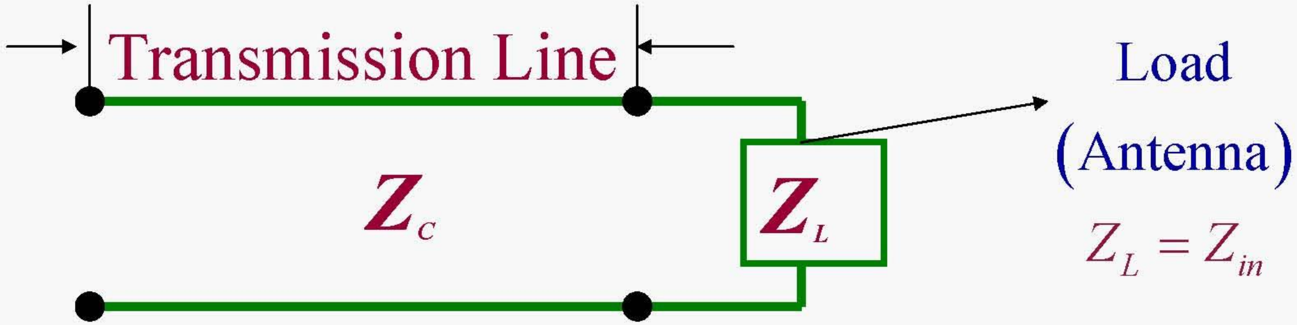
$$e_r = Reflection\ efficiency$$

$$e_{cd} = Radiation\ efficiency$$

Transmission Line and Load

Z_c = Characteristic Impedance of Line

Z_L = Load Impedance



$$\Gamma_{in} = \frac{Z_L - Z_c}{Z_L + Z_c} = \frac{Z_{in} - Z_c}{Z_{in} + Z_c}$$

$$e_r = (1 - |\Gamma_{in}|^2)$$

$$\Gamma_{in} = \frac{Z_{in} - Z_c}{Z_{in} + Z_c}$$

$$VSWR = \frac{1 + |\Gamma_{in}|}{1 - |\Gamma_{in}|}$$

$$|\Gamma_{in}| = \frac{VSWR - 1}{VSWR + 1}$$

Gain

$$P_{rad} = e_c e_d \, P_{in}$$

$$P_{rad} = e_{cd} \, P_{in}$$

$$\text{Gain} = G = 4\pi \frac{\text{Radiation intensity}}{\text{Total input (accepted) power}}$$

$$G = 4\pi \frac{U(\theta,\phi)}{P_{in}} \tag{2-46}$$

$$P_{rad} = e_{cd} P_{in} \Rightarrow P_{in} = \frac{P_{rad}}{e_{cd}} \tag{2-47}$$

$$G = 4\pi \frac{U(\theta,\phi)}{P_{rad}/e_{cd}} = e_{cd} \underbrace{\left[4\pi \frac{U(\theta,\phi)}{P_{rad}}\right]}_{D} \tag{2-48}$$

$$G = e_{cd} D$$

Absolute Gain G_{abs}

$$G_{abs}(\theta,\phi) = e_o D(\theta,\phi) = e_r e_{cd} D(\theta,\phi)$$

$$= (1 - |\Gamma_{in}|^2) e_{cd} D(\theta,\phi) \quad \text{(2-49b)}$$

$$e_o = antenna\ total\ efficiency$$

$$e_r = (1 - |\Gamma_{in}|^2) = Reflection\ efficiency$$

$$e_{cd} = e_c e_d = Radiation\ efficiency$$

$$e_r = Conduction\ efficiency$$

$$e_d = Dielectric\ efficiency$$

Gain : The directivity after considering the antennas efficiency .

G = D * [illegible]

Usually measured in dB .

Effective Aperture

So far, the properties of antennas have been described in terms of their radiation characteristics. A receiving antenna has directional properties also described by the radiation pattern, but in this case it refers to the ratio of received power normalized to the maximum value.

An important concept used to describe the reception properties of an antenna is that of *effective aperture*. Consider a TEM wave of a given power density Ψ at the receiving antenna. Let the load at the *antenna terminals* be a complex conjugate match so that maximum power transfer occurs and power P_{rec} is delivered to the load. Note that the power delivered to the actual receiver may be less than this as a result of feeder losses. With the receiving antenna aligned for maximum reception (including polarization alignment, which is described in detail later), the received power will be proportional to the power density of the incoming wave. The constant of proportionality is the effective aperture A_{eff} which is defined by the equation

ψ = W (W/m2)

$$P_{rec} = A_{eff}\,\Psi \tag{6.13}$$

For antennas which have easily identified physical apertures, such as horns and parabolic reflector types, the effective aperture is related in a direct way to the physical aperture. If the wave could uniformly illuminate the physical aperture, then this would be equal to the effective aperture. However, the presence of the antenna in the field of the incoming wave alters the field distribution, thereby preventing uniform illumination. The effective aperture is smaller than the physical aperture by a factor known as the *illumination efficiency*. Denoting the illumination efficiency by η_I gives

$$A_{eff} = \eta_I A_{physical} \tag{6.14}$$

The illumination efficiency is usually a specified number, and it can range between about 0.5 and 0.8. Of course, it cannot exceed unity, and a conservative value often used in calculations is 0.55.

A fundamental relationship exists between the power gain of an antenna and its effective aperture. This is

$$\frac{A_{eff}}{G} = \frac{\lambda^2}{4\pi} \tag{6.15}$$

where λ is the wavelength of the TEM wave, assumed sinusoidal (for practical purposes, this will be the wavelength of the radio wave carrier). The importance of this equation is that the gain is normally the known (measurable) quantity, but once this is known, the effective aperture is also known.

6.12 Horn Antennas

The horn antenna is an example of an aperture antenna which provides a smooth transition from a waveguide to a larger aperture that couples more effectively into space. Horn antennas are used directly as radiators aboard satellites to illuminate comparatively large areas of the earth, and they are also widely used as primary feeds for reflector-type antennas both in transmitting and receiving modes. The three most commonly used types of horns are illustrated in Fig. 6.10.

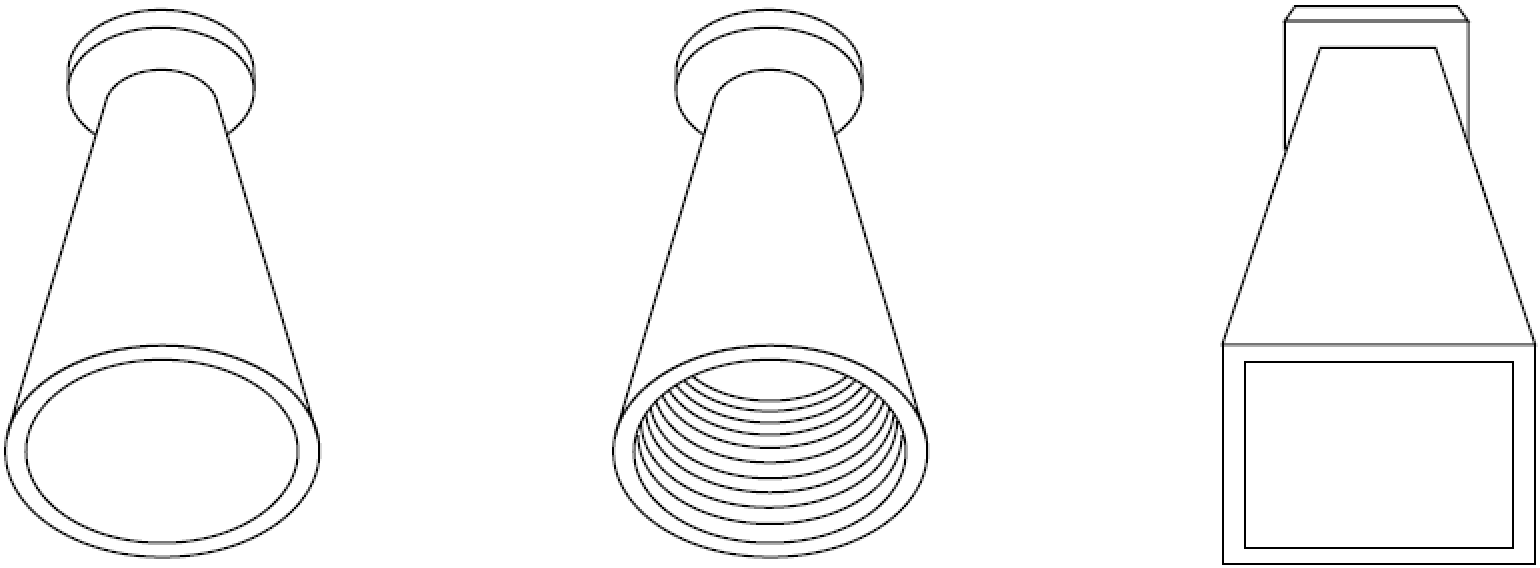


Figure 6.10 Horn antennas: (a) smooth-walled conical, (b) corrugated, and (c) pyramidal.

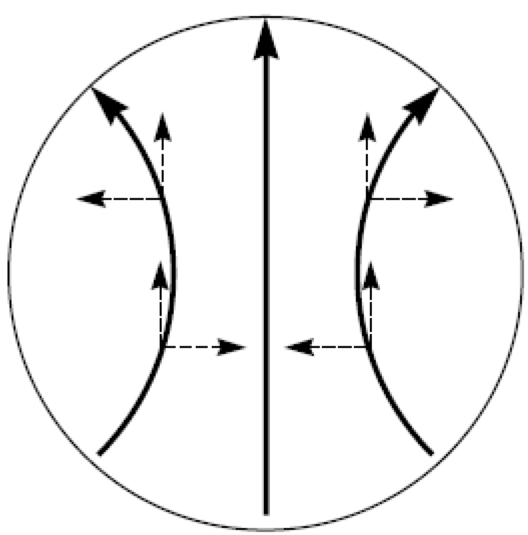
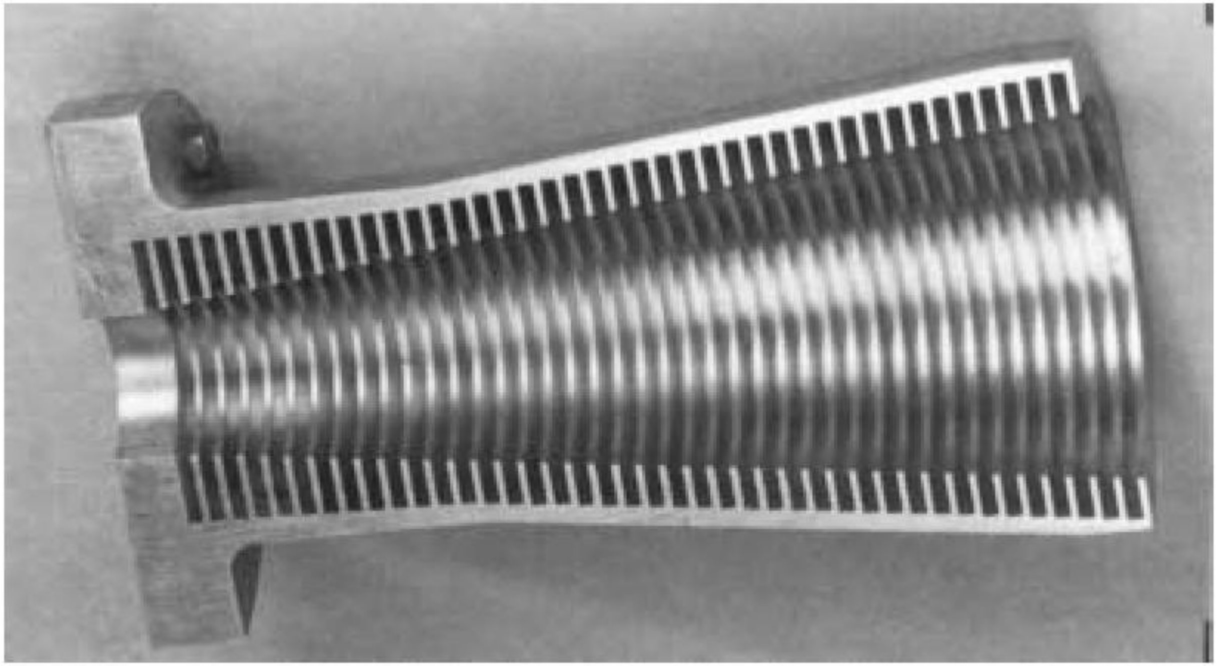
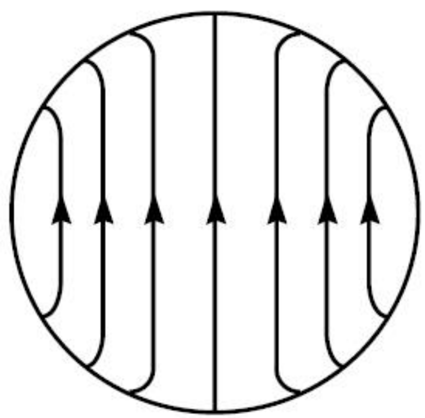


Figure 6.11 Aperture field in a smooth-walled conical horn.



(a)



(b)

Figure 6.12 (*a*) Cross section of a corrugated horn. (*From Alver, 1992, with permission.*) (*b*) Aperture field.

Reflector antenna

The first cassegrain
Reflector was designed
By Laurent cassegrain
In 1672 .

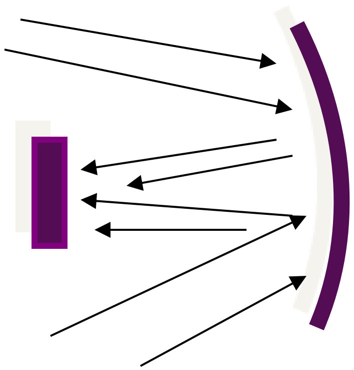
6.13 The Parabolic Reflector

Parabolic reflectors are widely used in satellite communications systems to enhance the gain of antennas. The reflector provides a focusing mechanism which concentrates the energy in a given direction. The most commonly used form of parabolic reflector has a circular aperture, as shown in Fig. 6.15. This is the type seen in many home installations for the reception of TV signals. The circular aperture configuration is referred to as a *paraboloidal reflector*.

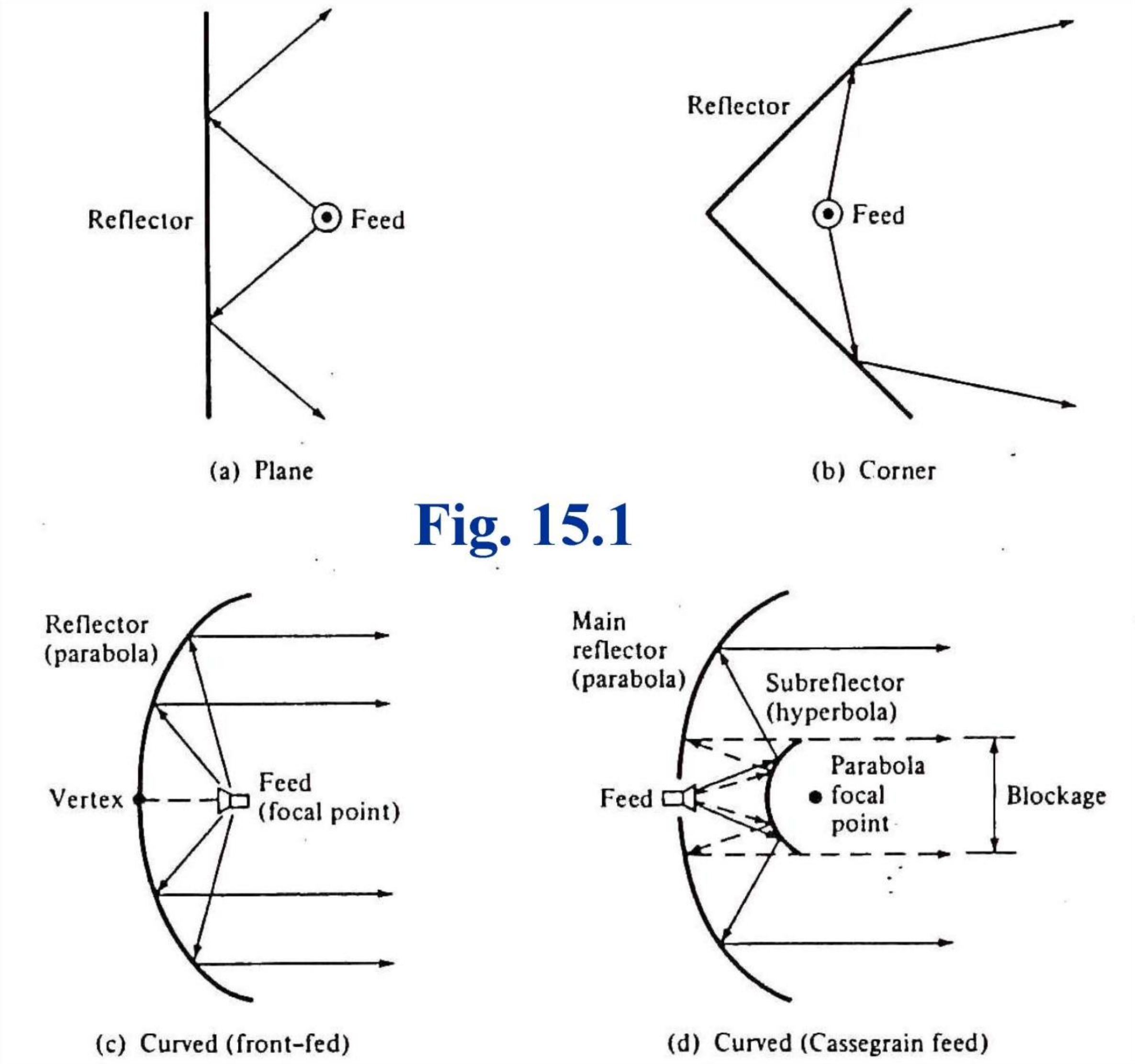
The main property of the paraboloidal reflector is its focusing property, normally associated with light, where parallel rays striking the reflector converge on a single point known as the *focus* and, conversely, rays originating at the focus are reflected as a parallel beam of light.

5.1. Why Reflectors ?

While using aperture antennas we always need to increase the aperture
Area to increase its directivity , but as this is not practical ,
instead we place a reflecting surface face to face with the aperture
(or any other antenna) , the reflecting surface collimates radiation to
The small aperture and thus we satisfied high directivity with a small
Aperture , and overcame space limitations.



A side view of
An aperture of
A small area
And a reflecting
Surface used.



Reflector
Feed
(a) Plane
Reflector
Feed
(b) Corner
Reflector
(parabola)
Vertex
Feed
(focal point)
(c) Curved (front-fed)
Main
reflector
(parabola)
Subreflector
(hyperbola)
Feed
Parabola
focal
point
Blockage
(d) Curved (Cassegrain feed)

Fig. 15.1

Paraboloid

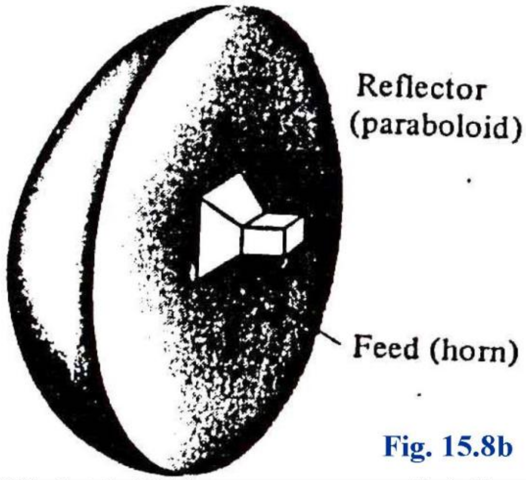
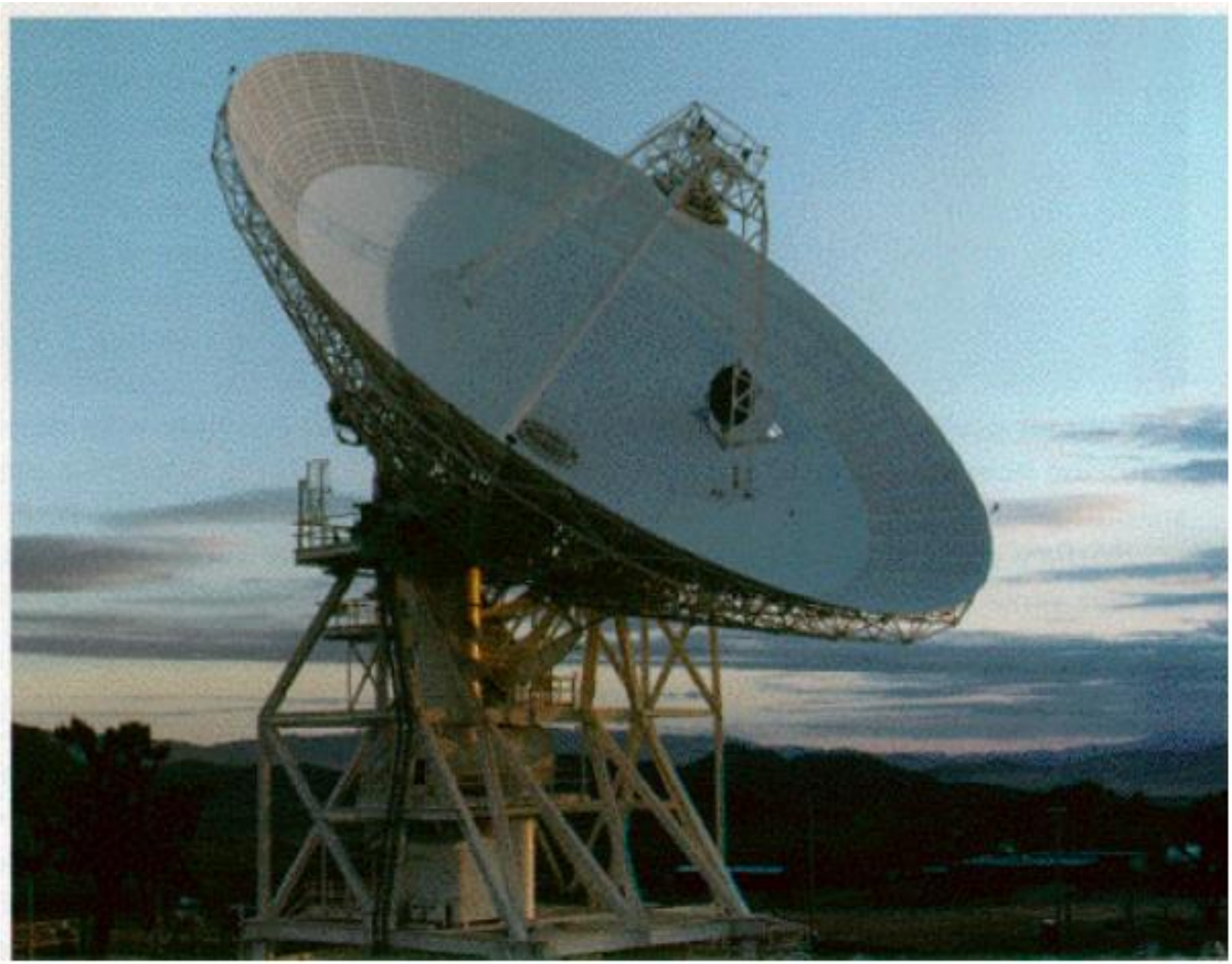


Fig. 15.8b

Figure 6.15 A parabolic reflector. (*Courtesy of Scientific Atlanta, Inc.*)

10 meter Earth Station Dual Reflector



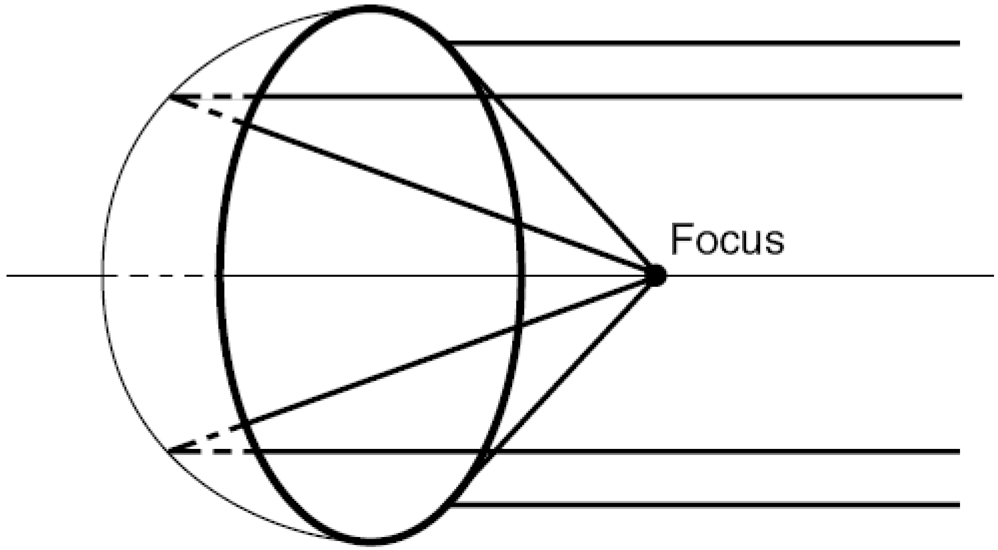
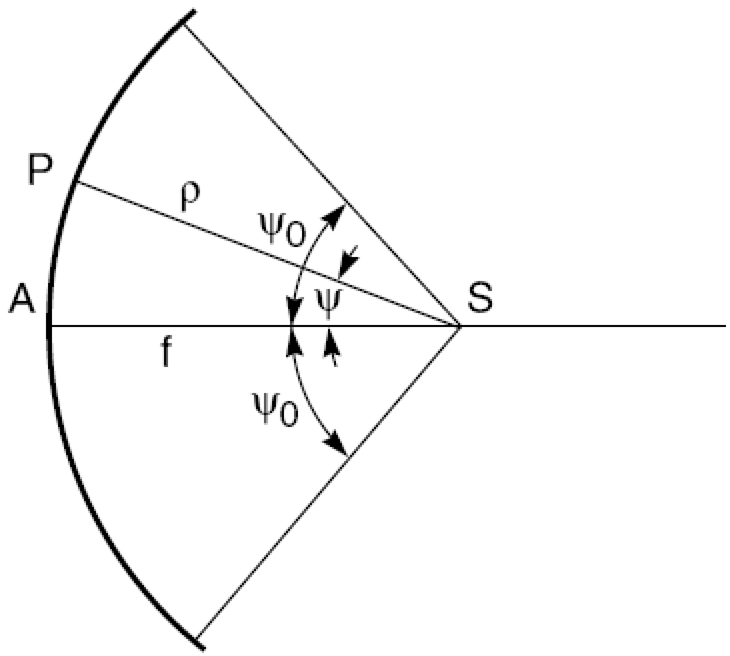
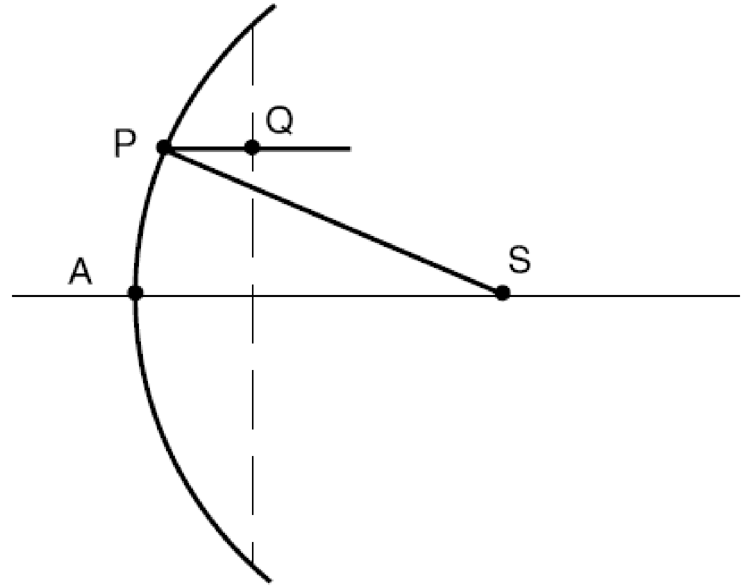


Figure 6.16 The focusing property of a paraboloidal reflector.

This is illustrated in Fig. 6.16. Light, of course, is a particular example of an electromagnetic wave, and the same properties apply to electromagnetic waves in general, including the radio waves used in satellite communications. The ray paths from the focus to the aperture plane (the plane containing the circular aperture) are all equal in length.



(b)



(a)

= SA and a ray path SPQ. (b) The focal distance ρ.

The geometric properties of the paraboloidal reflector of interest here are most easily demonstrated by means of the *parabola*, which is the curve traced by the reflector on any plane normal to the aperture plane and containing the focus. This is shown in Fig. 6.17*a*. The *focal point* or *focus* is shown as S, the *vertex* as A, and the axis is the line passing through S and A. SP is the *focal distance* for any point P and SA the *focal length*, usually denoted by f. (The parabola is examined in more detail in App. B). A ray path is shown as SPQ, where P is a point on the curve and Q is a point in the aperture plane. Length PQ lies parallel to the axis. For any point P, all path lengths SPQ are equal; that is, the distance $SP + PQ$ is a constant which applies for all such paths. The path equality means that a wave originating from an isotropic point source has a uniform phase distribution over the aperture plane. This property, along with the parallel-beam property, means that the wavefront is plane. Radiation from the paraboloidal reflector appears to originate as a plane wave from the plane normal to the axis and containing the directrix (see App. B). Although the characteristics of the reflector antenna are more readily described in terms of radiation, it should be kept in mind that the reciprocity theorem makes these applicable to the receiving mode as well.

Now although there are near- and far-field components present in the reflector region, the radio link is made through the far-field component, and only this need be considered. For this, the reflected wave is a plane wave, while the wave originating from the isotropic source and striking the reflector has a spherical wavefront. The power density in the plane wave is independent of distance. For the spherical wave, the power density of the far-field component decreases in inverse proportion to the distance squared, and therefore, the illumination at the edge of the reflector will be less than that at the vertex.

This gives rise to a nonuniform amplitude distribution across the aperture plane, which in effect means that the illumination efficiency is reduced. Denoting the focal distance by ρ and the focal length by f as in Fig. 6.17*b*, then, as shown in App. B,

$$\frac{\rho}{f} = \sec^2 \frac{\Psi}{2} \tag{6.27}$$

The *space attenuation function* (SAF) is the ratio of the power reaching point P to that reaching point A, and since the power density is inversely proportional to the square of the distance, the ratio is given by

$$SAF = \left(\frac{f}{\rho}\right)^2 = \cos^4 \frac{\Psi}{2} \tag{6.28}$$

For satellite applications, a high illumination efficiency is desirable. This requires that the radiation pattern of the primary antenna, which is situated at the focus and which illuminates the reflector, should approximate as closely as practical the inverse of the space attenuation factor.

An important ratio is that of aperture diameter to focal length. Denoting the diameter by D (do not confuse with the $\mathscr{D}$ used for directivity), then, as shown in App. B,

$$\boxed{\frac{f}{D} = 0.25 \cot \frac{\Psi_0}{2}} \tag{6.29}$$

The position of the focus in relation to the reflector for various values of *f/D* is shown in Fig. 6.18. For $f/D < 0.25$, the primary antenna lies in the space between the reflector and the aperture plane, and the illumination tapers away toward the edge of the reflector. For $f/D > 0.25$, the primary antenna lies outside the aperture plane, which results in more nearly uniform illumination, but *spillover* increases. In the transmitting mode, spillover is the radiation from the primary antenna which is directed toward the reflector but which lies outside the angle $2\Psi_0$. In satellite applications, the primary antenna is usually a horn (or an array of horns, as will be shown later) pointed toward the reflector. In order to compensate for the space attenuation described above, higher-order modes can be added to the horn feed so that the horn radiation pattern approximates the inverse of the space attenuation function (Chang, 1989).

The radiation from the horn will be a spherical wave, and the *phase center* will be the center of curvature of the wavefront. When used as the primary antenna for a parabolic reflector, the horn is positioned so that the phase center lies on the focus.

The focal length can be given in terms of the depth of the reflector and its diameter. It is sometimes useful to know the focal length for

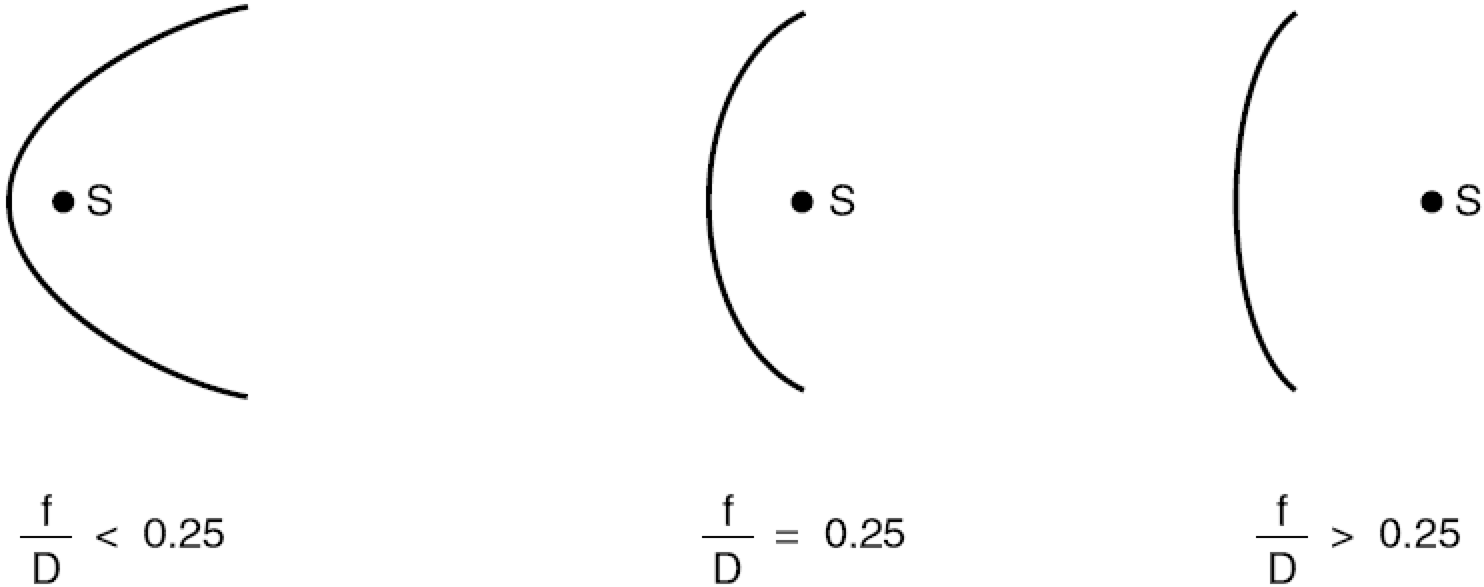


Figure 6.18 Position of the focus for various f/D values.

setting up a receiving system. The depth l is the perpendicular distance from the aperture plane to the vertex. This relationship is shown in App. B to be

$$f = \frac{D^2}{16l} \tag{6.30}$$

The gain and beamwidths of the paraboloidal antenna are as follows: The physical area of the aperture plane is

$$Area = \frac{\pi D^2}{4} \tag{6.31}$$

From the relationships given by Eqs. (6.14) and (6.15), the gain is

$$G = \frac{4\pi}{\lambda^2} \eta_I \text{ area}$$
$$= \eta_I \left(\frac{\pi D}{\lambda} \right)^2 \tag{6.32}$$

The radiation pattern for the paraboloidal reflector is similar to that developed in Example 6.1 for the rectangular aperture, in that there is a main lobe and a number of sidelobes, although there will be differences in detail. In practice, the sidelobes are accounted for by an envelope function as described in Chap. 13. Useful approximate formulas for the half-power beamwidth and the beamwidth between the first nulls (BWFN) are

$$\boxed{\mathrm{HPBW} \cong 70\ \frac{\lambda}{D}} \tag{6.33}$$

$$\mathrm{BWFN} \cong 2\mathrm{HPBW} \tag{6.34}$$

In these relationships, the beamwidths are given in degrees. The paraboloidal antenna described so far is *center-fed* in that the primary horn is pointed toward the center of the reflector. With this arrangement the primary horn and its supports present a partial blockage to the reflected wave. The energy scattered by the blockage is lost from the main lobe, and it can create additional sidelobes. One solution is to use an *offset feed* as described in the next section.

6.14 The Offset Feed

Figure 6.21*a* shows a paraboloidal reflector with a horn feed at the focus. In this instance the radiation pattern of the horn is offset so that

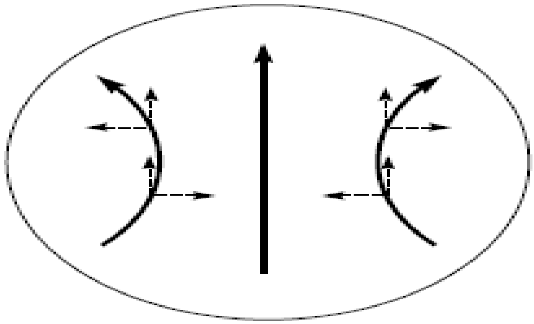


Figure 6.19 Current paths in a paraboloidal reflector for linear polarization.

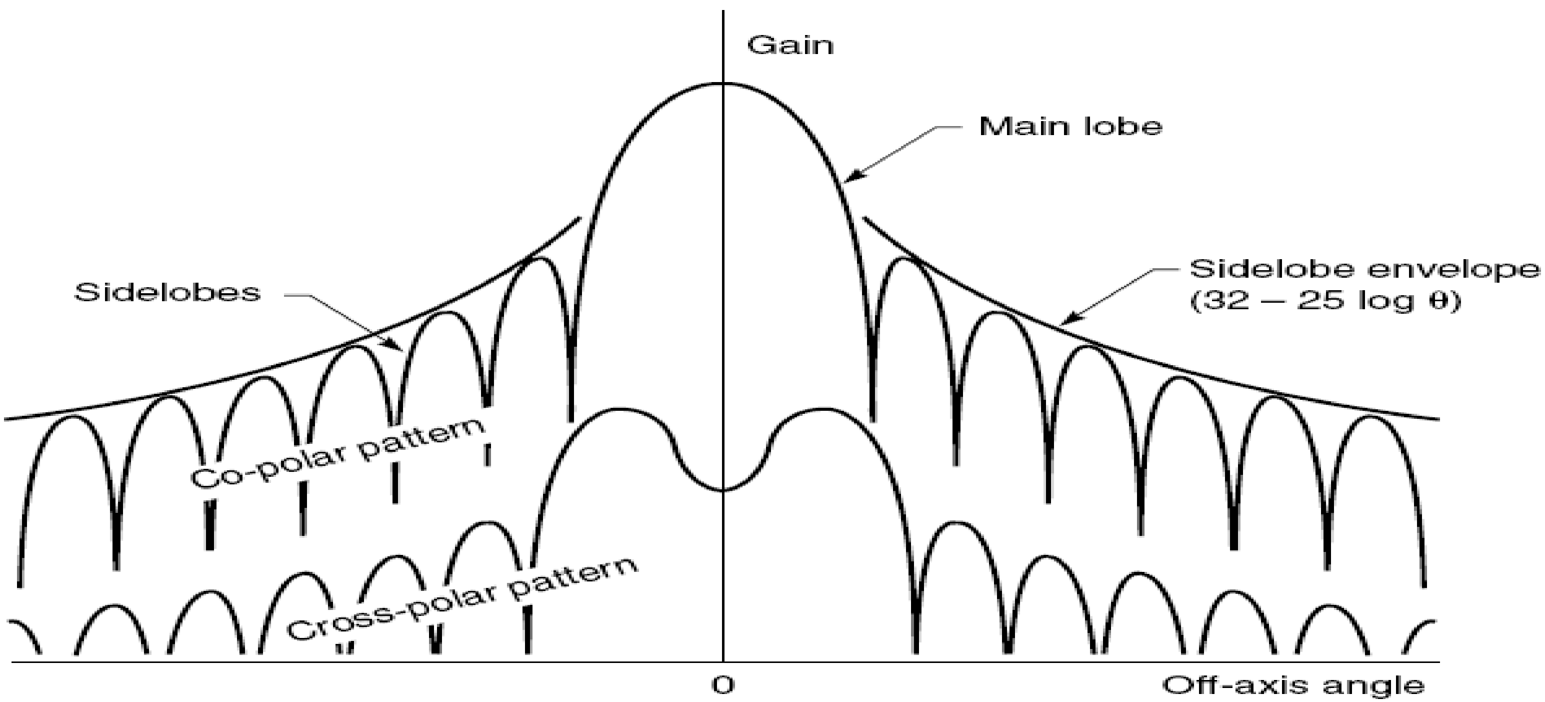
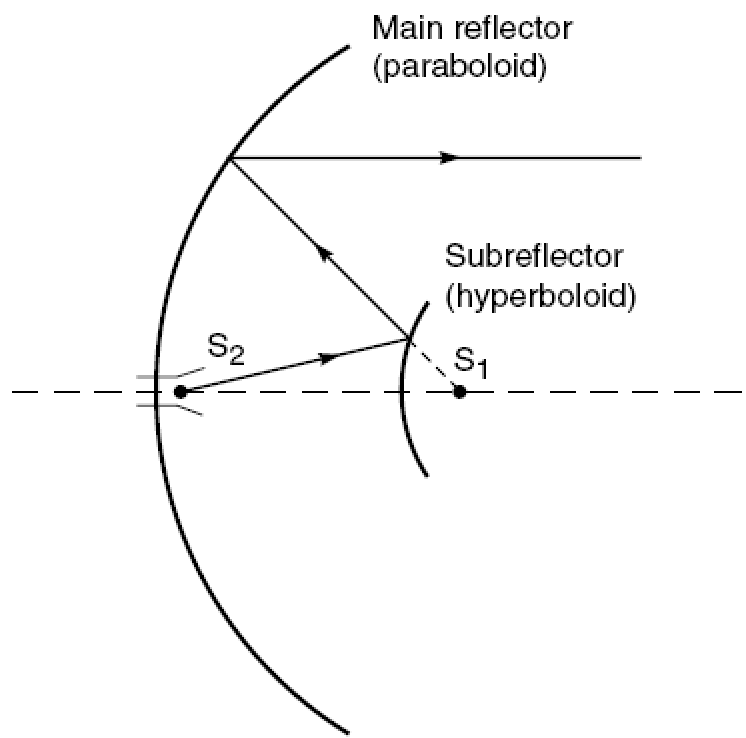


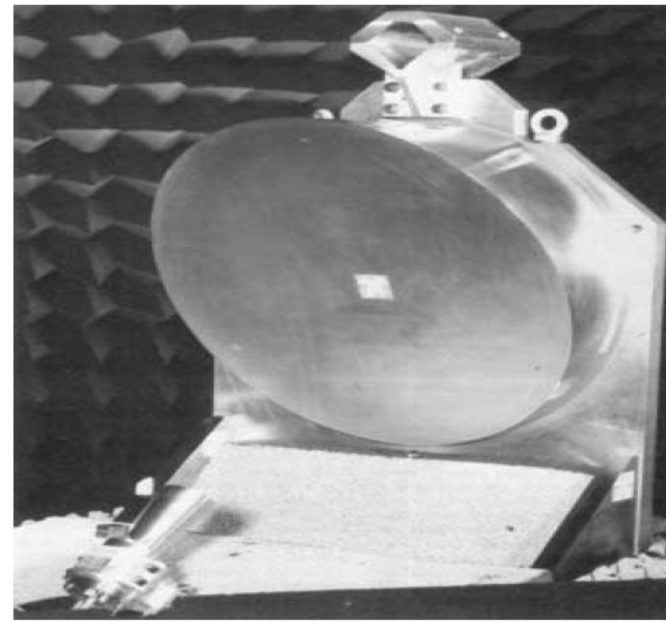
Figure 6.20 Copolar and cross-polar radiation patterns. (*From FCC Report FCC/OST R83-2, 1983.*)

Cassegrain antenna. The basic Cassegrain form consists of a main paraboloid and a subreflector, which is a hyperboloid (see App. B). The subreflector has two focal points, one of which is made to coincide with that of the main reflector and the other with the phase center of the



(a)

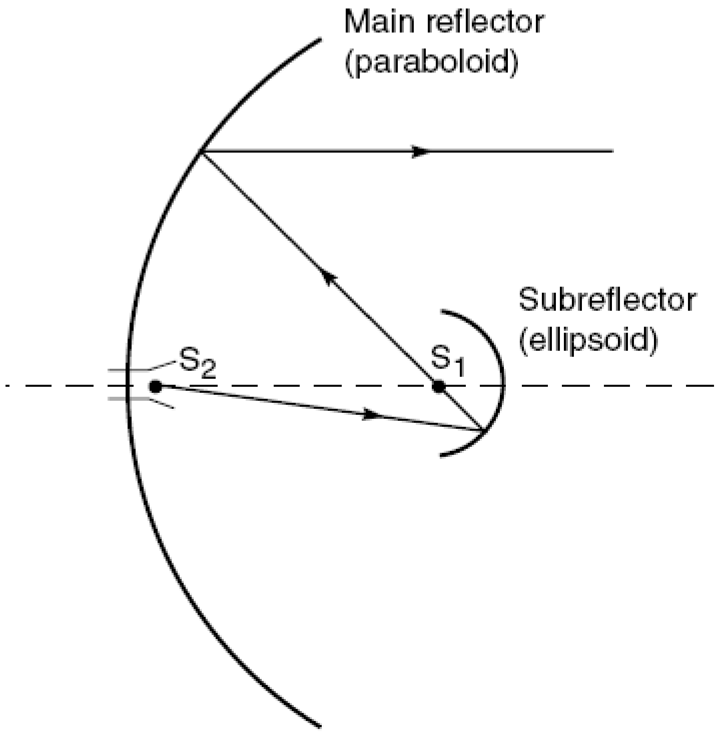
Figure 6.22 A 19-m Cassegrain antenna. (*Courtesy of TIW Systems, Inc.*)



(b)

Figure 6.21 (*a*) Ray paths for an offset reflector. (*b*) The offset feed for a paraboloidal reflector. (*From Brain and Rudge, 1984, with permission.*)

Gregorian antenna. The basic Gregorian form consists of a main paraboloid and a subreflector, which is an ellipsoid (see App. B). As



(b)

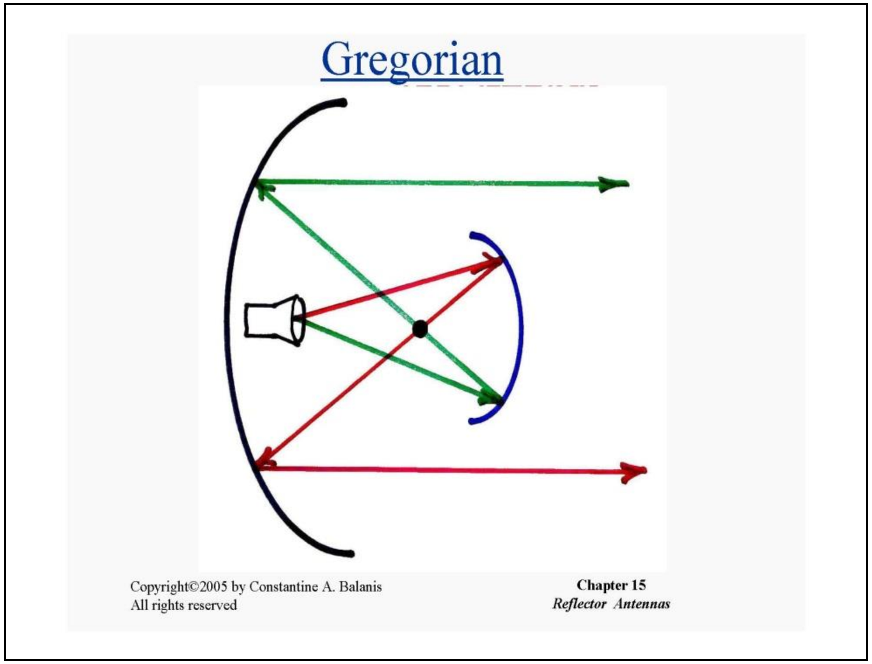


Figure 6.23 Ray paths for (*a*) Cassegrain and (*b*) Gregorian antennas.

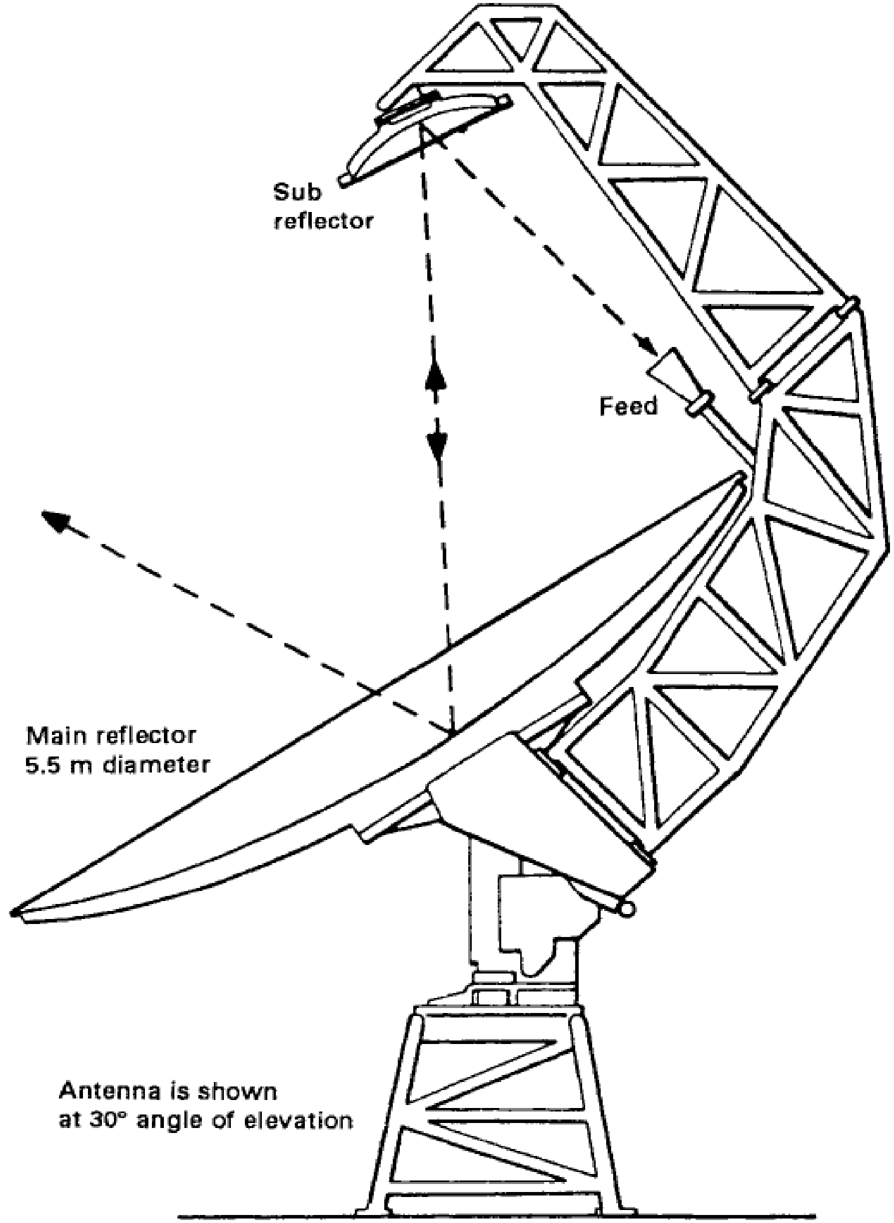


Figure 6.24 Offset Gregorian antenna *From Radio Electr. Eng., vol. 54, No. 3, Mar. 1984, p. 112, with permission.)*

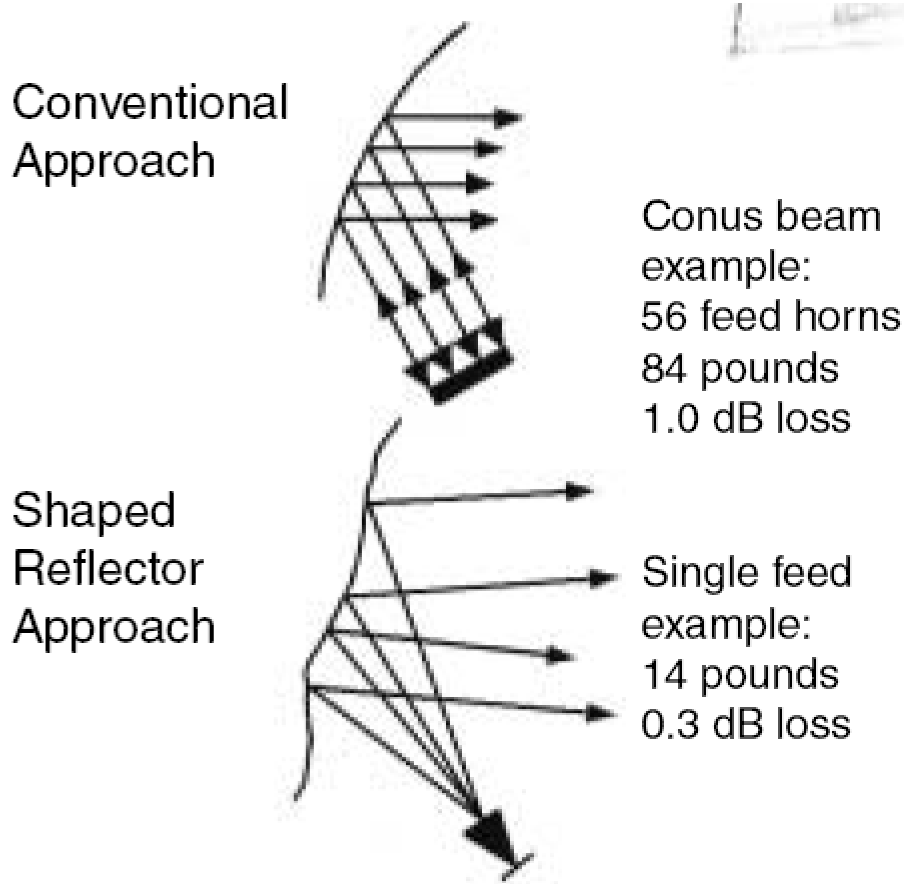


Figure 6.25 Shaped-beam reflector, showing ray paths. (*Courtesy of Hughes Space and Communications Company. Reproduced from Vectors XXXV(3):14, 1993. © Hughes Aircraft Co.*)

Input And Output Bundle

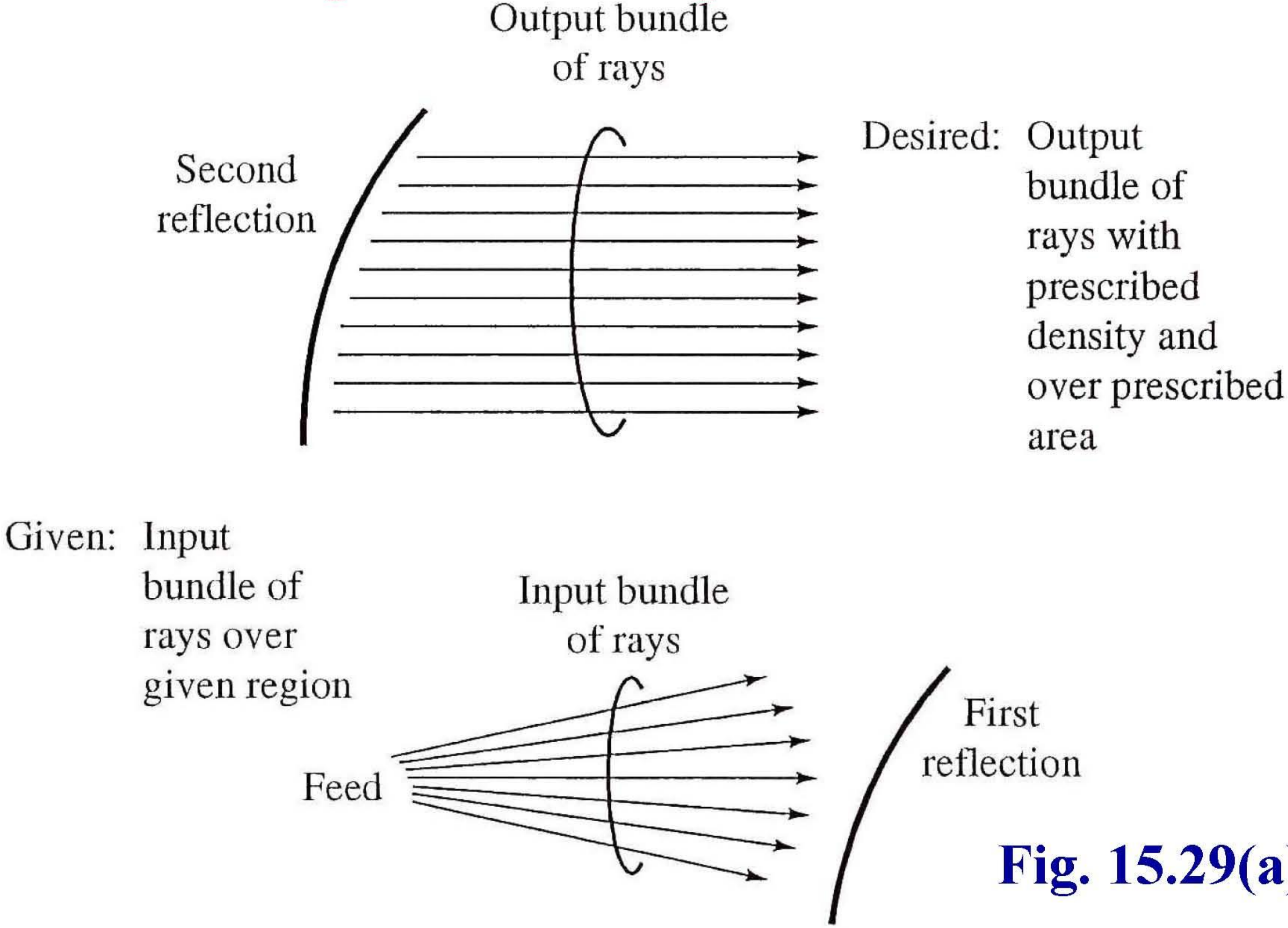


Fig. 15.29(a)

Specified Feed And Prescribed Output

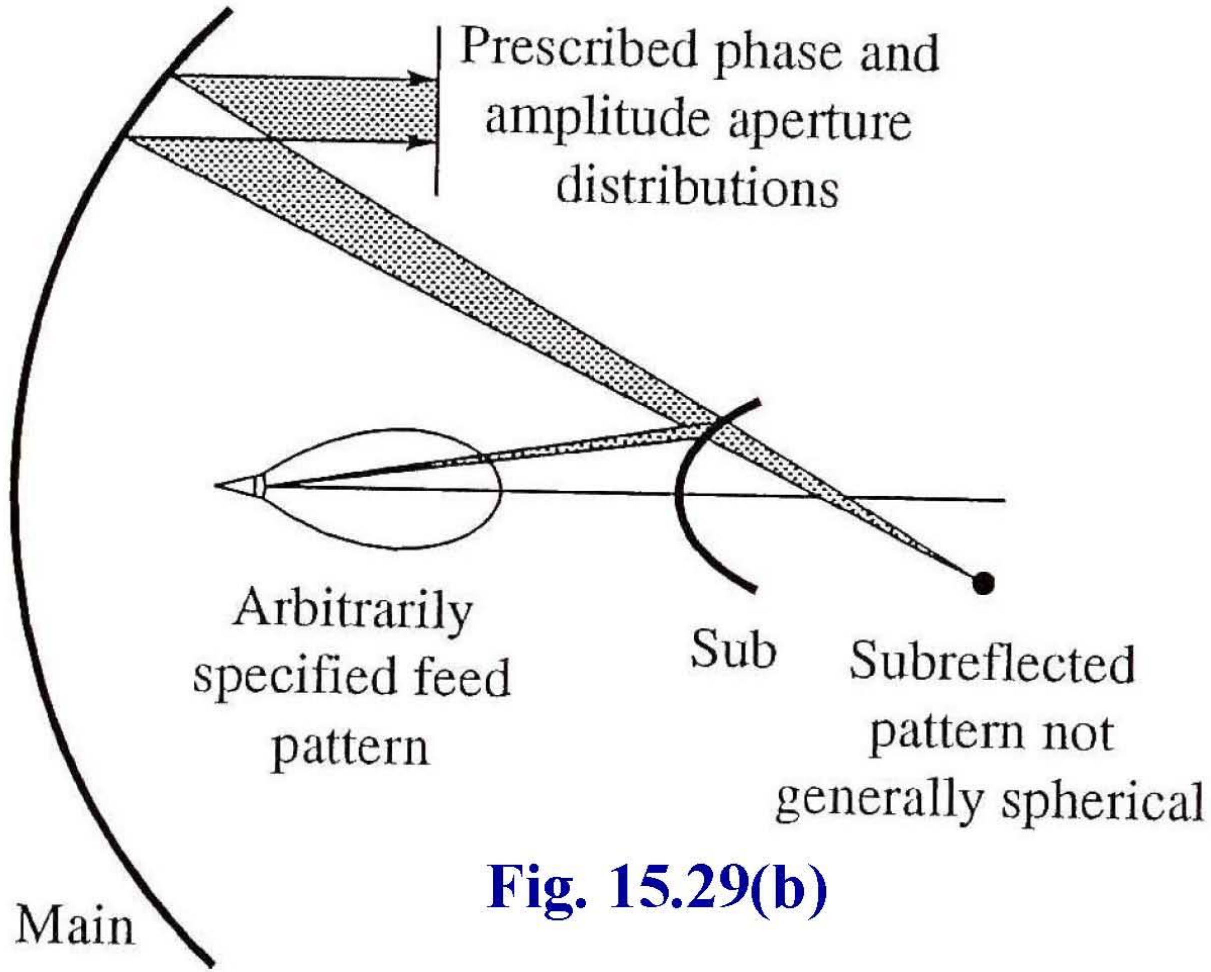
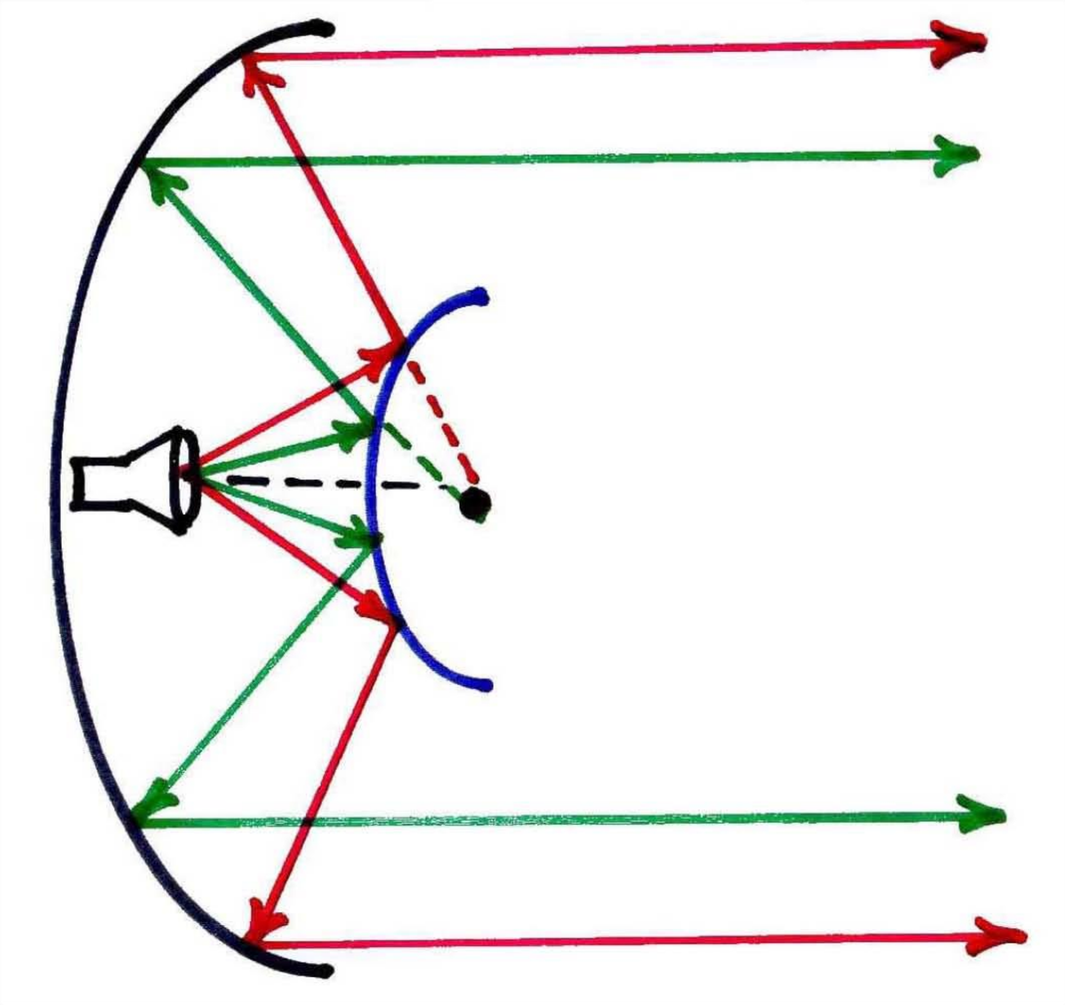


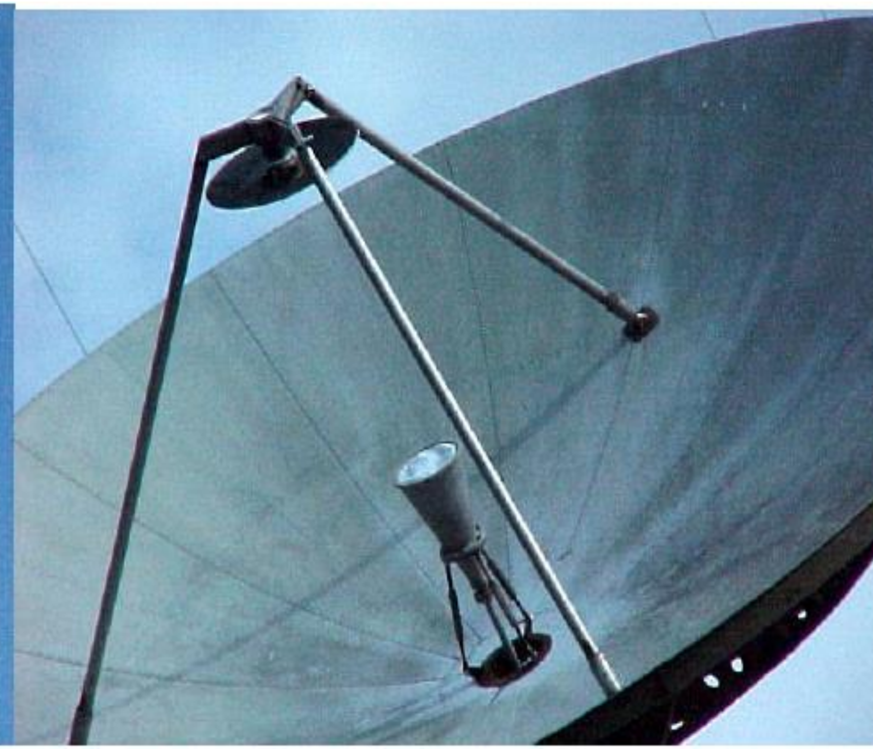
Fig. 15.29(b)

Cassegrain Reflectors

1. Classical form
2. Gregorian form

Classical



3.6 SATELLITE ANTENNAS

Basic Antenna Types and Relationships

Four main types of antennas are used on satellites. These are

1. *Wire antennas*: monopoles and dipoles.
2. *Horn antennas.*
3. *Reflector antennas.*
4. *Array antennas.*

Wire antennas are used primarily at VHF and UHF to provide communications for the TTC&M systems. They are positioned with great care on the body of the satellite in an attempt to provide *omnidirectional* coverage. Most satellites measure only a few wavelengths at VHF frequencies, which makes it difficult to get the required antenna patterns, and there tend to be some orientations of the satellite in which the sensitivity of the TTC&M system is reduced by *nulls* in the antenna pattern.

An *antenna pattern* is a plot of the field strength in the far field of the antenna when the antenna is driven by a transmitter. It is usually measured in *decibels* (dB) below the maximum field strength. The *gain* of an antenna is a measure of the antenna's capability to direct energy in one direction, rather than all around. *Antenna gain* is defined in Chapter 4, Section 4.2. At this point, it will be used with the simple definition given above. A useful principle in antenna theory is *reciprocity*. Reciprocity means that an antenna has the same gain and pattern at any given frequency whether it transmits or receives. An antenna pattern measured when receiving is identical to the pattern when transmitting.

Figure 3.15 shows typical satellite antenna coverage zones. The pattern is frequently specified by its 3-dB beamwidth, the angle between the directions in which the radiated

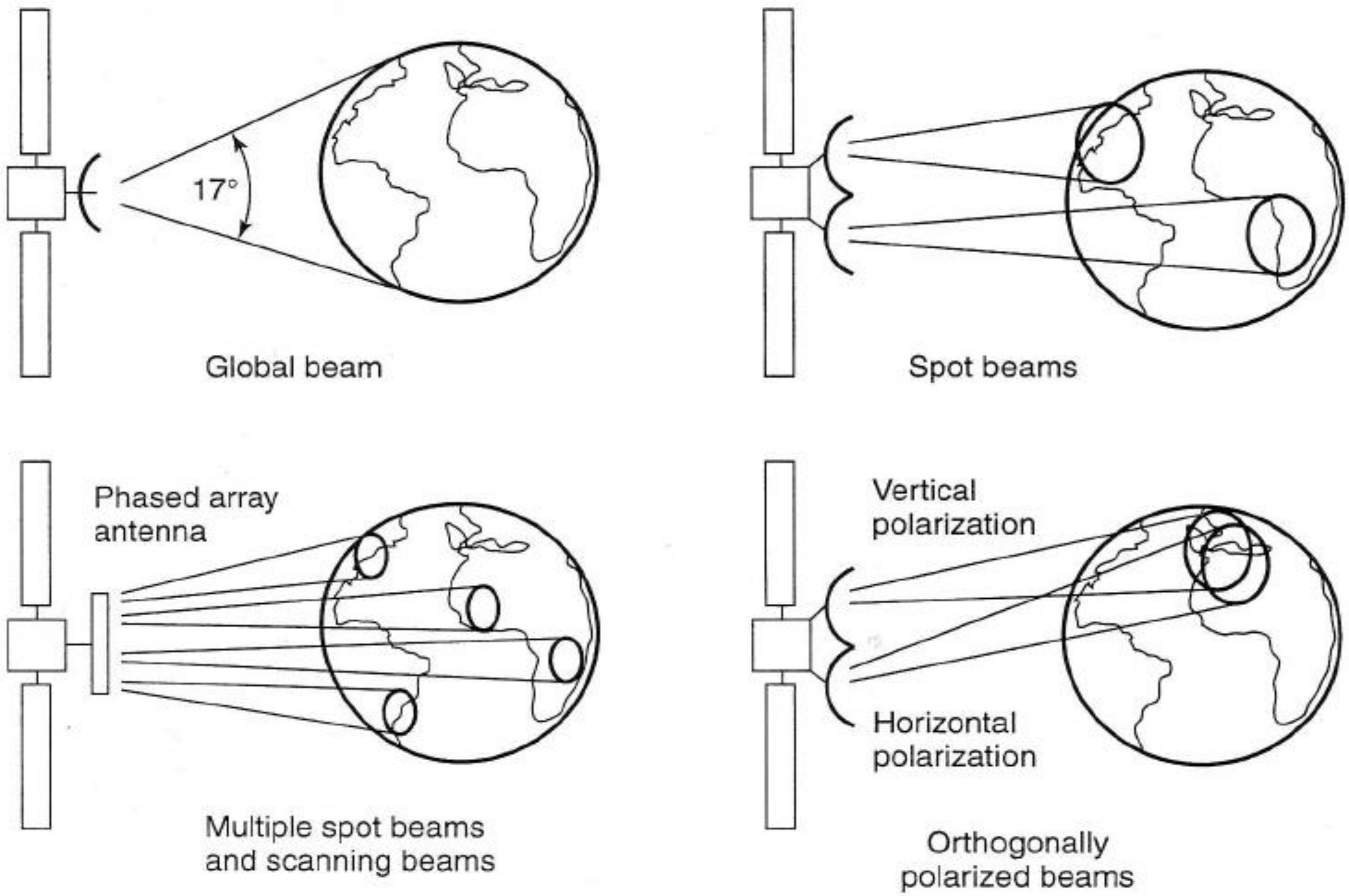


FIGURE 3.15 Typical satellite antenna patterns and coverage zones. The antenna for the global beam is usually a waveguide horn. Scanning beams and shaped beams require phased array antennas or reflector antennas with phased array feeds.

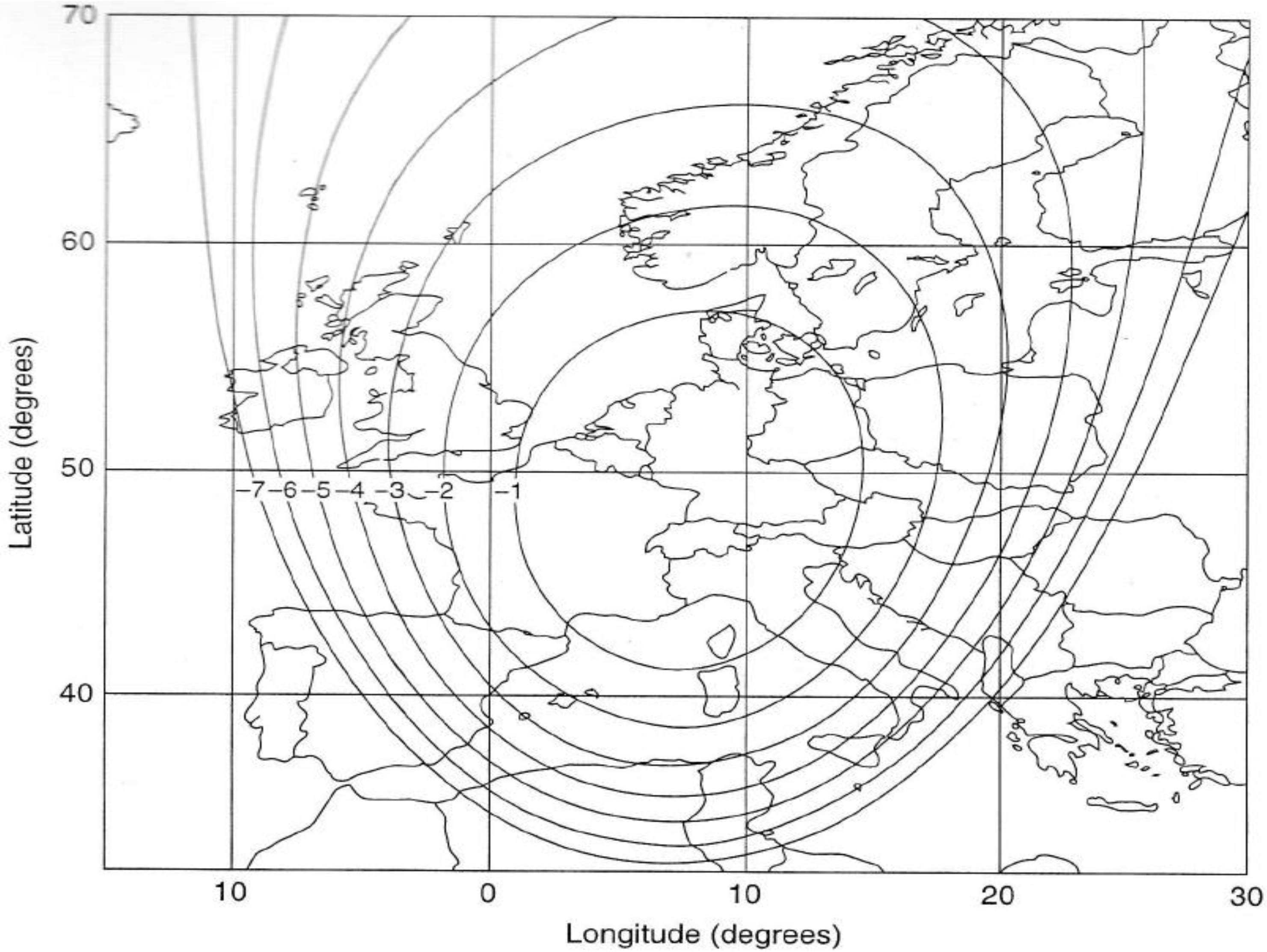


FIGURE 3.16 Contour plot of the spot beam of ESA's OTS satellite projected onto the earth. The contours are in 1 dB steps, normalized to 0 dB at the center of the beam. (Courtesy of ESA.)

(or received) field falls to half the power in the direction of maximum field strength. However, a satellite antenna is used to provide coverage of a certain area, or *zone* on the earth's surface, and it is more useful to have contours of antenna gain as shown in Figure 3.16.

When computing the signal power received by an earth station from the satellite, it is important to know where the station lies relative to the satellite transmit antenna contour pattern, so that the exact EIRP can be calculated. If the pattern is not known, it may be possible to estimate the antenna gain in a given direction if the antenna *boresight* or *beam axis* direction and its beamwidth are known.

Horn antennas are used at microwave frequencies when relatively wide beams are required, as for global coverage. A horn is a flared section of waveguide that provides an aperture several wavelengths wide and a good match between the waveguide impedance and free space. Horns are also used as feeds for reflectors, either singly or in clusters. Horns and reflectors are examples of *aperture antennas* that launch a wave into free space from a waveguide. It is difficult to obtain gains much greater than 23 dB or beamwidths narrower than about 10° with horn antennas. For higher gains or narrow beamwidths a reflector antenna or array must be used.

Reflector antennas are usually illuminated by one or more horns and provide a larger aperture than can be achieved with a horn alone. For maximum gain, it is necessary to generate a plane wave in the aperture of the reflector. This is achieved by choosing a reflector profile that has equal path lengths from the feed to the aperture, so that all the energy radiated by the feed and reflected by the reflector reaches the aperture with the same phase angle and creates a uniform phase front. One reflector shape that achieves this with a point source of radiation is the paraboloid, with a feed placed at its focus. The paraboloid is the basic shape for most reflector antennas, and is commonly used for earth station antennas. Satellite antennas often use modified paraboloidal reflector profiles to tailor the beam pattern to a particular coverage zone. Phased array antennas are also used on satellites to create multiple beams from a single aperture, and have been used by Iridium and Globalstar to generate up to 16 beams from a single aperture for their LEO mobile telephone systems[11].

Some basic relationships in aperture antennas can be used to determine the approximate size of a satellite antenna for a particular application, as well as the antenna gain. More accurate calculations are needed to determine the exact gain, efficiency, and pattern of a satellite antenna, and the interested reader should refer to one of the many excellent texts in this field for details.[12–14] The following approximate relationships will be used here to guide the selection of antennas for a communications satellite.

An aperture antenna has a gain G given by

$$G = \eta_A 4\pi A/\lambda^2 \tag{3.1}$$

where A is the area of the antenna aperture in square meters, λ is the operating wavelength in meters, and η_A is the *aperture efficiency* of the antenna. The aperture efficiency η_A is not easily determined, but is typically in the range 55 to 68% for reflector antennas with single feeds, lower for antennas with shaped beams. Horn antennas tend to have higher efficiencies than reflector antennas, typically in the range 65 to 80%. If the aperture is circular, as is often the case, Eq. (3.1) can be written as

$$G = \eta_A(\pi D/\lambda)^2 \tag{3.2}$$

where D is the diameter of the circular aperture in meters and λ is in meters.

The beamwidth of an antenna is related to the aperture dimension in the plane in which the pattern is measured. A useful rule of thumb is that the 3 dB beamwidth in a given plane for an antenna with dimension D in that plane is

$$\theta_{3\ \mathrm{dB}} \approx 75\lambda/D \text{ degrees} \tag{3.3}$$

where $\theta_{3\ \mathrm{dB}}$ is the beamwidth between half power points of the antenna pattern and D is the aperture dimension in the same units as the wavelength λ. The beamwidth of a horn antenna may depart from Eq. (3.2) quite radically. For example, a small rectangular horn will produce a narrower beam than suggested by Eq. (3.2) in its E plane and a wider beamwidth in the H plane.

Since both Eqs. (3.2) and (3.3) contain antenna dimension parameters, the gain and beamwidth of an aperture antenna are related. For antennas with $\eta_A \approx 60\%$, the gain is approximately

$$G \approx 33{,}000/(\theta_{3\ \mathrm{dB}})^2 \tag{3.4}$$

where $\theta_{3\ \mathrm{dB}}$ is in degrees and G is not in decibels. If the beam has different beamwidths in orthogonal planes, $\theta_{3\ \mathrm{dB}}$ should be replaced by the product of the two 3 dB beamwidths. Values of the constant in Eq. (3.4) vary between different sources, with a range 28,000 to 35,000. The value 33,000 is typical for reflector antennas used in satellite communication systems.

EXAMPLE 3.6.1 *Global Beam Antenna*

The earth subtends an angle of 17° when viewed from geostationary orbit. What are the dimensions and gain of a horn antenna that will provide global coverage at 4 GHz?

If we design our horn to give a circularly symmetric beam with a 3-dB beamwidth of 17° using Eq. (3.3)

$$D/\lambda = 75/(\theta_{3\text{ dB}}) = 4.4$$

At 4 GHz, $\lambda = 0.075$ m, so $D = 0.33$ m (just over 1 ft). If we use a circular horn excited in the TE_{11} mode, the beamwidths in the E and H planes will not be equal and we may be forced to make the aperture slightly smaller to guarantee coverage in the E plane. A *corrugated horn* designed to support the HE hybrid mode has a circularly symmetric beam and could be used in this application. Waveguide horns are generally used for global beam coverage. Reflector antennas are not efficient when the aperture diameter is less than 8λ.

Using Eq. (3.3), the gain of the horn is approximately 100, or 20 dB, at the center of the beam. However, in designing our communication system we will have to use the edge of beam gain figure of 17 dB, since those earth stations close to the earth's horizon, as viewed from the satellite, are close to the 3 dB contour of the transmitted beam. ■

EXAMPLE 3.6.2 *Regional Coverage Antenna*

The continental United States (48 contiguous states) subtends an angle of approximately $6° \times 3°$ when viewed from geostationary orbit. What dimension must a reflector antenna have to illuminate half this area with a circular beam 3° in diameter at 11 GHz?
Can a reflector be used to produce a $6° \times 3°$ beam? What gain would the antenna have?

Using Eq. (3.2), we have for a 3° circular beam

$$D/\lambda = 75/3 = 25$$

and with $\lambda = 0.0272$ m, $D = 0.68$ m (just over 2 ft). The gain of this antenna, from Eq. (3.3) is approximately 35 dB.

To generate a beam with different beamwidths in orthogonal planes we need an aperture with different dimensions in the two planes. In this case, a rectangular aperture $25\lambda \times 12.5\lambda$ would generate a beam $6° \times 3°$, and would have a gain of 32 dB, approximately. In order to illuminate such a reflector, a horn with unequal beamwidths is required, since the reflector must intercept most of the radiation from the feed if it is to have an acceptable efficiency. Rectangular, or more commonly elliptical, outline reflectors are used to generate unequal beamwidths. When orthogonal polarizations are to be transmitted or received, it is better to use a circular reflector with a distorted profile to broaden the beam in one plane, or a feed cluster to provide the appropriate amplitude and phase distribution across the reflector. ■

Satellite Antennas in Practice

The antennas of a communications satellite are often a limiting element in the complete system. In an ideal satellite, there would be one antenna beam for each earth station, completely isolated from all other beams, for transmit and receive. However, if two earth stations are 300 km apart on the earth's surface and the satellite is in geostationary orbit, their angular separation at the satellite is 0.5°. For $\theta_{3\text{ dB}}$ to be 0.5°, D/λ must be 150, which requires an aperture diameter of 11.3 m at 4 GHz. Antennas this large have been flown on satellites (ATS-6 deployed a 2.5 GHz, 10-m diameter antenna, for example), and large unfurled antennas are used to create multiple spot beams from GEO satellites serving mobile users. However, at 20 GHz, an antenna with $D/\lambda = 150$ is only 1.5 m wide, and such an antenna can readily be flown on a 30/20 GHz satellite. A phased array feed is used to create many 0.5° beams which can be clustered to serve the coverage zone of the satellite.

To provide a separate beam for each earth station would also require one antenna feed per earth station if a multiple-feed antenna with a single reflector were used. A compromise between one beam per station and one beam for all stations has been used in many satellites by using zone-coverage beams and orthogonal polarizations within the

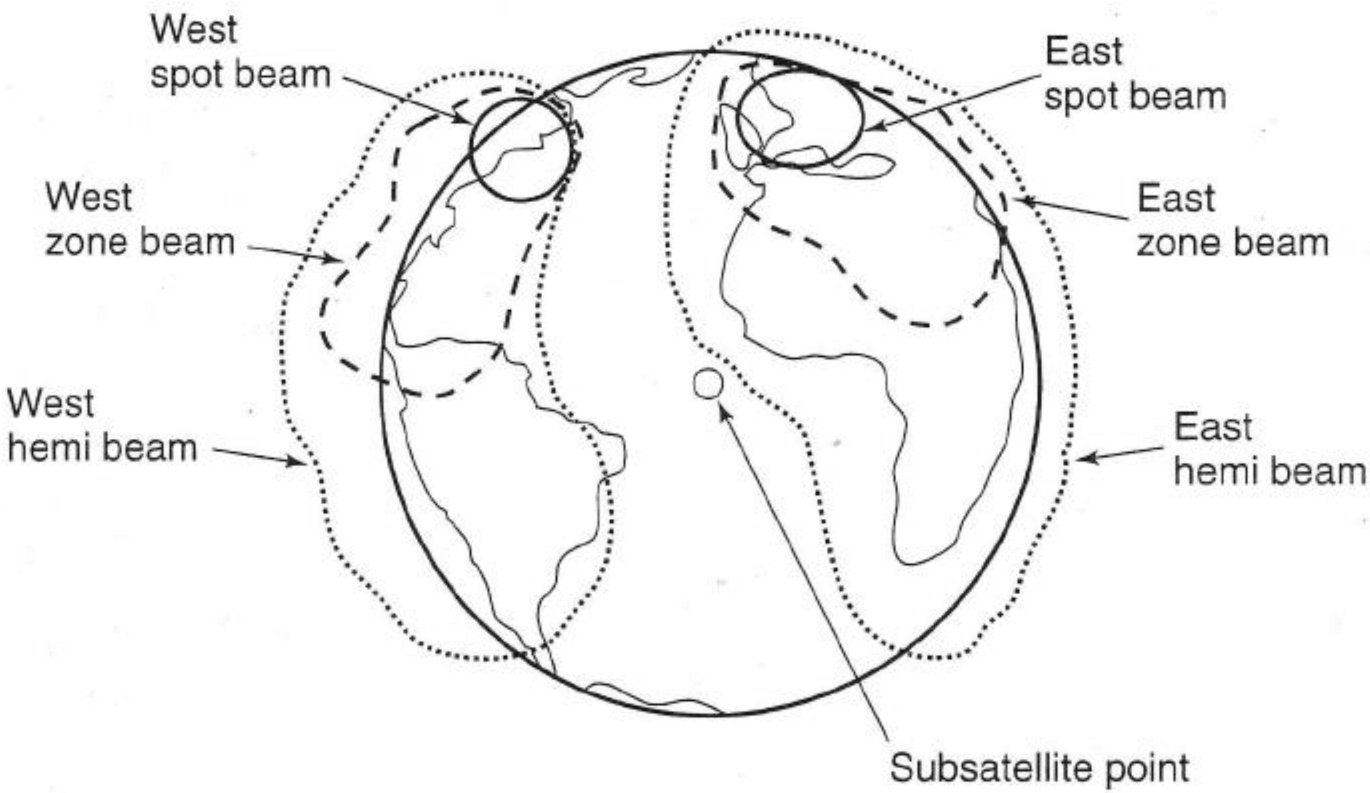


FIGURE 3.17 Typical coverage patterns for Intelsat satellites over the Atlantic Ocean.

same beam to provide more channels per satellite. Figure 3.3b shows a GEO satellite that has four reflector antennas. Each reflector is illuminated by a complex feed that provides the required beam shape to permit communication between earth stations within a given coverage zone. Figure 3.17 shows the coverage zones provided by a typical Intelsat satellite. The largest reflector on the satellite transmits at 4 GHz and produces the “peanut” shaped patterns for the zone beams, which are designed to concentrate the transmitted energy onto densely populated areas such as North America and western Europe where much telecommunications traffic is generated. The smaller antennas are used to provide hemisphere transmit and receive beams, and the 14/11 GHz spot beams. In addition, there are horn antennas providing global beam coverage.

Countries such as the United States create an enormous demand for communication services, and a number of domestic satellite communication systems have been established to meet that demand. In 2000 the geostationary orbit had domestic satellites spaced every 2°, operating at 6/4 GHz and 14/11 GHz from longitude 60° W to 140° W. This encompasses all orbital locations that can be simultaneously viewed by earth stations in the United States and Canada, and each operator has been given a limited number of *orbital slots* in which to place a satellite. As a result, there is a great deal of pressure on the operating companies to obtain the maximum number of channels per satellite in order to give the operator the greatest possible revenue-earning capacity. This has encouraged the development of frequency reuse antennas by means of orthogonal polarizations and multiple beams, the combination of 6/4 and 14/11 GHz communication systems on one satellite, and the use of multilevel digital modulation and TDMA to increase capacity.

The requirements of narrow antenna beams with high gain over a small coverage zone leads to large antenna structures on the satellite. Frequently, the antennas in their operating configuration are too large to fit within the shroud dimensions of the launch vehicle, and must be folded down during the launch phase. Once in orbit, the antennas then can be deployed. In many larger satellites, the antennas use offset paraboloidal reflectors with clusters of feeds to provide carefully controlled beam shapes. The feeds mount on the body of the satellite, close to the communications subsystem, and the reflector is mounted on a hinged arm. Figure 3.18 shows an example of this design of antenna for the INTELSAT VI satellite. For launch, the solid reflectors fold down to provide a compact structure; in orbit, the hinged arms are swung out and locked in place to hold the reflectors in the correct position. When the satellite is in geostationary orbit it is weightless, so very little energy is required to move the large reflector.

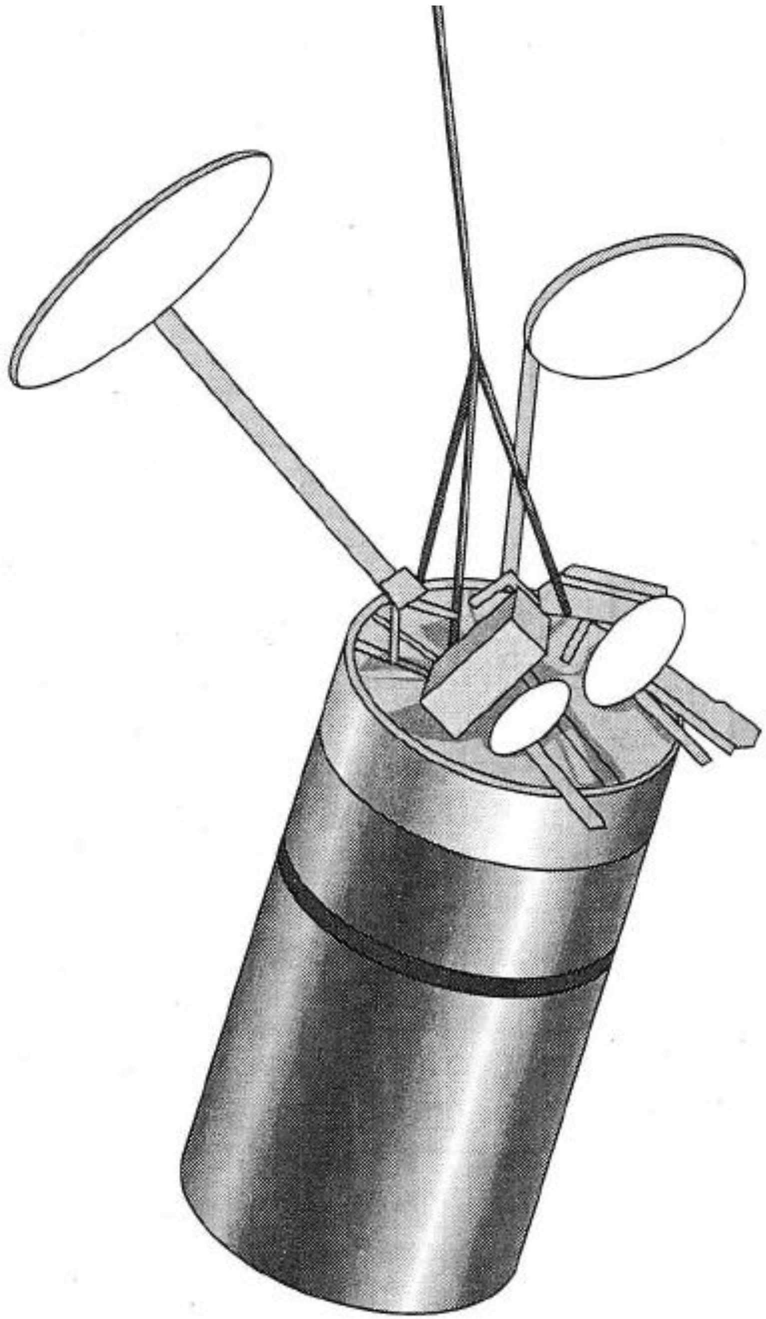


FIGURE 3.18 Intelsat VI satellite on station.

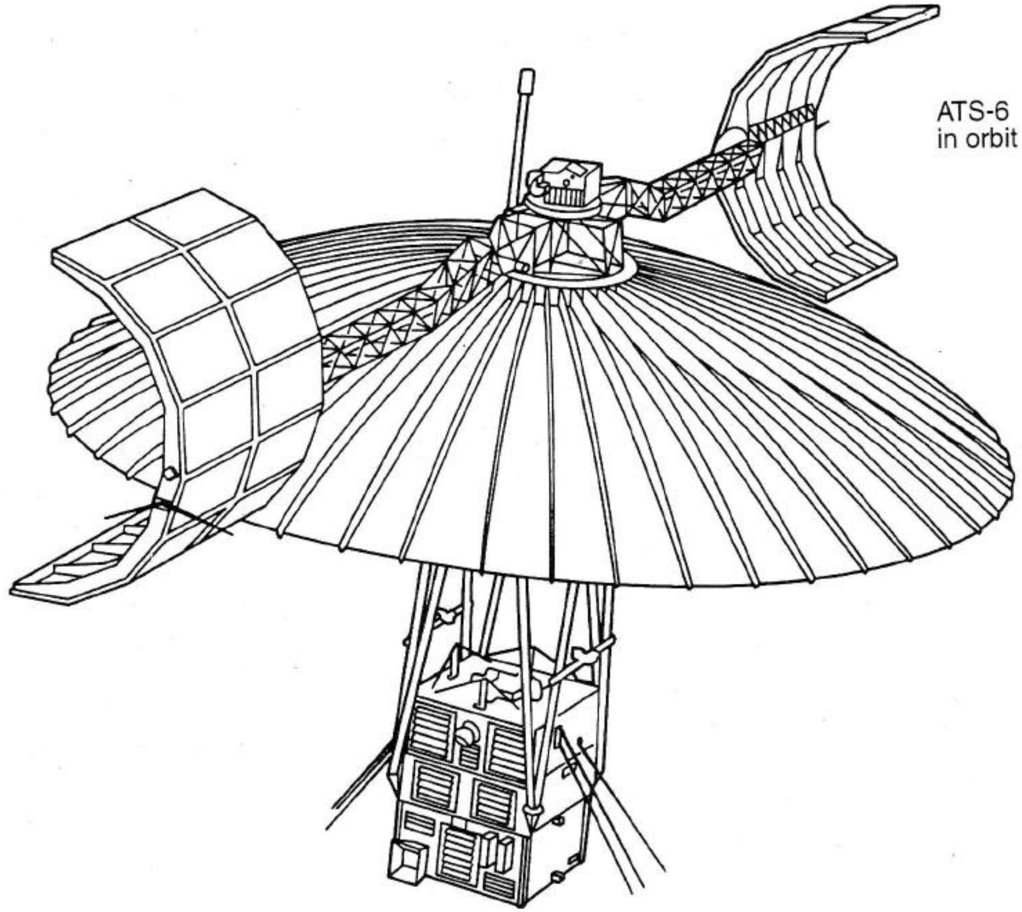


Figure 3.19 shows the deployment sequence used for the 30-ft antenna carried by ATS-6: the antenna was built as a series of petals that folded over each other to make a compact unit during launch, which then unfurled in orbit. The solar sails folded down over the antenna, and were deployed first. Springs or pyrotechnic devices can be used to provide the energy for deployment of antennas or solar sails, with a locking device to ensure correct positioning after deployment. Similar unfurlable antennas are used on GEO satellites that provide satellite telephone service at L band using multiple narrow beams.

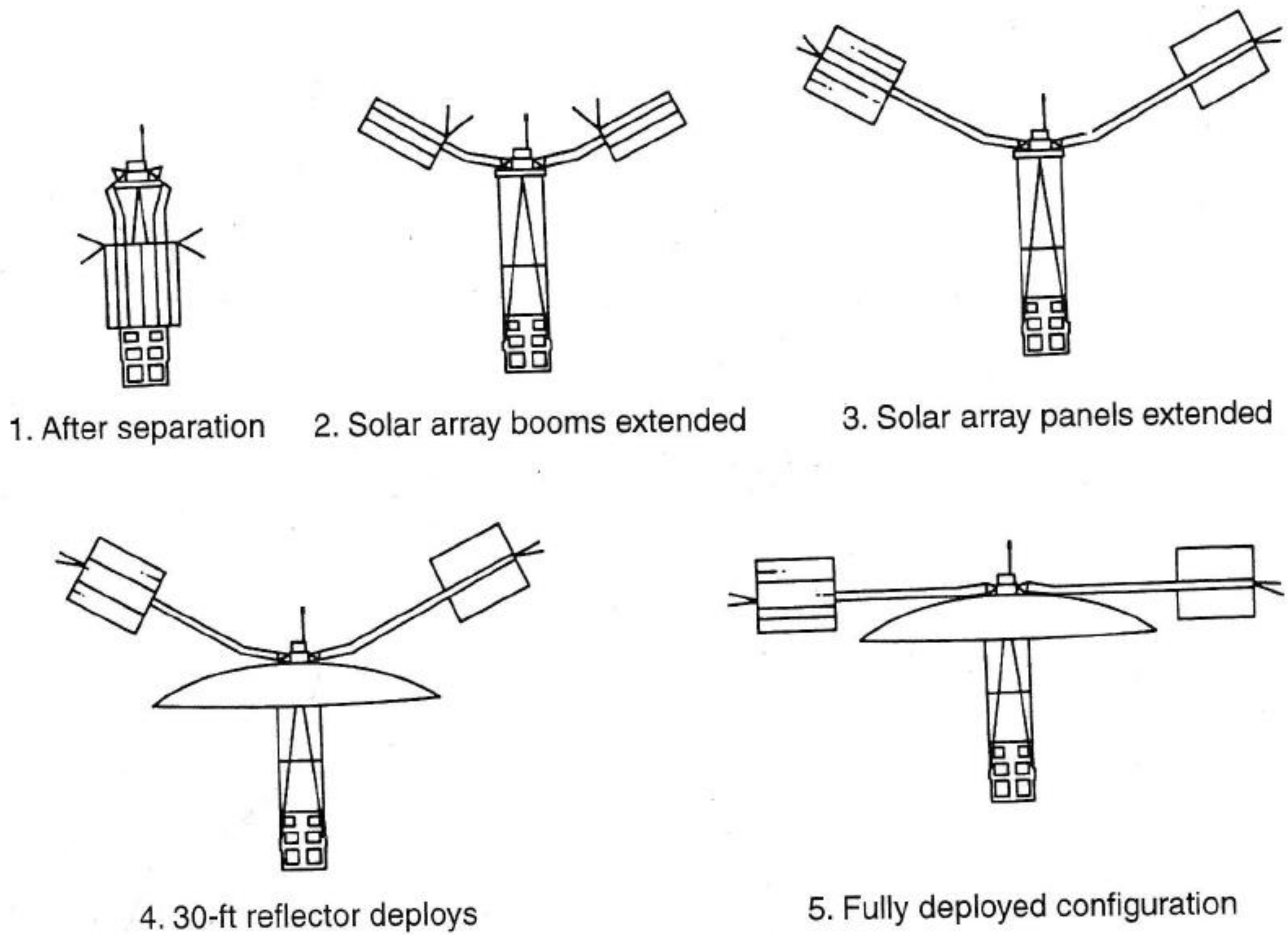


FIGURE 3.19 Deployment sequence of ATS-6 10-m antenna. (Courtesy of NASA.)

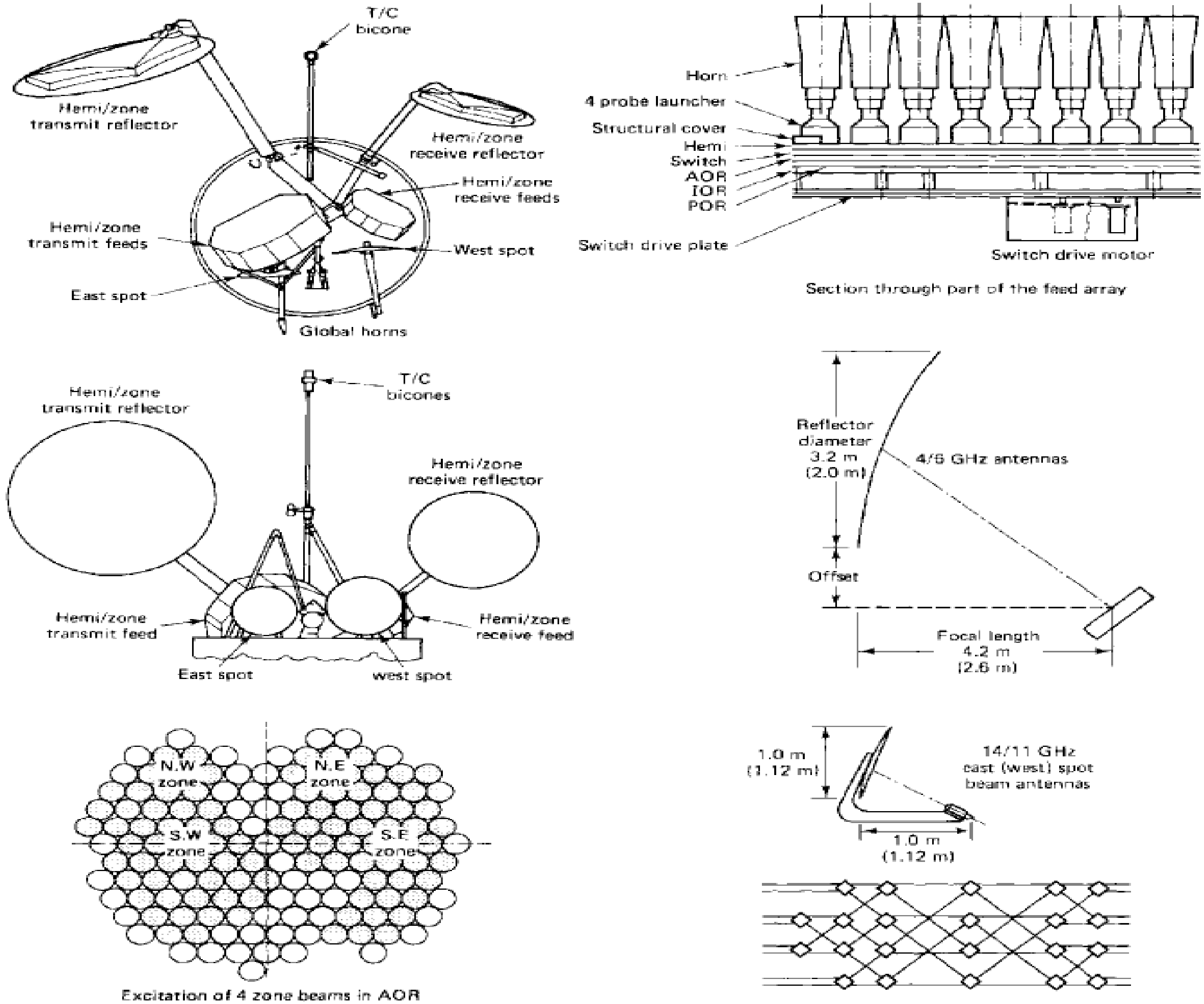


Figure 7.22 The antenna subsystem for the INTELSAT VI satellite. (*From Johnston and Thompson, 1982, with permission.*)

Appendix

Geometry Of A Parabola

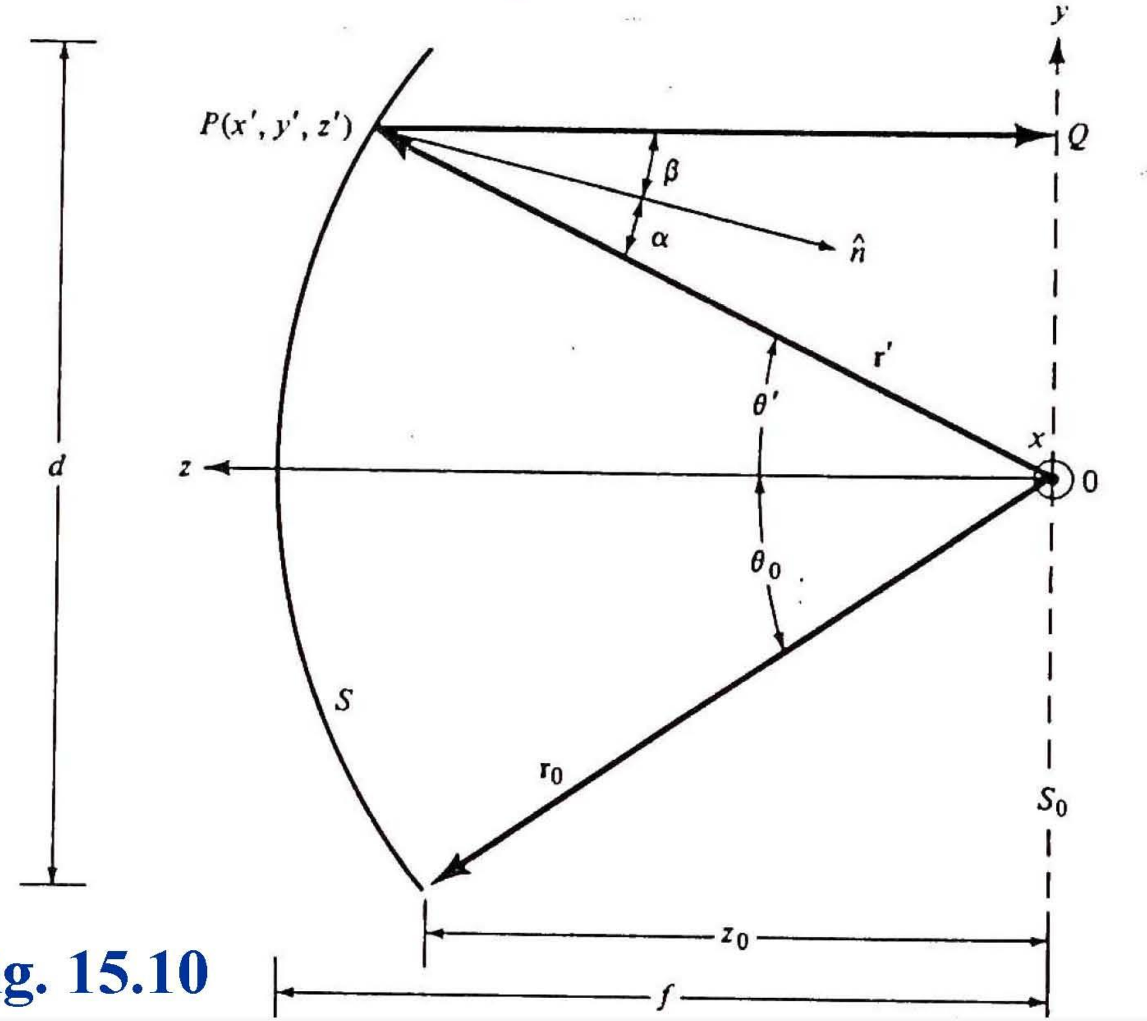


Fig. 15.10

Parabola

$$OP + PQ = \text{Constant} = 2f \quad (15\text{-}12)$$

$$OP = r' \quad (15\text{-}13)$$

$$PQ = r' \cos\theta'$$

$$r' + r' \cos\theta' = r'(1 + \cos\theta') = 2f \quad (15\text{-}14)$$

$$r' = \frac{2f}{1+\cos\theta'} = \frac{f}{\left(\dfrac{1+\cos\theta'}{2}\right)} = \frac{f}{\cos^2\left(\dfrac{\theta'}{2}\right)}$$

$$\boxed{r' = f\sec^2\left(\frac{\theta'}{2}\right)} \qquad \theta \le \theta_o \quad (15\text{-}14a)$$

Parabola Written In Rectangular Coordinates

$$r' + r' \cos\theta' = 2f \qquad (15\text{-}15)$$

$$\sqrt{(x')^2 + (y')^2 + (z')^2} + z' = 2f$$

$$(x')^2 + (y')^2 + (z')^2 = (2f - z')^2 = 4f^2 - 4fz' + z'^2$$

$$\boxed{(x')^2 + (y')^2 = 4f(f - z')} \qquad (x')^2 + (y')^2 \le (d/2)^2$$

(15-15a)

Univector Normal To Surface

$$r' = f\sec^2\left(\frac{\theta'}{2}\right) \qquad \theta \le \theta_0$$

$$f = \frac{r'}{\sec^2\left(\frac{\theta'}{2}\right)} = r'\cos^2\left(\frac{\theta'}{2}\right)$$

$$\boxed{f - r'\cos^2\left(\frac{\theta'}{2}\right) = S = 0} \qquad (15\text{-}16)$$

$$\underline{N} = \nabla S = \nabla\left[f - r'\cos^2\left(\frac{\theta'}{2}\right)\right]$$

$$= \hat{a}'_r \frac{\partial S}{\partial r'} + \hat{a}'_\theta \frac{1}{r'}\frac{\partial S}{\partial \theta'}$$

$$\underline{N} = -\hat{a}'_r \cos^2\left(\frac{\theta'}{2}\right) + \hat{a}'_\theta \cos\left(\frac{\theta'}{2}\right)\sin\left(\frac{\theta'}{2}\right) \quad (15\text{-}17)$$

$$\hat{n} = \frac{\underline{N}}{|\underline{N}|} = -\hat{a}'_r \cos\left(\frac{\theta'}{2}\right) + \hat{a}'_\theta \sin\left(\frac{\theta'}{2}\right) \quad (15\text{-}18)$$

$$\cos\alpha = -\hat{a}'_r \cdot \hat{n} = -\hat{a}'_r \cdot \left[-\hat{a}'_r \cos\left(\frac{\theta'}{2}\right) + \hat{a}'_\theta \sin\left(\frac{\theta'}{2}\right) \right] = \cos\left(\frac{\theta'}{2}\right) \tag{15-19}$$

$$\cos\beta = -\hat{a}_z \cdot \hat{n} = -\hat{a}_z \cdot \left[-\hat{a}'_r \cos\left(\frac{\theta'}{2}\right) + \hat{a}'_\theta \sin\left(\frac{\theta'}{2}\right) \right] \tag{15-20}$$

$$= -\left(\hat{a}'_r \cos\theta' - \hat{a}'_\theta \sin\theta'\right) \cdot \left[-\hat{a}'_r \cos\left(\frac{\theta'}{2}\right) + \hat{a}'_\theta \sin\left(\frac{\theta'}{2}\right) \right]$$

$$= \cos\left(\frac{\theta'}{2}\right) \tag{15-21}$$

$$\cos(\alpha) = \cos(\beta) = \cos\left(\frac{\theta'}{2}\right) \Rightarrow \therefore \ \alpha = \beta = \frac{\theta'}{2}$$ Snell's Law

From Fig. 15.10

$$\theta_0 = \tan^{-1}\left(\frac{d/2}{z_0}\right) \qquad (15\text{-}22)$$

$$z_0 = f - \frac{x_0^2 + y_0^2}{4f} = f - \frac{(d/2)^2}{4f} = f - \frac{d^2}{16f}$$

(15-23)

$$\theta_0 = \tan^{-1}\left[\frac{\dfrac{d}{2}}{f - \dfrac{d^2}{16f}}\right] = \tan^{-1}\left[\frac{\dfrac{1}{2}\left(\dfrac{f}{d}\right)}{\left(\dfrac{f}{d}\right)^2 - \dfrac{1}{16}}\right] \quad (15\text{-}24)$$

$$f = \frac{d}{4}\cot\left(\frac{\theta_0}{2}\right)$$

(15-25)

$$\frac{f}{d} = \frac{1}{4}\cot\left(\frac{\theta_0}{2}\right)$$

θ_o (degrees)		f/d
10°		2.857
20°		1.418
30°		0.933
40°		0.687
50°		0.536
60°		0.433
70°		0.357
80°		0.298
90°		0.250
100°		0.210
120°		0.144