Module 3: VHF, UHF AND MICROWAVE ANTENNAS

Helical Antennas - Helical geometry, Helix modes, Practical design considerations for monofilar helical antenna in axial and normal modes, Horn antenna, Rectangular Microstrip antennas - Introduction, Geometrical Features, Characteristics Advantages and limitations, Reflector types-paraboloidal, cassegrain, feed methods for parabolic reflectors; RF radiation hazards and solutions, Illustrative problems.

Helical Antennas

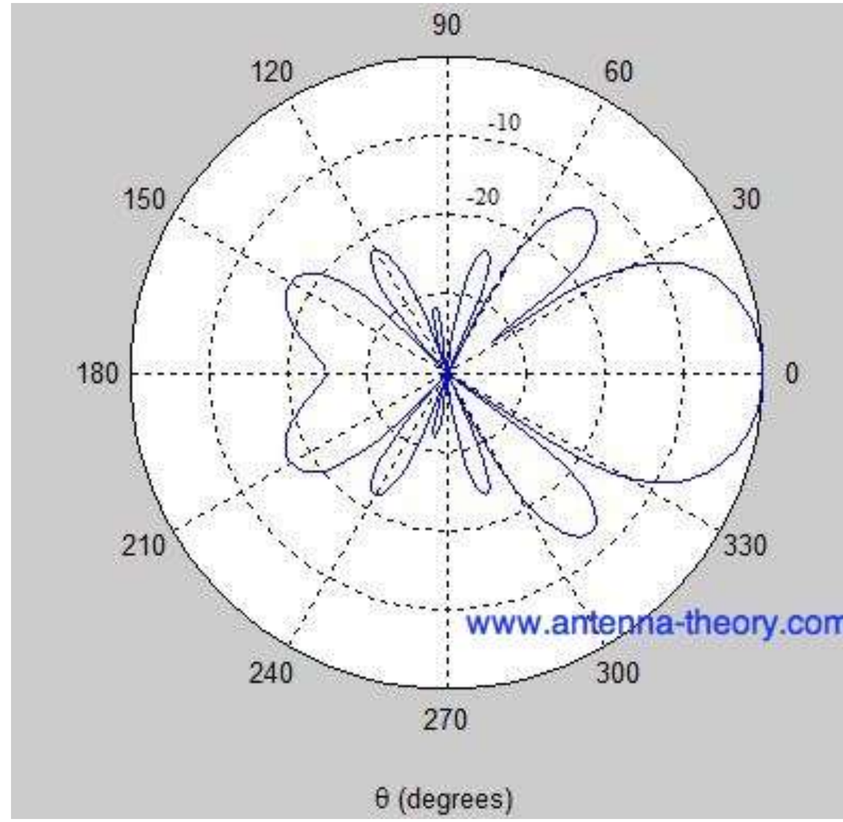
- A conducting wire wound in the form of a screw thread forming a helix can be used as a broad band antenna.

- Its construction requires
  - Thick Copper wire or tube
  - Metal (Ground) Plane
  - Coaxial Cable

- The ground plane can take different forms - flat with typical the diameter of the ground plane should be at least $3\lambda/4$, cupped in the form of a cylindrical cavity or in the form of a frustrum cavity.

- The helix is usually connected to the center conductor of a coaxial transmission line at the feed point with the outer conductor of the line attached to the ground plane.


www.antenna-theory.com


90
120
60
-10
150
30
-20
180
0
210
330
www.antenna-theory.com
240
300
270
θ (degrees)

Construction

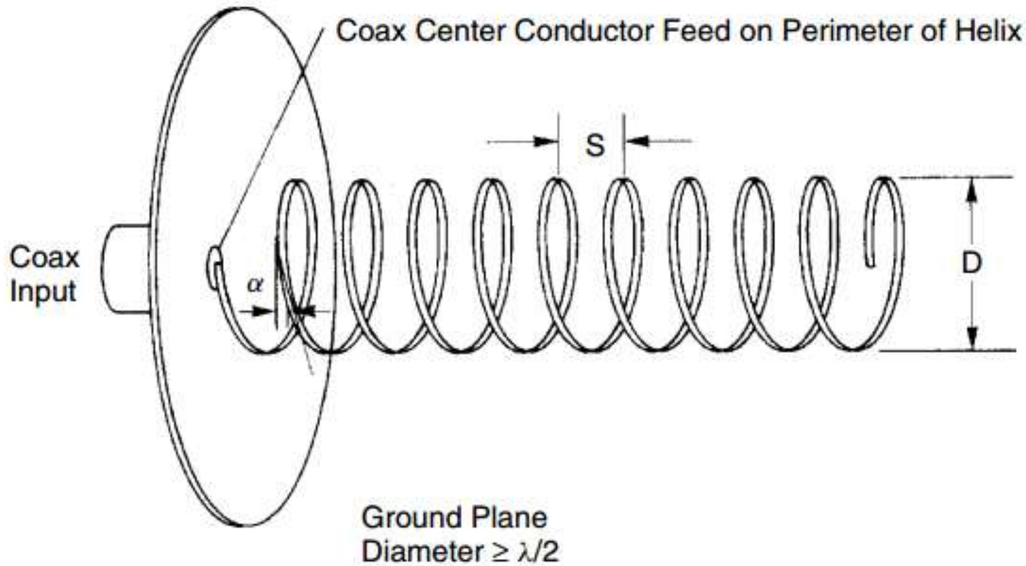
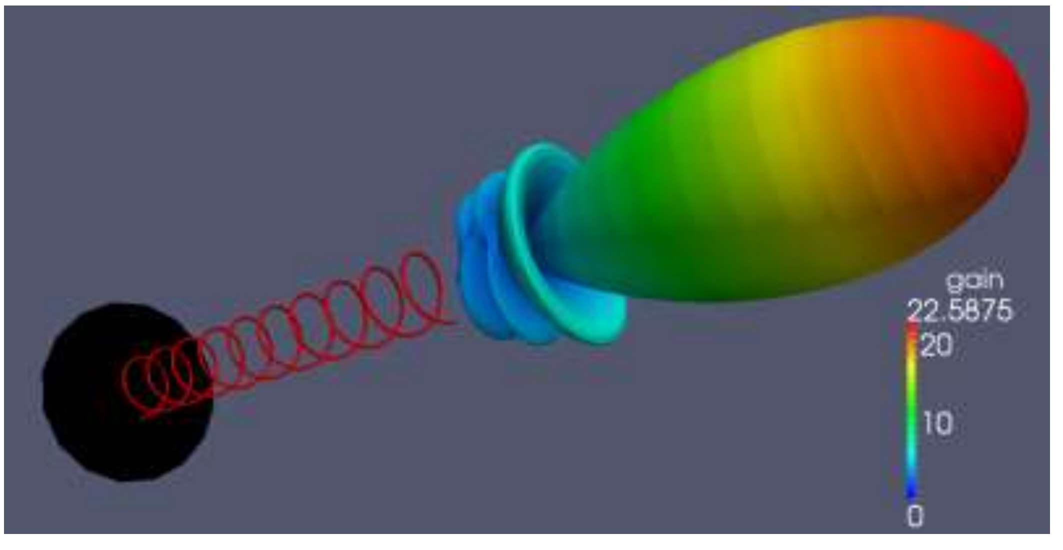


FIGURE 10-17 Axial mode helical wire antenna (RHC).



Array of axial mode helical antennas for satellite tracking and Acquisition

Rubber Ducky antenna used in portable radios

Wireless communication LAN antenna

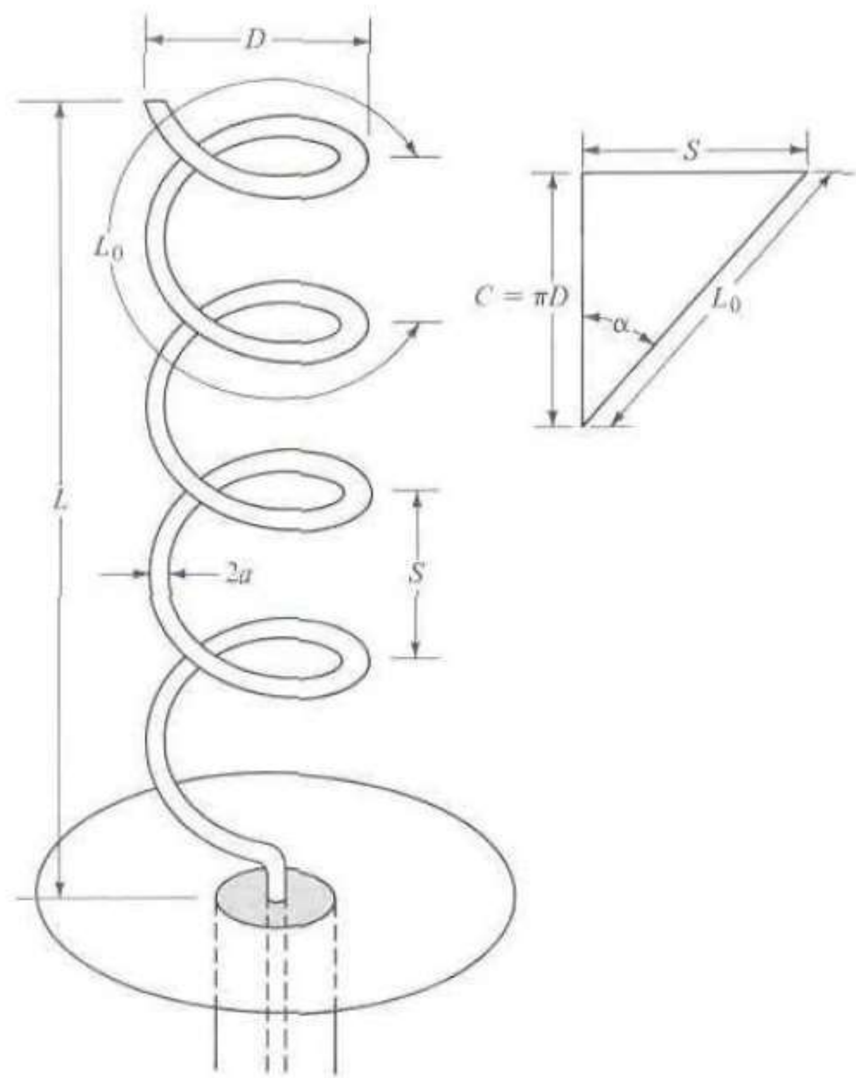
Contd.,

- The geometrical configuration of a helix consists usually of N turns, diameter D and spacing S between each turn.

- The total length of the antenna is L = NS while the total length of the wire is

$$L_n = NL_o = N\sqrt{S^2 + C^2}$$

- where Lo is the length of the wire between each turn and C = πD is the circumference of the helix.



Contd.,

- The pitch angle is the angle formed by a line tangent to the helix wire and a plane perpendicular to the helix axis.

$$\alpha = \tan^{-1}\left(\frac{S}{\pi D}\right) = \tan^{-1}\left(\frac{S}{C}\right)$$

- When α = 0°, then the winding is flattened and the helix reduces to a loop antenna of N turns.

- When α = 90° then the helix reduces to a linear wire.

- When 0° < α < 90°. then a true helix is formed with a circumference greater than zero but less than the circumference when the helix is reduced to a loop (α = 0°)

- The radiation characteristics of the antenna can be varied by controlling the size of its geometrical properties compared to the wavelength. T

- The input impedance is critically dependent upon the pitch angle and the size of the conducting wire, especially near the feed point, and it can be adjusted by controlling their values.

- The general polarization of the antenna is elliptical. However circular and linear polarizations can be achieved over different frequency ranges.

Contd.,

- The helical antenna can operate in many modes; however the two principal ones are the normal (broadside) and the axial (End - Fire) modes.

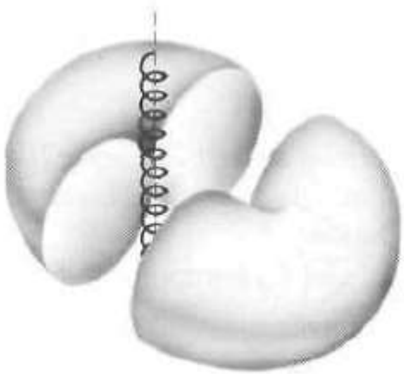
- The axial (End - Fire) mode is usually the most practical because it can achieve circular polarization over a wider bandwidth (usually 2:1) and it is more efficient.

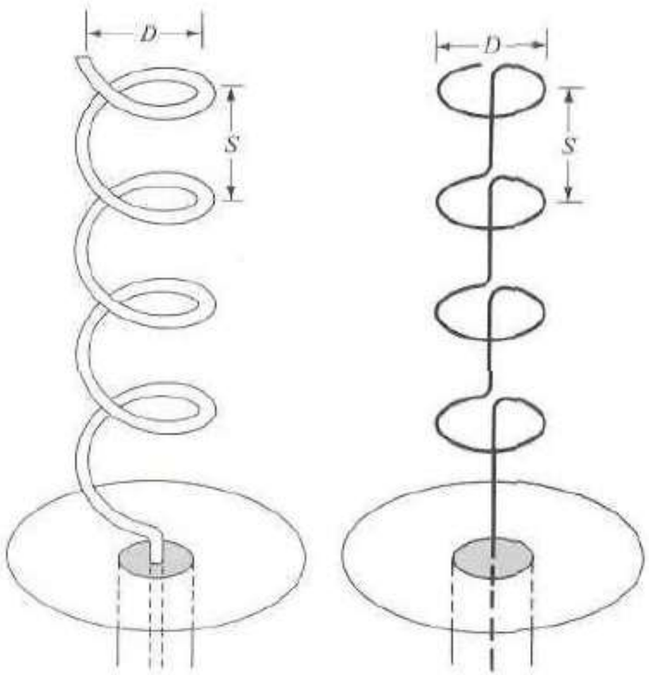
- Because an elliptically polarized antenna can be represented as the suns of two orthogonal linear components in time-phase quadrature, a helix can always receive a signal transmitted from a rotating linearly polarized antenna.

- Therefore helices are usually positioned on the ground for space telemetry applications of satellites, space probes, and ballistic missiles to transmit or receive signals that have undergone Faraday rotation by traveling through the ionosphere.

Modes of Operation

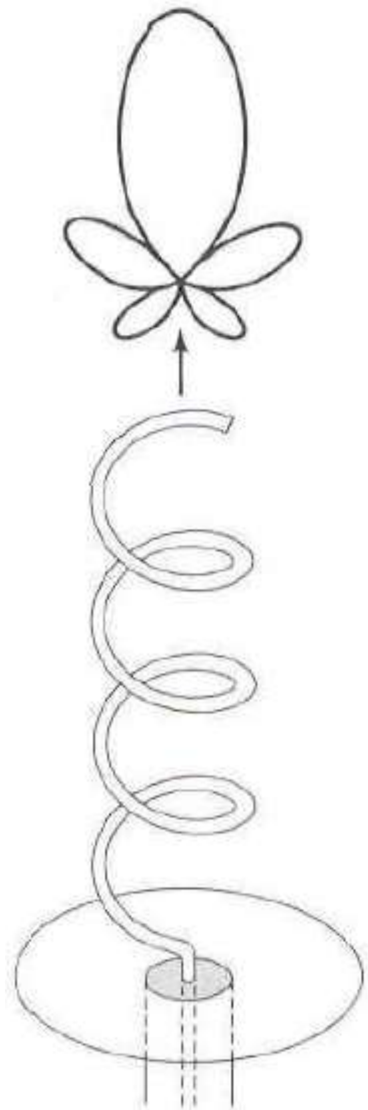
Normal (Broadside) Mode

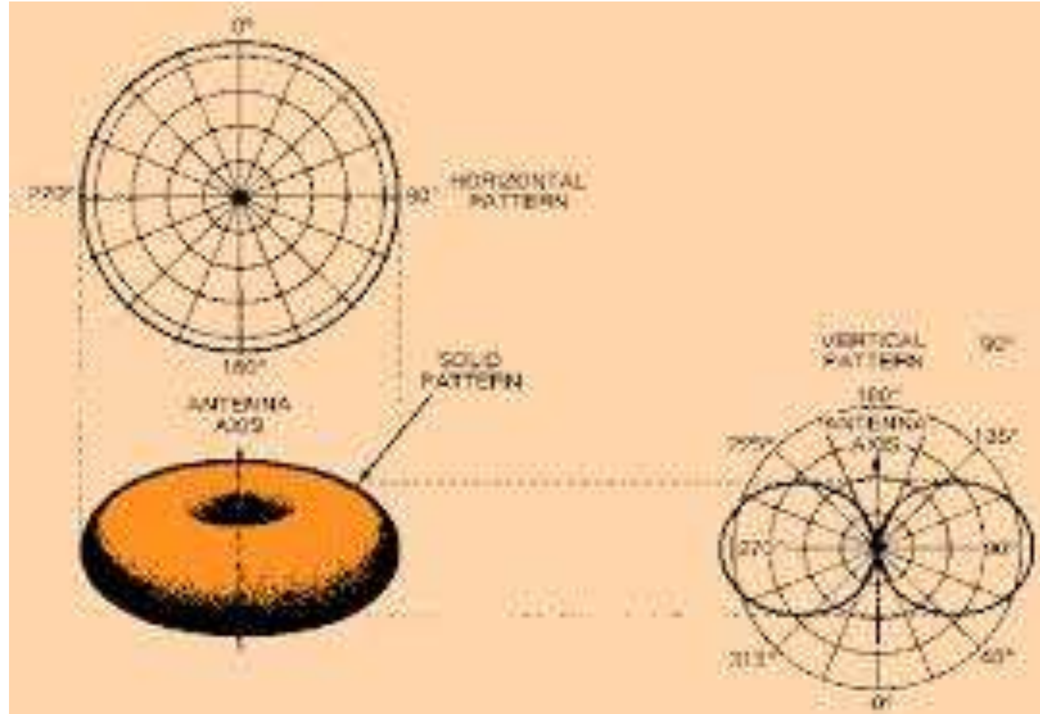




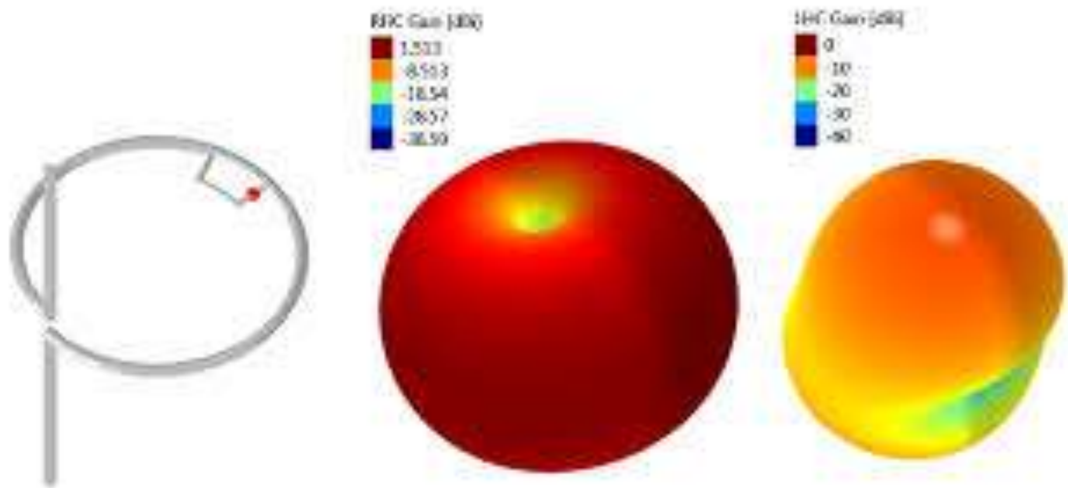
(a) Normal mode (b) Equivalent

Axial (End-Fire) Mode





HORIZONTAL PATTERN
SOLID PATTERN
ANTENNA AXIS
VERTICAL PATTERN
ANTENNA AXIS



Normal (Broadside) Mode

- In the normal mode of operation the field radiated by the antenna is maximum in a plane normal to the helix axis and minimum along its axis, which is a figure-eight rotated about its axis similar to that of a linear dipole of $l < \lambda$ or a small loop ($a << \lambda$).

- To achieve the normal mode of operation, the dimensions of the helix are usually small compared to the wavelength ($NL_o << \lambda$).

- The geometry of the helix reduces to a loop of diameter D when the pitch angle approaches zero and to a linear wire of length S when it approaches 90^o.

- Since the limiting geometries of the helix are a loop and a dipole, the far-field radiated by a small helix in the normal mode can be described in terms of E_θ and E_Φ, components of the dipole and loop, respectively.

- In the normal mode, it can be thought that the helix consists of N small loops and N short dipoles connected together in series as shown in fig.b

Contd.,

- The fields are obtained by superposition of the fields from these elemental radiators.
- The planes of the loops are parallel to each other and perpendicular to the axes of the vertical dipoles.
- The axes of the loops and dipoles coincide with the axis of the helix.
- Since in the normal mode the helix dimensions are small, the current throughout its length can be assumed to be constant and its relative far-field pattern to be independent of the number of loops and short dipoles.
- Thus its operation can be described accurately by the sum of the fields radiated by a small loop of radius D and a short dipole of length S, with its axis perpendicular to the plane of the loop, and each with the same constant current distribution.
- The far-zone electric field radiated by a short dipole of length S and constant current I_0 is E_θ, and it is given by

$$E_\theta = j\eta \frac{kI_0 S e^{-jkr}}{4\pi r}\sin\theta$$

- Where l is being replaced by S. In addition the electric field radiated by a loop is E_Φ, and it is given by

$$E_\Phi = \eta \frac{k^2 a^2 I_0 e^{-jkr}}{4r}\sin\theta = \eta \frac{k^2\left(\frac{D}{2}\right)^2 I_0 e^{-jkr}}{4r}\sin\theta$$

- The two components are in time-phase Quadrature, a necessary but not sufficient condition for circular or elliptical polarizations.

Contd.,

- The ratio of the magnitudes of the Eθ and EΦ components is defined as the axial ratio (AR), is given by

$$AR = \frac{|E_\theta|}{|E_\phi|} = \frac{\frac{\eta\beta I_o S}{4\pi r}}{\frac{\eta\beta^2\left(\frac{D}{2}\right)^2 I_o}{4r}} = \frac{2\lambda S}{(\pi D)^2}$$

- By varying the D and/or S the axial ratio attains values of $0 \le AR \le \infty$.

- The value of AR = 0 is a special case and occurs when $E_\theta = 0$ leading to a linearly polarized wave of horizontal polarization (the helix is a loop).

- When AR = ∞, $E_\Phi = 0$ and the radiated wave is linearly polarized with vertical polarization (the helix is a vertical dipole).

- Another special case is the one when AR is unity (AR = 1) and occurs when

$$\frac{2\lambda S}{(\pi D)^2} = 1$$

$$C = \pi D = \sqrt{2S\lambda}$$

$$\tan\alpha = \frac{S}{\pi D} = \frac{\pi D}{2\lambda}$$

Contd.,

- When the dimensional parameters of the helix satisfy the above relation, the radiated field is circularly polarized in all directions other than 0 = 0° where the fields vanish, otherwise not circularly polarized.

- The progression of polarization change can be described geometrically by beginning with the pitch angle of zero degrees ($\alpha = 0^{0}$), which reduces the helix to a loop with linear horizontal polarization.

- As α increases, the polarization becomes ellipitical with the major axis being horizontally polarized.

- When α is such that $\frac{C}{\lambda} = \sqrt{\frac{2S}{\lambda}}$, AR = I and we have circular polarization.

- For greater values of α, the polarization again becomes elliptical but with the major axis vertically polarized.

- Finally when a = 90° the helix reduces to a linearly polarized vertical dipole.

- To achieve the normal mode of operation. it has been assumed that the current throughout the length of the helix is of constant magnitude and phase. This is satisfied to a large extent provided the total length of the helix wire NL_0 is very small compared to the wavelength ($L_n \ll \lambda$) and its end is terminated properly to reduce multiple reflections.

- Because of the critical dependence of its radiation characteristics on its geometrical dimensions, which must be very small compared to the wavelength, this mode of operation is very narrow in bandwidth and its radiation efficiency is very small.

- Practically this mode of operation is limited, and it is seldom utilized.

Axial Mode

- A more practical mode of operation, which can be generated with great ease, is the axial or endfire mode.
- In this mode of operation, there is only one major lobe and its maximum radiation intensity is along the axis of the helix.
- The minor lobes are at oblique angles to the axis.
- To excite this mode, the diameter D and spacing S must be large fractions of the wavelength.
- To achieve circular polarization, primarily in the major lobe, the circumference of the helix must be in the $3/4 < C/\lambda < 4/3$ range (with $C/\lambda = 1$ near optimum), and the spacing about $S \approx \lambda/4$. The pitch angle is usually $12° < \alpha < 14°$.
- Most often the antenna is used in conjunction with a ground plane, whose diameter is at least $\lambda/2$, and it is fed by a coaxial line.
- However other types of feeds (such as waveguides and dielectric rods) are possible, especially at microwave frequencies.
- The dimensions of the helix for this mode of operation are not as critical, thus resulting in a greater bandwidth.

Design Procedure

- The terminal impedance of a helix radiating in the axial mode is nearly resistive with values between 100 and 200 ohms.
- Smaller values, even near 50 ohms, can be obtained by properly designing the feed.
- Empirical expressions, based on a large number of measurements, have been derived and they are used to determine a number of parameters.
- The input impedance (purely resistive) is obtained by ->
- All these relations are approximately valid provided 12° < α< 14°, ¾ < C/λ < 4/3 and N > 3.
- The far-field pattern of the helix, has been developed by assuming that the helix consists of an array of N identical turns (each of nonuniform current and identical to that of the others), a uniform spacing S between them, and the elements are placed along the z-axis.
- The cosθ term represents the field pattern of a single turn, and the last term is the array factor of a uniform array of N elements.
- The total field is obtained by multiplying the field from one turn with the array factor (pattern multiplication).

$$\boxed{R \simeq 140\left(\frac{C}{\lambda}\right)}$$

which is accurate to about ±20%, the half-power beamwidth by

$$\boxed{\text{HPBW (degrees)} \simeq \frac{52\lambda^{3/2}}{C\sqrt{NS}}}$$

the beamwidth between nulls by

$$\boxed{\text{FNBW (degrees)} \simeq \frac{115\lambda^{3/2}}{C\sqrt{NS}}}$$

the directivity by

$$\boxed{D_0 \text{ (dimensionless)} \simeq 15N\frac{C^2S}{\lambda^3}}$$

the axial ratio (for the condition of increased directivity) by

$$\boxed{\text{AR} = \frac{2N+1}{2N}}$$

and the normalized far-field pattern by

$$E = \sin\left(\frac{\pi}{2N}\right)\cos\theta\frac{\sin[(N/2)\psi]}{\sin[\psi/2]}$$

where

$$\psi = k_0\left(S\cos\theta - \frac{L_0}{p}\right)$$

$$p = \frac{L_0/\lambda_0}{S/\lambda_0 + 1} \quad \text{For ordinary end-fire radiation}$$

$$p = \frac{L_0/\lambda_0}{S/\lambda_0 + \left(\frac{2N+1}{2N}\right)} \quad \text{For Hansen-Woodyard end-fire radiation}$$

Contd.,

- The value of p is the ratio of the wave velocity along the helix wire to that in free space, and it is selected according for ordinary end-fire radiation or for Hansen-Woodyard end-fire radiation.

- For ordinary end-fire the relative phase among the various turns of the helix (elements of the array) is given by

$$\psi = k_0 S \cos\theta + \beta$$

- where d = S is the spacing between the turns of the helix.

- For an end-fire design, the radiation from each one of the turns along θ=0° must be in-phase. Since the wave along the helix wire between turns travels a distance L_o with a wave velocity $v = pv_0$ (p> 1) where v_o is the wave velocity in free space) and the desired maximum radiation is along 0 = 0°, then for ordinary endfire radiation is equal to

$$\psi = (k_0 S \cos\theta - kL_0)_{\theta=0^\circ} = k_0\left(S - \frac{L_0}{p}\right) = -2\pi m, \quad m = 0, 1, 2, \ldots$$

- Solving for p leads to

$$p = \frac{L_0/\lambda_0}{S/\lambda_0 + m}$$

- in For m = 0 and p = 1, L_0 = S. This corresponds to a straight wire (α = 90°), and not a helix. Therefore the next value is m=1 and its corresponding first transmission mode for a helix.

$$p = \frac{L_0/\lambda_0}{S/\lambda_0 + 1}$$

Contd.,

- In a similar manner, for Hansen woodyard end-fire radiation,

$$\psi = (k_0 S \cos\theta - kL_0)_{\theta=0^\circ} = k_0\left(S - \frac{L_0}{p}\right) = -\left(2\pi m + \frac{\pi}{N}\right), \quad m = 0, 1, 2, \ldots$$

- When solved for p leads to

$$p = \frac{L_0/\lambda_0}{S/\lambda_0 + \left(\frac{2mN + 1}{2N}\right)}$$

- For m=1, reduces to

$$p = \frac{L_0/\lambda_0}{S/\lambda_0 + \left(\frac{2N + 1}{2N}\right)}$$

Feed Design

- The nominal impedance of a helical antenna operating in axial mode computed using 100-200 ohm.
- Practical transmission lines such as coaxial cables, have characteristic impedance of about 50 ohms.
- In order to provide a better match, the input impedance of the helix must be reduced to near that value.
- One way to effectively control the input impedance of the helix is to properly design the first 1/4 turn of the helix which is next to the feed.
- To bring the input impedance of the helix from nearly 150 ohms down to 50 ohms, the wire of the first 1/4 turn should be flat in the form of a strip and the transition into a helix should be very gradual.
- This is accomplished by making the wire from the feed, at the beginning of the formation of the helix, in the form of a strip of width W by flattening it and nearly matching the ground plane which is covered with a dielectric slab of height

$$h = \frac{w}{\dfrac{377}{\sqrt{\epsilon_r}\, Z_0} - 2}$$

- w = width of strip conductor of the helix starting at the feed
- ε_r= dielectric constant of the dielectric slab covering the ground plane
- Zo = characteristic impedance of the input transmission line

Contd.,

- Typically the strip configuration of the helix transitions from the strip to the-regular circular wire and the designed pitch angle of the helix very gradually within the first 1/4-1/2 turn.

- This modification decreases the characteristic impedance of the conductor-ground plane effective transmission line, and it provides a lower impedance over a substantial but reduced bandwidth.

- Eg., a 50-ohm helix has a VSWR of less than 2:1 over a 40% bandwidth compared to a 70% bandwidth for a 140-ohm helix. In addition, the 50-ohm helix has a VSWR of less than 1.2:1 over a 12% bandwidth as contrasted to a 20% bandwidth for one of 140 ohms.

- A simple and effective way of increasing the thickness of the conductor near the feed point will be to bond a thin metal strip to the helix conductor.

- Eg., a metal strip 70-mm wide was used to provide a 50-ohm impedance in ahelix whose conducting wire was 13-mm in diameter and it was operating at 230.77MHz.

- Helical antennas can be made to operate one of two ways, normal-mode helical and axial-mode helical, which both have specific advantages over a straight monopole antenna.

- Normal-mode helical:

  - The antenna is the same length as the monopole, still produces as omnidirectional radiation pattern, and is linearly polarized.

  - The antenna is low profile because it is curled up in a helical nature such that the total height is much less. This variant operates as an electrically small antenna (ESA).

  - Low profile antennas are desired for many vehicular, aerospace, and handheld applications because it is often burdensome to have a large antenna sticking out.

- Axial-mode helical:

  - This antenna is larger and has specific ring size and spacing.

  - It is a traveling wave antenna that radiates in an end-fire (in the direction of the helix) fashion.

  - This mode is advantageous because it can be circularly polarized (CP), instead of linearly polarized as a monopole is.

  - Many communication systems, including many satellite communication systems operate in CP so it is advantageous to use a CP antenna over a linearly polarized antenna to efficiently communicate.

- **Advantages**
  - Very directive
  - Good gain
  - Omnidirectional
  - Robust
  - Less sensitivity
  - Wide bandwidth
  - These aerials are usually used where the polarisation of the wave is circular, or linear but either unknown or varying. In applications where one or both ends are rotating, circular polarisation will reduce (the effects of) signal fading due to mismatched polarisation between sender and receiver.
  - they are less fussy about being cut to exactly the right length
  - less prone to being detuned by nearby metalwork, height above ground - common problems with "straight" aerials.

- **Disadvantages**
  - Mechanical construction
  - Bulky size, Space requirements
  - Costly
  - Easily detuned by nearby objects

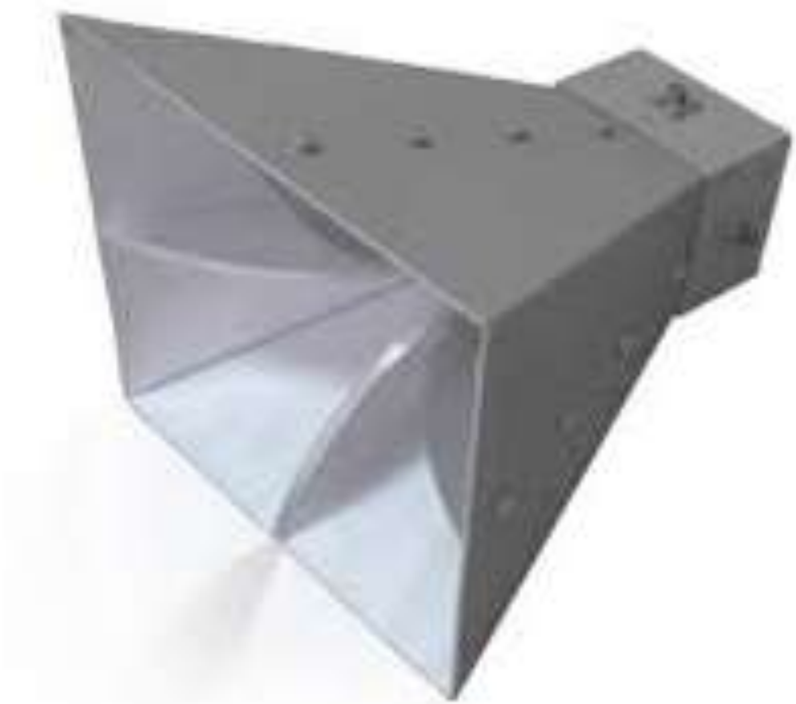
Applications

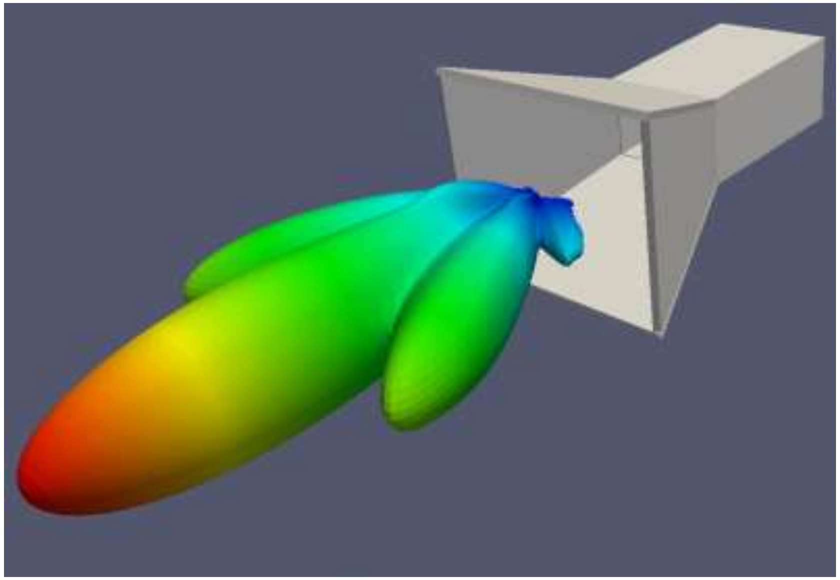
- Broadside Helical
  - Used as standard FM receiving antennas for many factory produced motor vehicles
  - Used in basic style of aftermarket HF and VHF mobile Helical
  - Used in situations where a smaller antenna is an important operational factor. These antennas were first used with Citizen's Band Radios n the US and Australia during the late 1960s and are widely used today as typical FM receiving antennas in automobiles.
- Axial (End - Fire) Helical
  - Best suited for animal tracking and space communications
  - Used with hand held satellite communication devices such as telephones, radios, and global positioning systems.

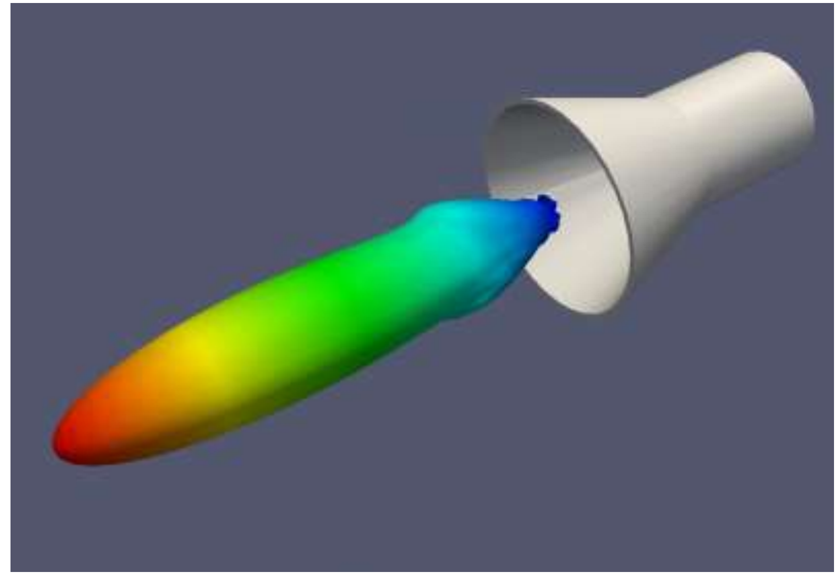
Horn Antenna

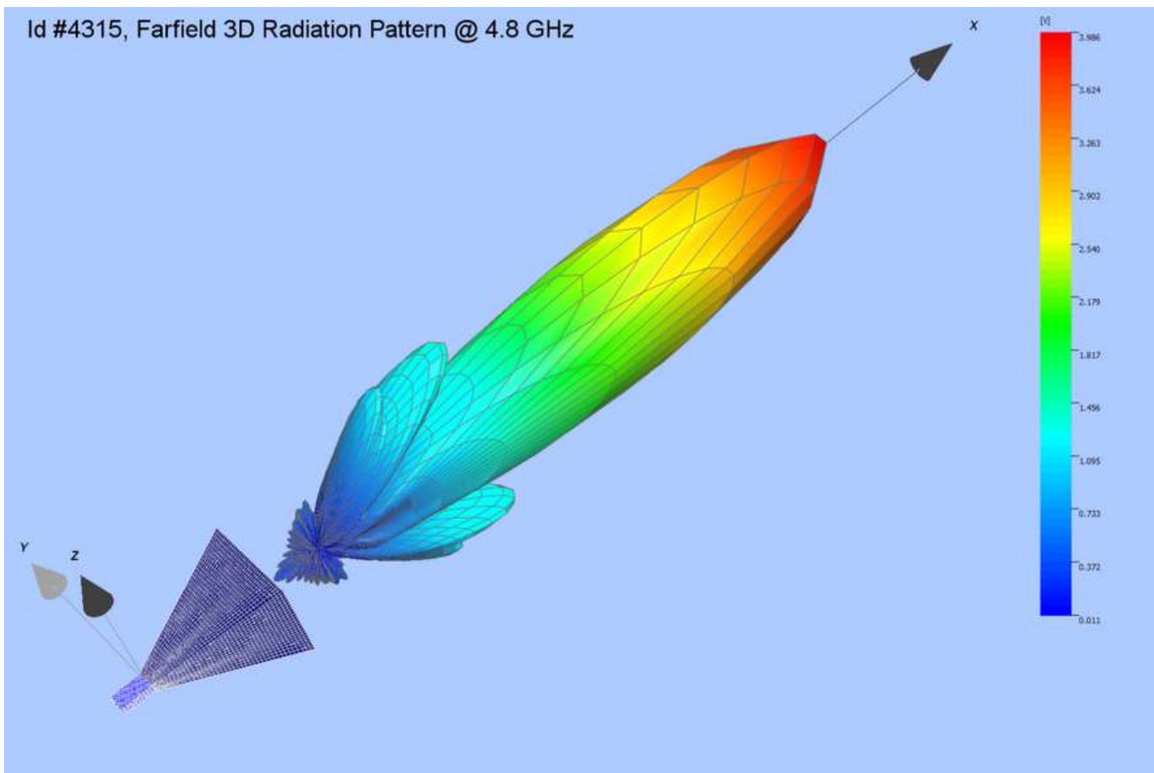
- Construction
- Radiation Pattern
- Characteristics
- Advantages
- Disadvantages
- Applications

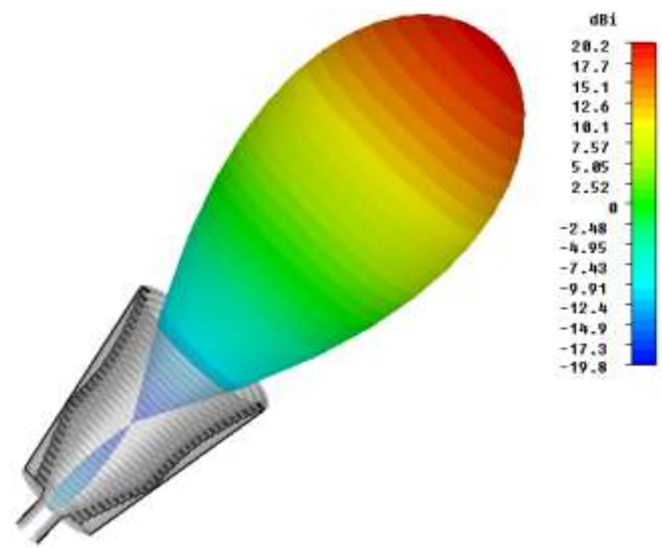


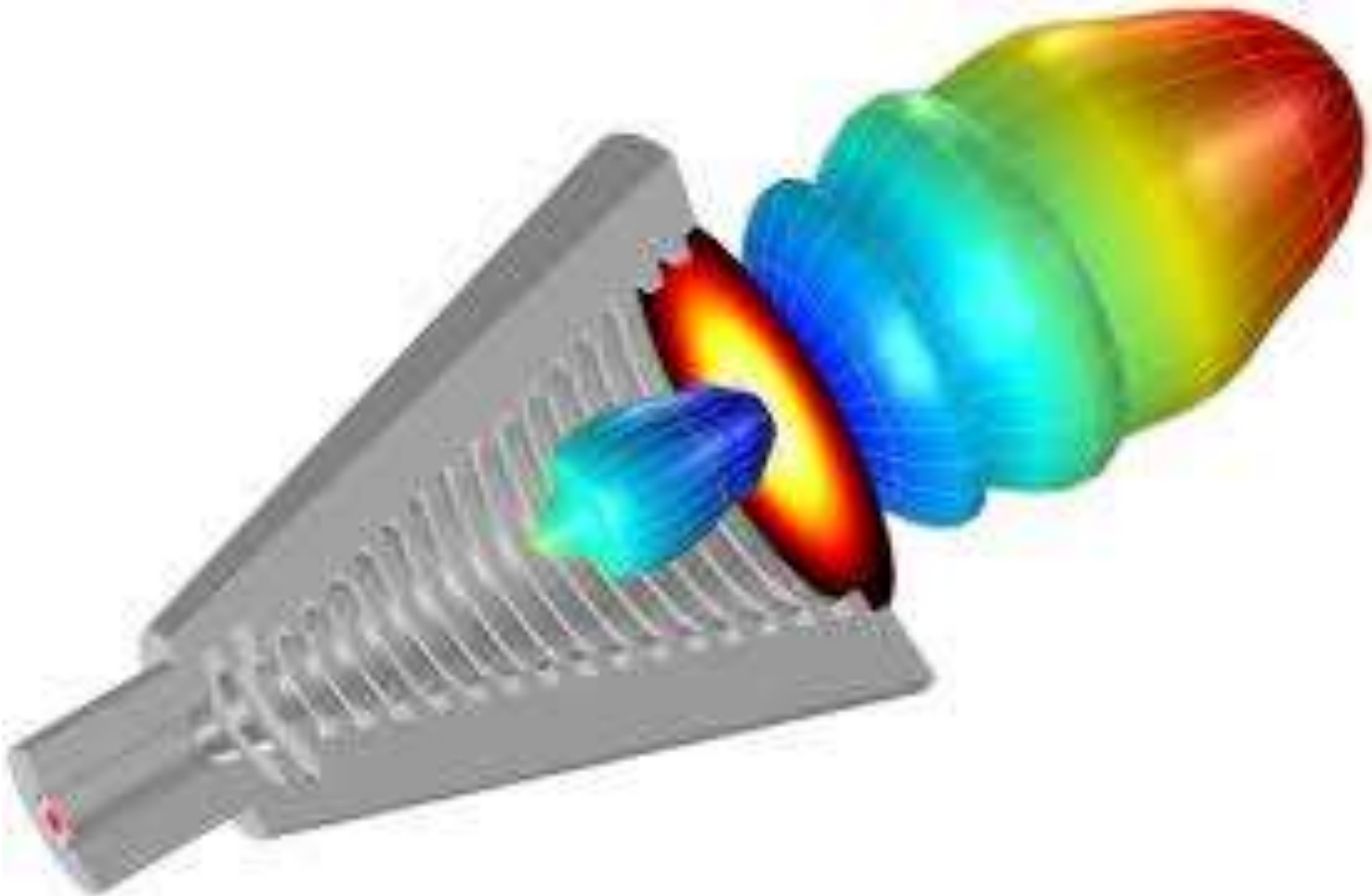





Id #4315, Farfield 3D Radiation Pattern @ 4.8 GHz
X
Y
Z
[V]
3.624
3.263
2.902
2.540
2.179
1.817
1.456
1.095
0.733
0.372
0.011


dBi
20.2
17.7
15.1
12.6
10.1
7.57
5.05
2.52
0
-2.48
-4.95
-7.43
-9.91
-12.4
-14.9
-17.3
-19.8



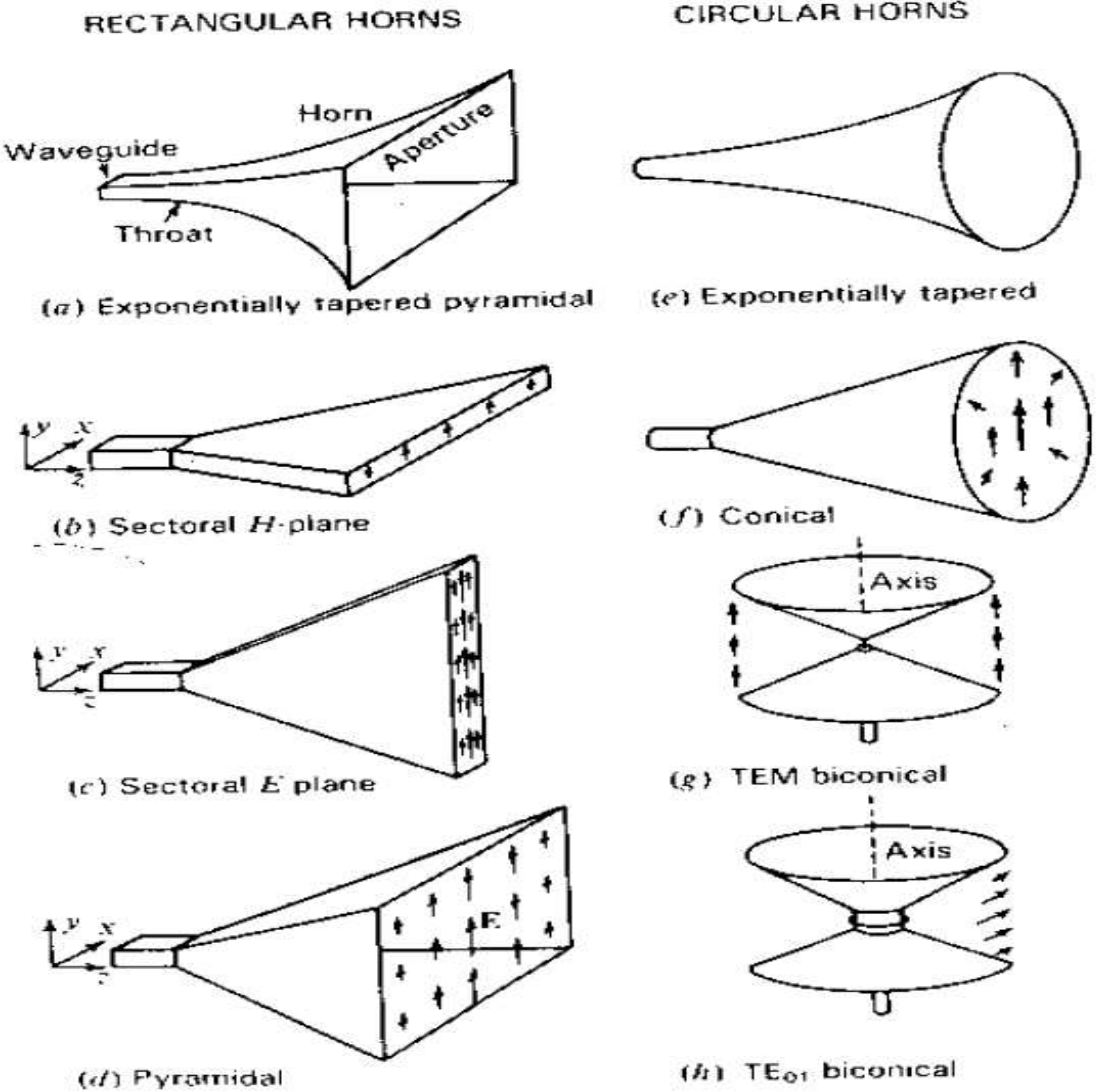


Figure 13-20 Types of rectangular and circular horn antennas. Arrows indicate E-field directions.

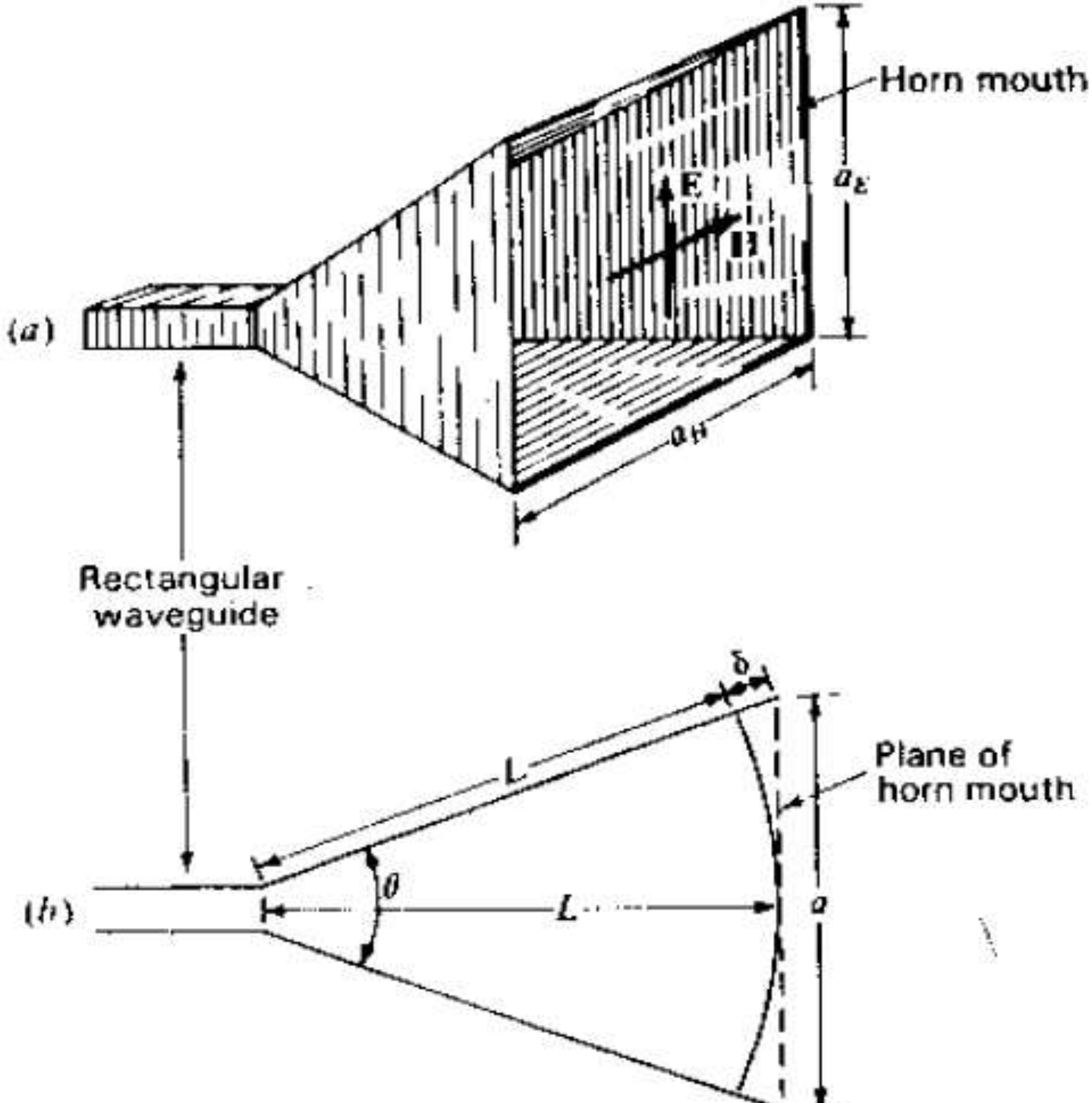


Figure 13-21 (*a*) Pyramidal horn antenna. (*b*) Cross section with dimensions used in analysis. The diagram can be for either *E*-plane or *H*-plane cross sections. For the *E* plane the flare angle is θ_E and aperture a_E. For the *H* plane the flare angle is θ_H and the aperture a_H. See Fig. 13-22.

$$\cos\frac{\theta}{2} = \frac{L}{L+\delta} \qquad \delta_0 = \frac{L}{\cos(\theta/2)} - L$$

$$\sin\frac{\theta}{2} = \frac{a}{2(L+\delta)} \qquad L = \frac{\delta_0\cos(\theta/2)}{1-\cos(\theta/2)}$$

$$\tan\frac{\theta}{2} = \frac{a}{2L}$$

where θ = flare angle (θ_E for E plane, θ_H for H plane)
a = aperture (a_E for E plane, a_H for H plane)
L = horn length

From the geometry we have also that

$$L = \frac{a^2}{8\delta} \qquad (\delta \ll L)$$

and

$$\theta = 2\tan^{-1}\frac{a}{2L} = 2\cos^{-1}\frac{L}{L+\delta}$$

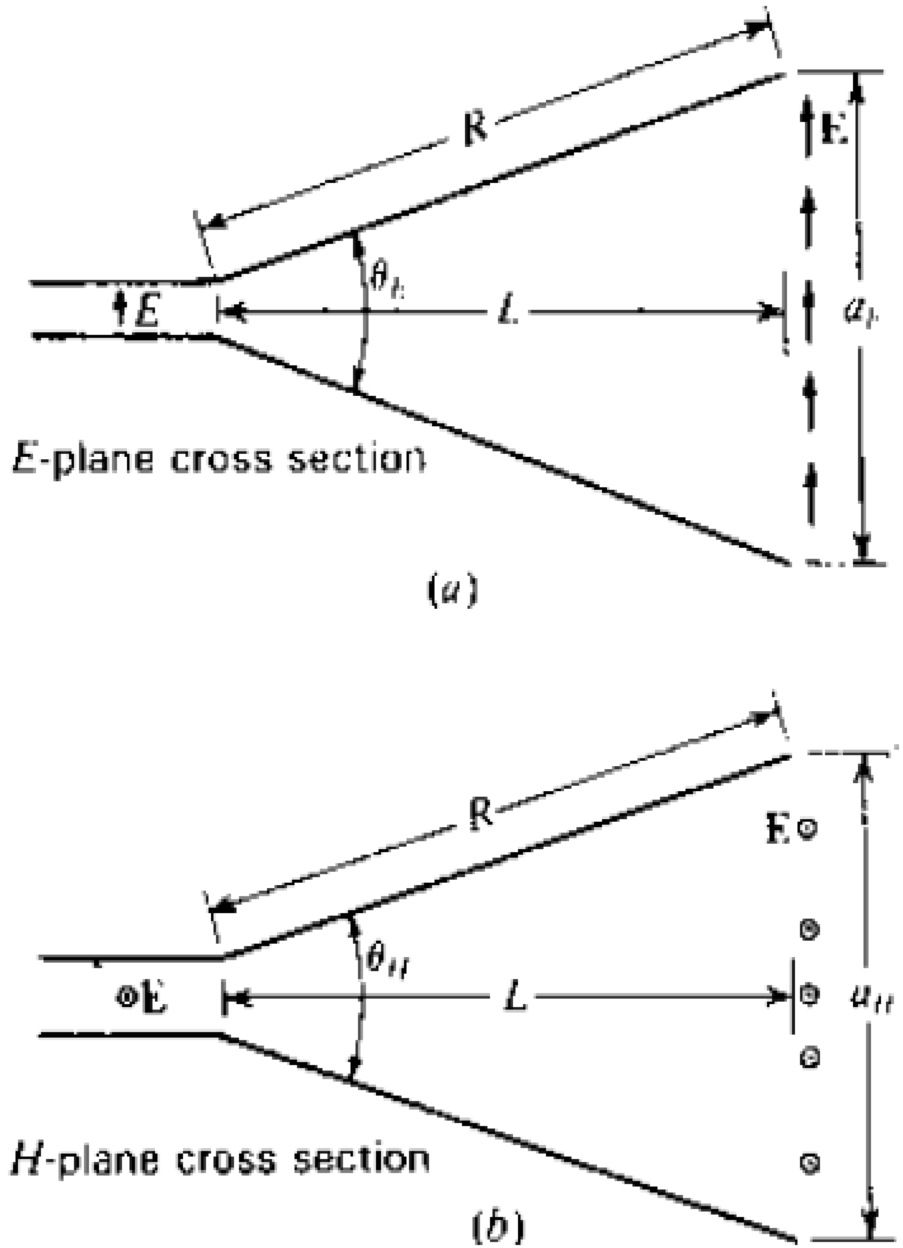


Figure 13-22 *E*-plane and *H*-plane cross sections.

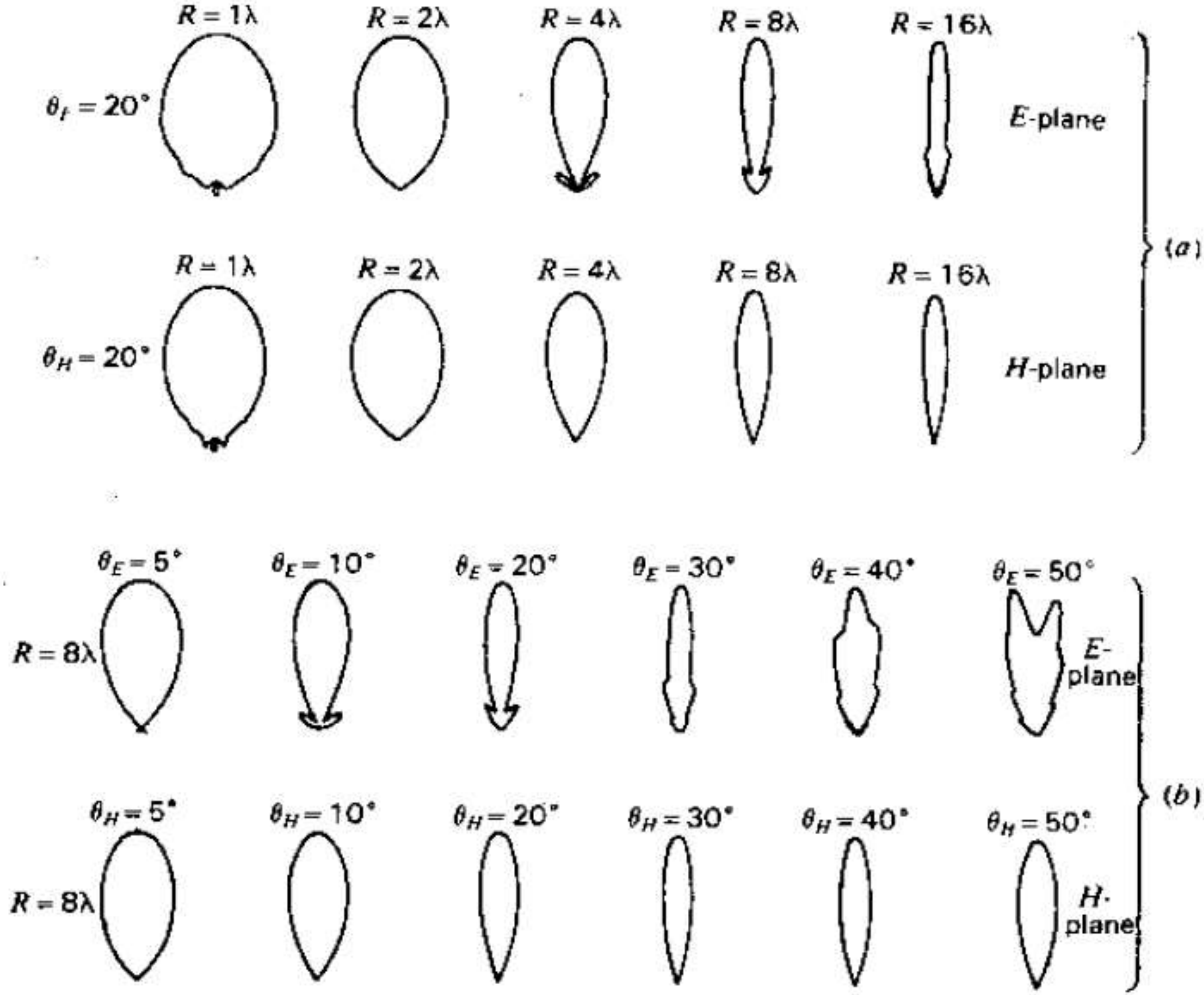


Figure 13-23 Measured E- and H-plane field patterns of rectangular horns as a function of flare angle and horn length. (*After D. R. Rhodes, "An Experimental Investigation of the Radiation Patterns of Electromagnetic Horn Antennas,"* Proc. IRE, **36**, *1101–1105, September 1948.*)

The directivity (or gain, assuming no loss) of a horn antenna can be expressed in terms of its effective aperture. Thus,

$$D = \frac{4\pi A_e}{\lambda^2} = \frac{4\pi \varepsilon_{ap} A_p}{\lambda^2} \tag{1}$$

where A_e = effective aperture, m^2
A_p = physical aperture, m^2
ε_{ap} = aperture efficiency = A_e/A_p
λ = wavelength, m

For a rectangular horn $A_p = a_E a_H$ and for a conical horn $A_p = \pi r^2$, where r = aperture radius. It is assumed that a_E, a_H or r are all at least 1λ. Taking $\varepsilon_{ap} \simeq 0.6$, (1) becomes

$$D \simeq \frac{7.5 A_p}{\lambda^2} \tag{2}$$

or

$$D \simeq 10 \log\left(\frac{7.5 A_p}{\lambda^2}\right) \qquad \text{(dBi)} \tag{3}$$

For a pyramidal (rectangular) horn (3) can also be expressed as

$$D \simeq 10 \log\left(7.5 a_{E\lambda} a_{H\lambda}\right) \tag{4}$$

where $a_{E\lambda}$ = E-plane aperture in λ
$a_{H\lambda}$ = H-plane aperture in λ

Type of aperture	Beam width, deg	
	Between first nulls	Between half-power points
Uniformly illuminated rectangular aperture or linear array	$\dfrac{115}{L_\lambda}$	$\dfrac{51}{L_\lambda}$
Uniformly illuminated circular aperture	$\dfrac{140}{D_\lambda}$	$\dfrac{58}{D_\lambda}$
Optimum E-plane rectangular horn	$\dfrac{115}{a_{E\lambda}}$	$\dfrac{56}{a_{E\lambda}}$
Optimum H-plane rectangular horn	$\dfrac{172}{a_{H\lambda}}$	$\dfrac{67}{a_{H\lambda}}$

† L_λ = length of rectangular aperture or linear array in free-space wavelengths
D_λ = diameter of circular aperture in free-space wavelengths
$a_{E\lambda}$ = aperture in E plane in free-space wavelengths
$a_{H\lambda}$ = aperture in H plane in free-space wavelengths

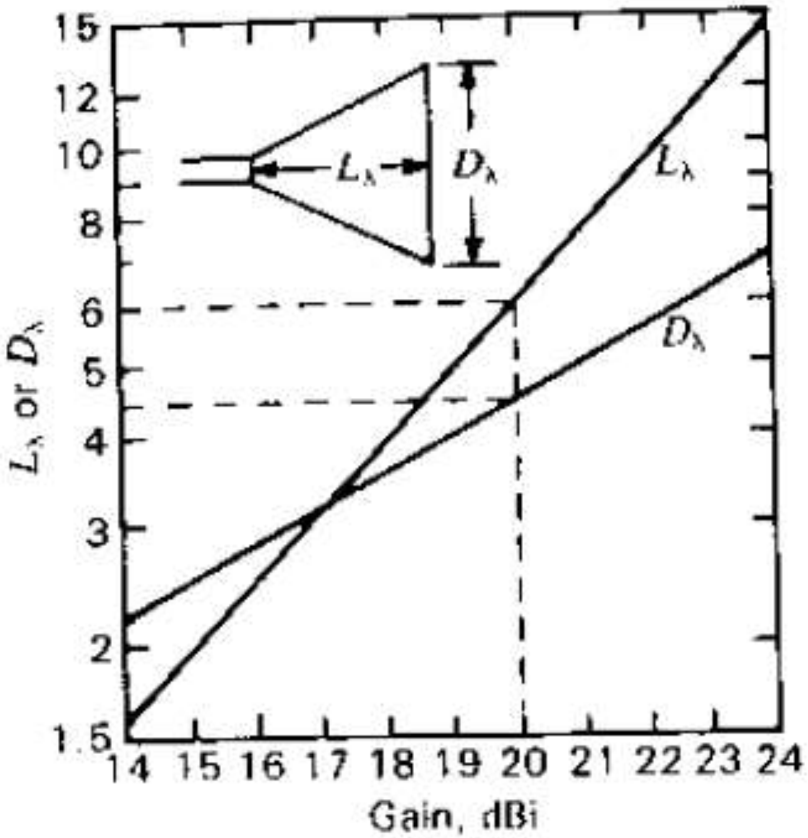


Figure 13-25b Dimensions of conical horn (in wavelengths) versus directivity (or gain, if no loss). Thus, noting the dashed lines, a gain of 20 dBi requires a horn length $L_\lambda = 6.0$ and a diameter $D_\lambda = 4.3$. These (inside) dimensions are close to optimum.

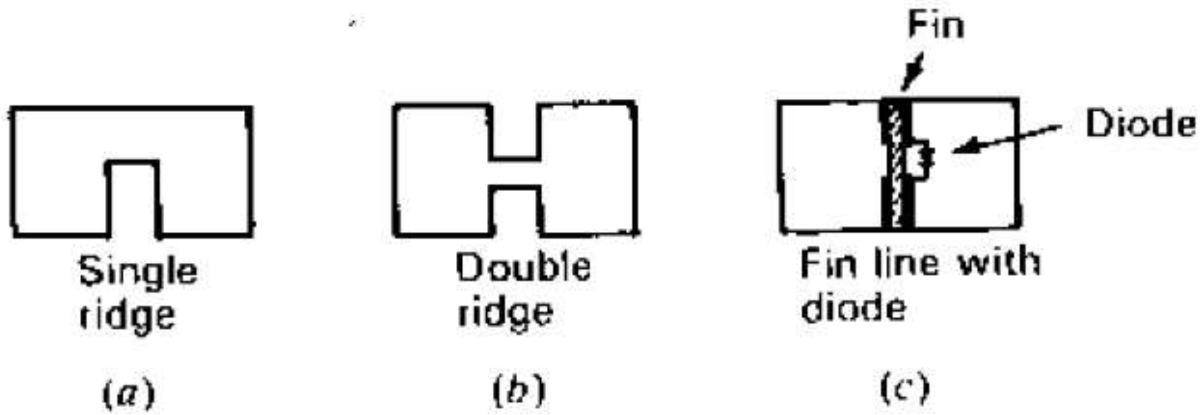


Figure 13-26 Single- and double-ridge rectangular waveguide and fin-line with diode.

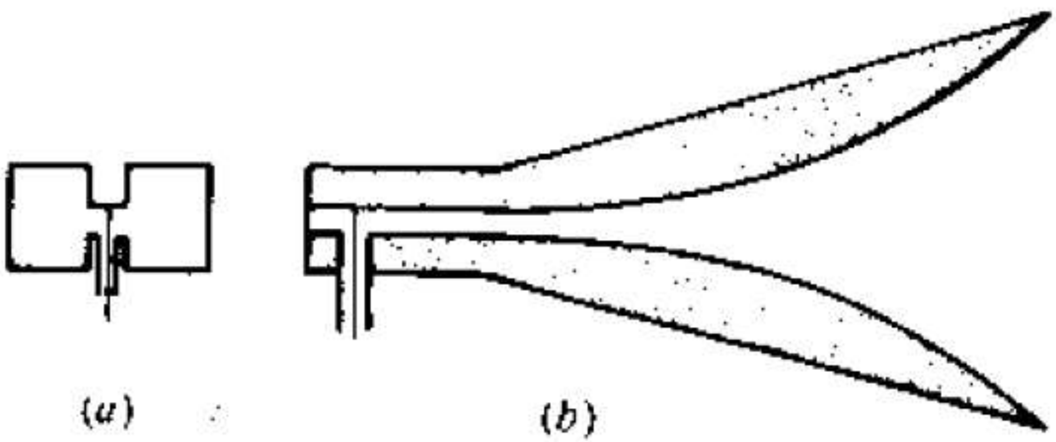


Figure 13-27 Double-ridge horn with coaxial feed. The view at (a) is a cross section at the feed point.

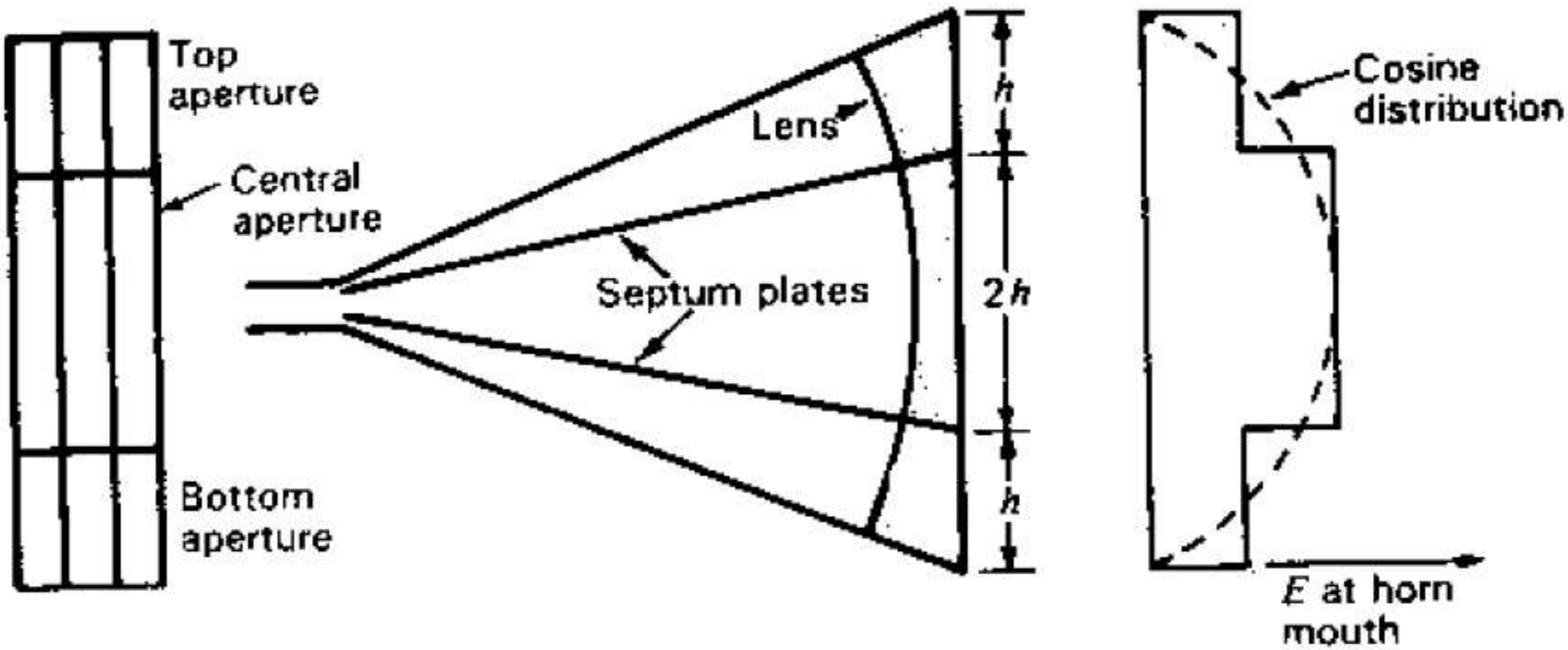


Figure 13-28 Two-septum horn with 1 : 2 : 1 stepped amplitude distribution in field intensity at mouth of horn (approximating a cosine distribution).

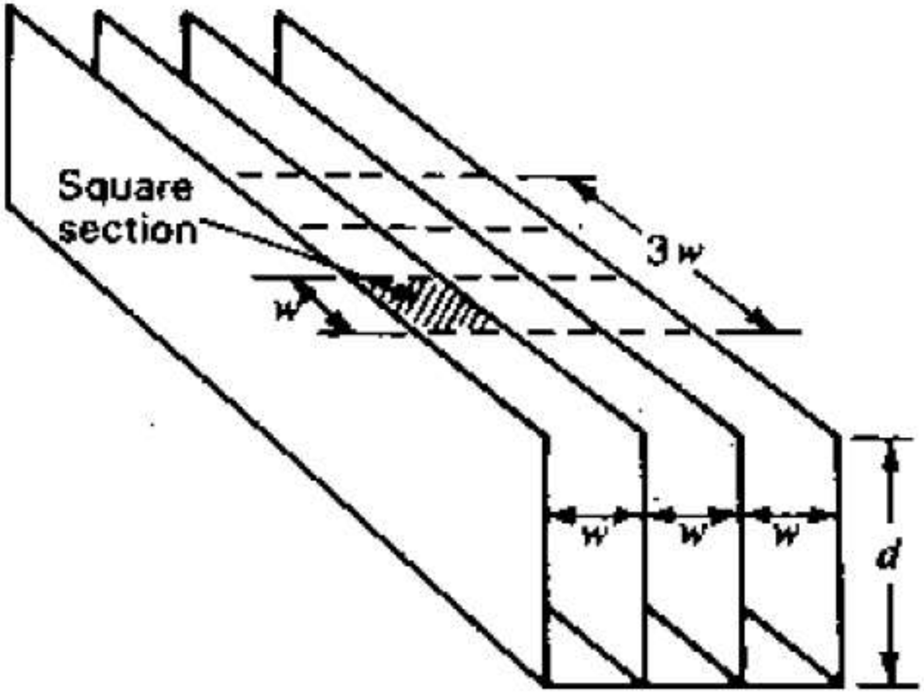


Figure 13-30 Corrugations of width w and depth d.

$$X \simeq 377 \tan\left(\frac{2\pi d}{\lambda}\right) \qquad (\Omega)$$

where d = depth, λ

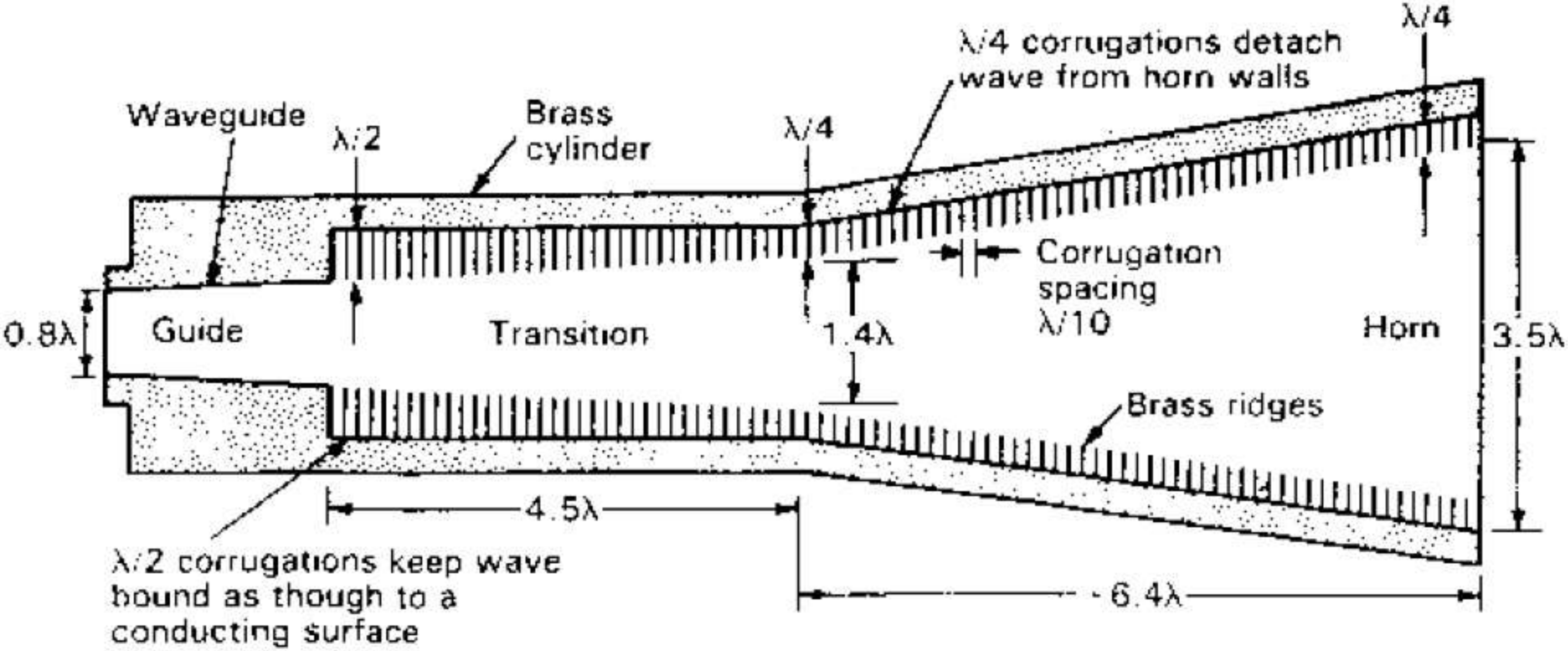


Figure 13-31 Cross section of circular waveguide-fed corrugated horn with corrugated transition. Corrugations with depth of $\lambda/2$ at waveguide act like a conducting surface while corrugations with $\lambda/4$ depth in horn present a high impedance. (*After T. S. Chu* et al., "*Crawford Hill 7-m Millimeter Wave Antenna*," Bell Sys. Tech. J., **57**, *1257–1288, May–June 1978.*)

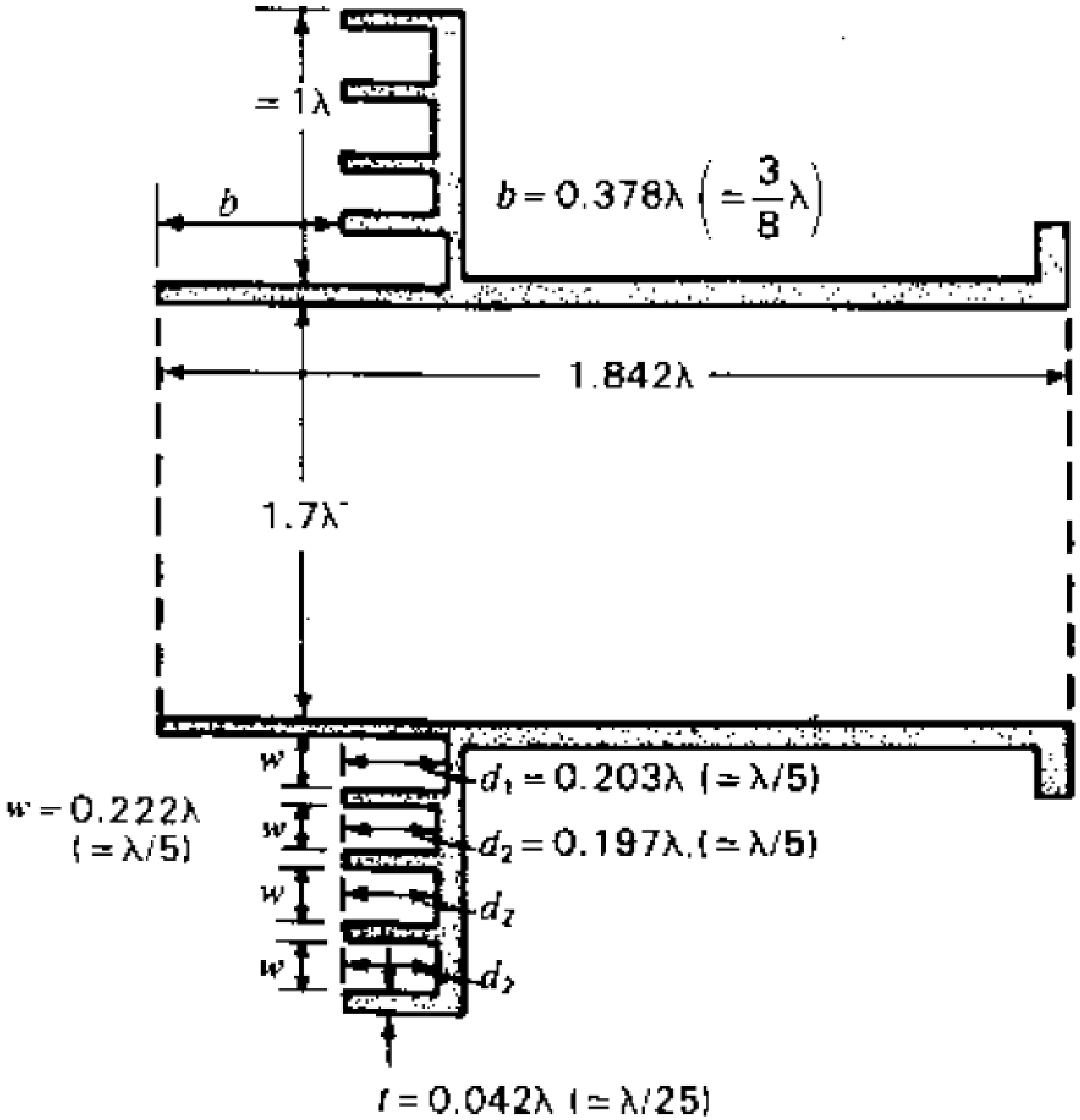


Figure 13-32 Cross section of circular waveguide with flange and 4 chokes for wide-beam-width high-efficiency feed of low F/D parabolic reflectors. (*After R. Wohlleben, H. Mattes and O. Lochner, "Simple, Small, Primary Feed for Large Opening Angles and High Aperture Efficiency,"* Electronics Letters, 8, *19, September 1972.*)

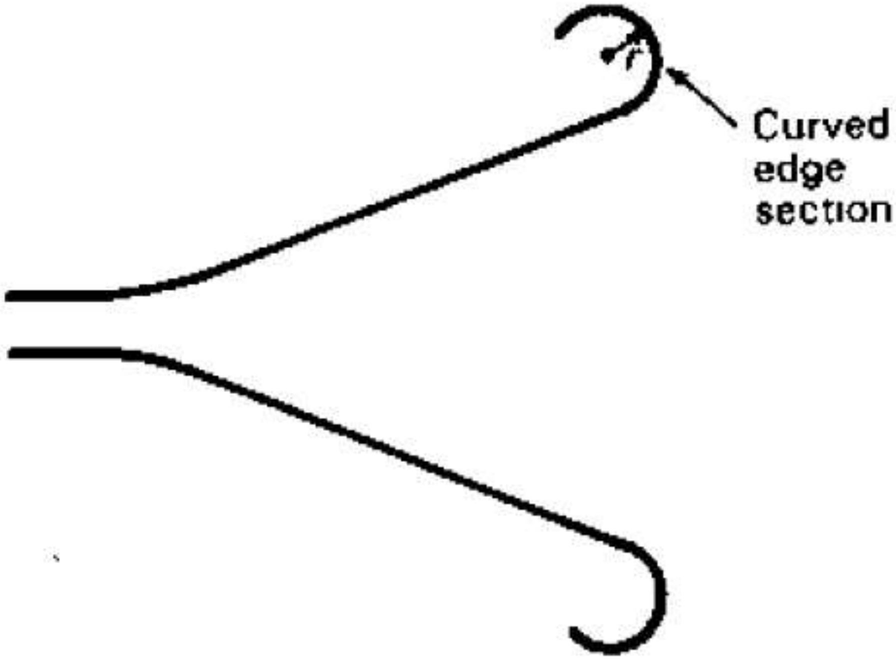


Figure 13-33 Cross section of Burnside and Chuang's aperture-matched horn. The radius of curvature r of the rolled edge should be at least $\lambda/4$.

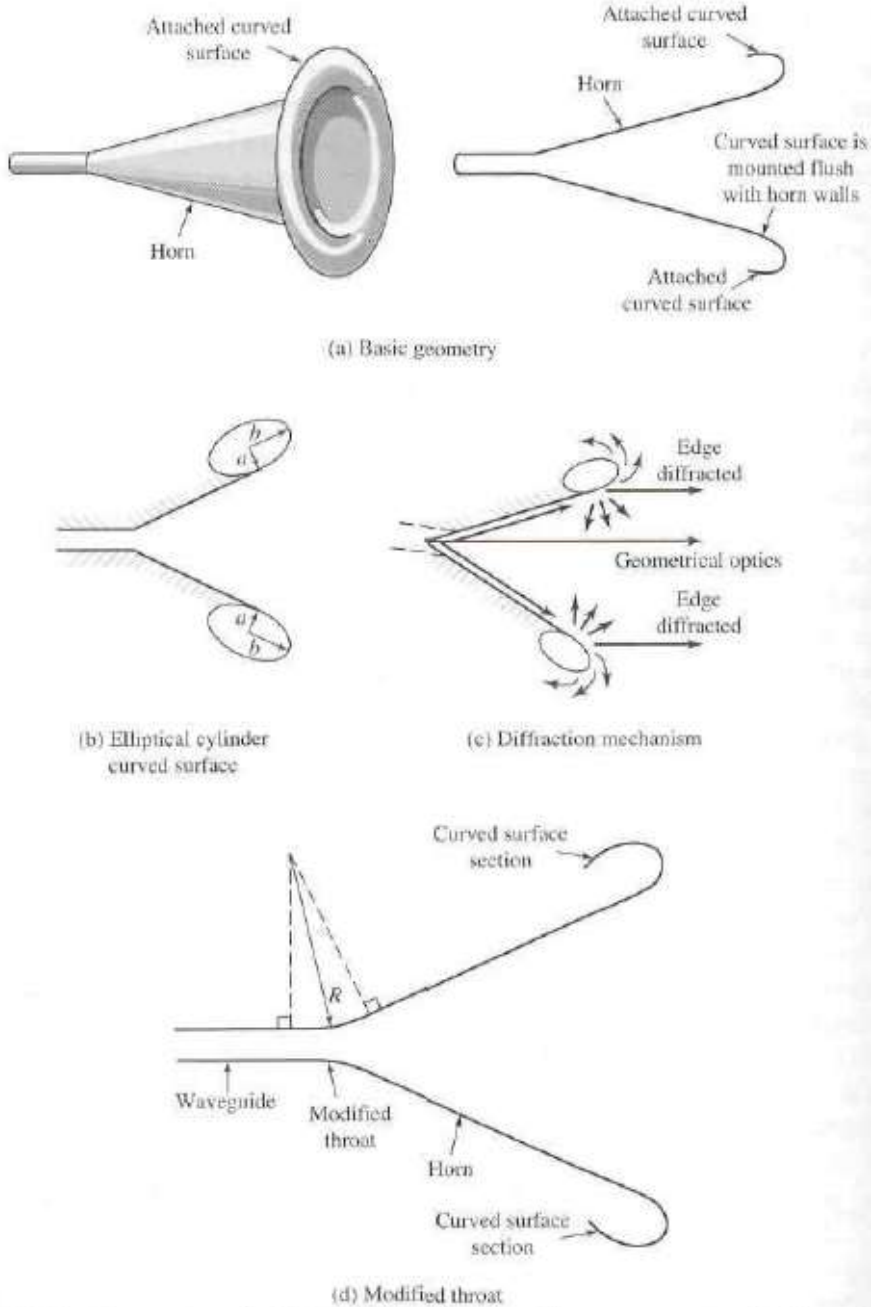


Figure 13.34 Geometry and diffraction mechanism of an aperture-matched horn. (SOURCE: W. D. Burnside and C. W. Chuang, "An Aperture-Matched Horn Design," *IEEE Trans. Antennas Propagat.*, Vol. AP-30, No. 4, pp. 790–796, July 1982. © (1982) IEEE)

Horn Antenna

- The horn antenna is used in the transmission and reception of RF microwave signals, and the antenna is normally used in conjunction with waveguide feeds.

- A horn antenna form of antenna that consists of a flared waveguide which is shaped like a horn and it has the effect that it enables a transition between the waveguide and free space and it also directs radio waves in a beam.

- Horn antenna history

- One of the first recorder instances of the appearance of the horn antenna was in 1897 when one was constructed in by an Indian radio researcher named Jagadish Chandra Bose who was experimenting with wireless signals we now call microwaves.

- Then around 1936 much experimental research was undertaken by Southworth and Barrow. Later this was followed by theoretical analysis of the horn antenna in 1939 by Barrow and Chu.

- The onset of World War 2 provided a significant level of impetus to the development of the horn antenna as it was suitable for use as a feed horn in radar antennas that were starting to be used.

- Then in1962 Kay invented the corrugated horn which has become widely used as a feed horn for microwave antennas such as satellite dishes and radio telescopes.

Basic Horn Antenna Concept

- The horn antenna may be considered as an RF transformer or impedance match between the waveguide feeder and free space which has an impedance of 377 ohms.

- By having a tapered or having a flared end to the waveguide the horn antenna is formed and this enables the impedance to be matched.

- Although the waveguide will radiate without a horn antenna, this provides a far more efficient match.

- **Microwave horn antenna**
- In addition to the improved match provided by the horn antenna, it also helps suppress signals travelling via unwanted modes in the waveguide from being radiated.

- However the main advantage of the horn antenna is that it provides a significant level of directivity and gain. For greater levels of gain the horn antenna should have a large aperture.

- Also to achieve the maximum gain for a given aperture size, the taper should be long so that the phase of the wave-front is as nearly constant as possible across the aperture.

- However there comes a point where to provide even small increases in gain, the increase in length becomes too large to make it sensible. Thus gain levels are a balance between aperture size and length. However gain levels for a horn antenna may be up to 20 dB in some instances.

Horn antenna types

- There are several types of horn antenna:
- ***Pyramid horn antenna*** As the name suggests, the pyramid horn antenna takes on a rectangular shape - the cross section through the antenna is rectangular, as is the end of the antenna. It is normally used with rectangular waveguide.
- ***Sectoral horn antenna:*** This form of horn antenna is one in which only one pair of sides flared whist the other remains parallel. This form of configuration produces a fan-shaped beam, which is narrow in the plane of the flared sides, but wide in the plane of the narrow sides.
  - ***E-plane horn antenna:*** This form of antenna is one that is flared in the direction of the electric or E-field in the waveguide.
  - ***H plan horn antenna :*** This form of antenna is one that is flared in the direction of the electric or H-field in the waveguide.
- ***Conical horn antenna*** Again, as the name indicates, the conical horn antenna has a circular cross section and end ito it. It is normally used with circular waveguide and is seen less frequently than the rectangular version.
- ***Exponential horn antenna*** This form of horn antenna is also called a scalar horn antenna and it is one that has curved sides. The separation of the sides increases as an exponential function of length. The antenna can come as either a pyramidal or conical cross sections. The advantage of exponential horns is that they have a minimum level of internal reflections, and almost constant impedance and other characteristics over a wide frequency range. They are used in applications requiring high performance, such as feed horns for communication satellite antennas and radio telescopes.
- ***Corrugated horn:*** The corrugated horn antenna has parallel slots or grooves along the inside surface of the horn, transverse to the axis. These corrugations are small when compared to the wavelength. Corrugated horns have several advantages including a wider bandwidth, and smaller side-lobes that other types. The corrugated horn provides a pattern that is nearly symmetrical, with the E and H plane beam-widths being nearly the same. As a result corrugated horn antennas are widely used as feed horns for satellite dishes and radio telescopes.

Advantages and Disadvantages

- Advantages
  - It can operate over broad bandwidth-The horn antenna possesses no resonant elements and therefore it is able to operate over a wide bandwidth.
  - Significant level of directivity and gain
  - Low standing wave ratio
  - Simple construction and adjustment-The horn antenna consists simply of a flared horn. As a result they are relatively easy to construct.
  - Superior power handling capability than other antennas
  - Easy interface to waveguide - By the very nature of their shape, these antennas are very easy to interface to waveguide, although they can also be designed with a transition so that standard coaxial feeder can also be used.

- Disadvantages
  - Its gain is subjected to major fluctuations particularly at low frequencies.
  - VSWR in lower frequency range is not favorable with values between 2 and 5.

Applications

- Used in remote sensing satellites, communication satellites, geographic information and weather satellite.

- Universal standard for calibration and gain measurements of other antennas due to its very wide bandwidth and its performance varies little over a wide frequency range.

- Used as feed element for parabolic reflector antennas
  - Radio astronomy
  - Satellite tracking
  - Communication dishes

- Short range radar systems in speed enforcement cameras.

Reflector Antennas

- Introduction
- Flat sheet and corner reflectors
- Paraboloidal reflectors
  - geometry
  - pattern characteristics
- Feed Methods
- Reflector Types.

Neelima Koppala

A parabolic satellite communicationsantenna at Erdfunkstelle Raisting, the biggest facility for satellite communication in the world, in Raisting, Bavaria, Germany. It has a Cassegrain type feed.

Wire grid-type parabolic antenna used for MMDS data link at a frequencyof 2.5-2.7 GHz. It is fed by a verticaldipole under the small aluminum reflector on the boom. It radiatesvertically polarized microwaves.

Dish parabolic antennas

Shrouded microwave relay dishes on a communications tower in Australia.

A satellite television dish, an example of an offset fed dish.

Cassegrain satellite communication antenna in Sweden.

Offset Gregorian antenna used in the Allen Telescope Array, a radio telescope at the University of California at Berkeley, USA.

Shaped-beam parabolic antennas

Vertical "orange peel" antenna for military height finder radar, Germany.

Early cylindrical parabolic antenna, 1931, Nauen, Germany.

Air traffic control radar antenna, near Hannover, Germany.

ASR-9 Airport surveillance radar antenna.

"Orange peel" antenna for air search radar, Finland.

Introduction

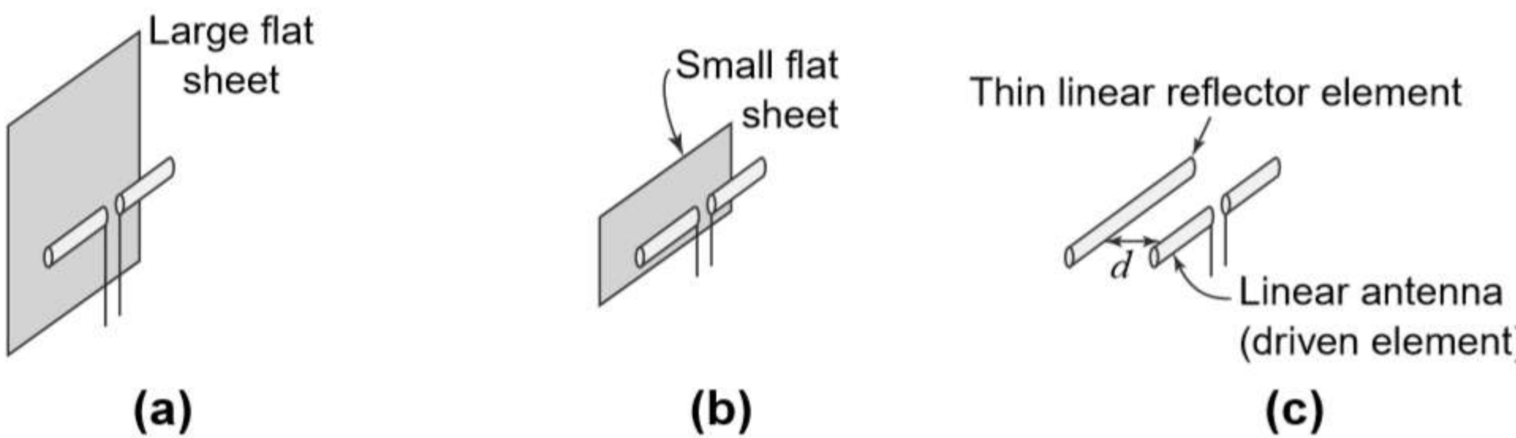
- Reflectors are widely used to modify the radiation pattern of a radiating element.

- The backward radiation from an antenna may be eliminated with a plane sheet reflector of large enough dimensions. In the more general case, a beam of predetermined characteristics may be produced by means of a large, suitably shaped, and illuminated reflector surface.

- The characteristics of antennas with sheet reflectors or their equivalent are considered.

- The arrangement in Fig. 1a has a large, flat sheet reflector near a linear dipole antenna to reduce the backward radiation (to the left in the figure). With small spacings between the antenna and sheet this arrangement also yields a substantial gain in the forward radiation.

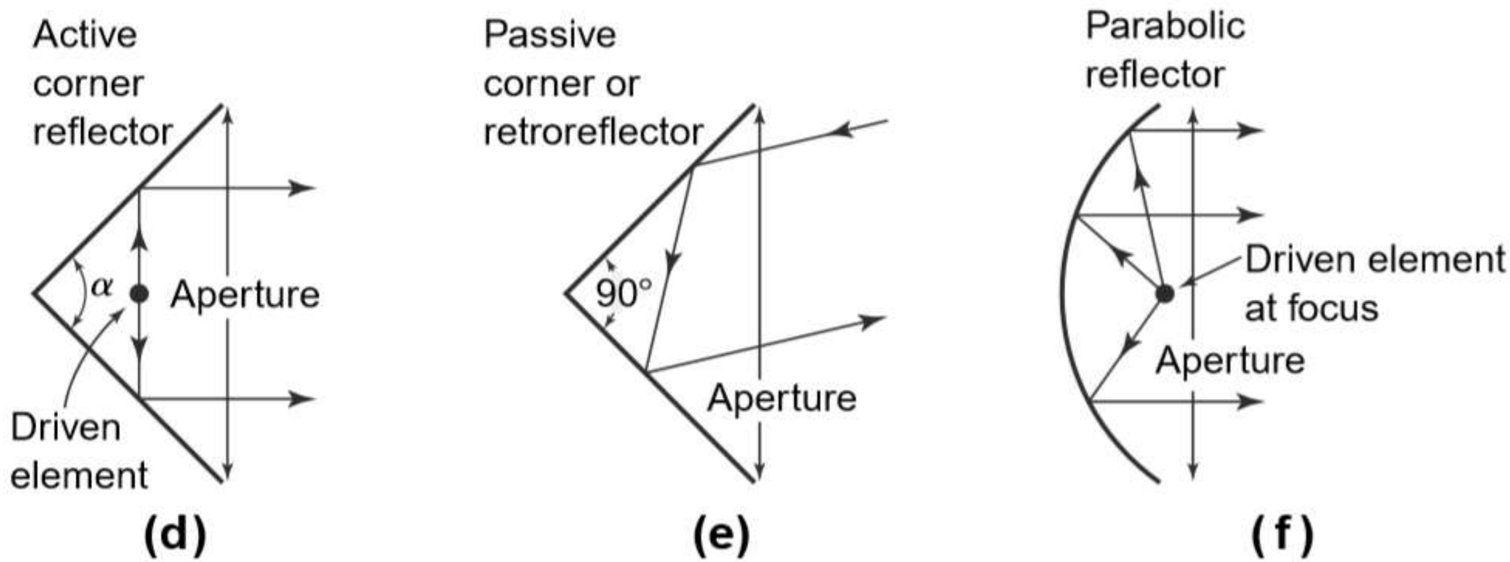
Contd.,



- The desirable properties of the sheet reflector may be largely preserved with the reflector reduced in size as in Fig. 1b and even in the limiting case of Fig. 1c.

- Here the sheet has degenerated into a thin reflector element. Whereas the properties of the large sheet are relatively insensitive to small frequency changes, the thin reflector element is highly sensitive to frequency changes.

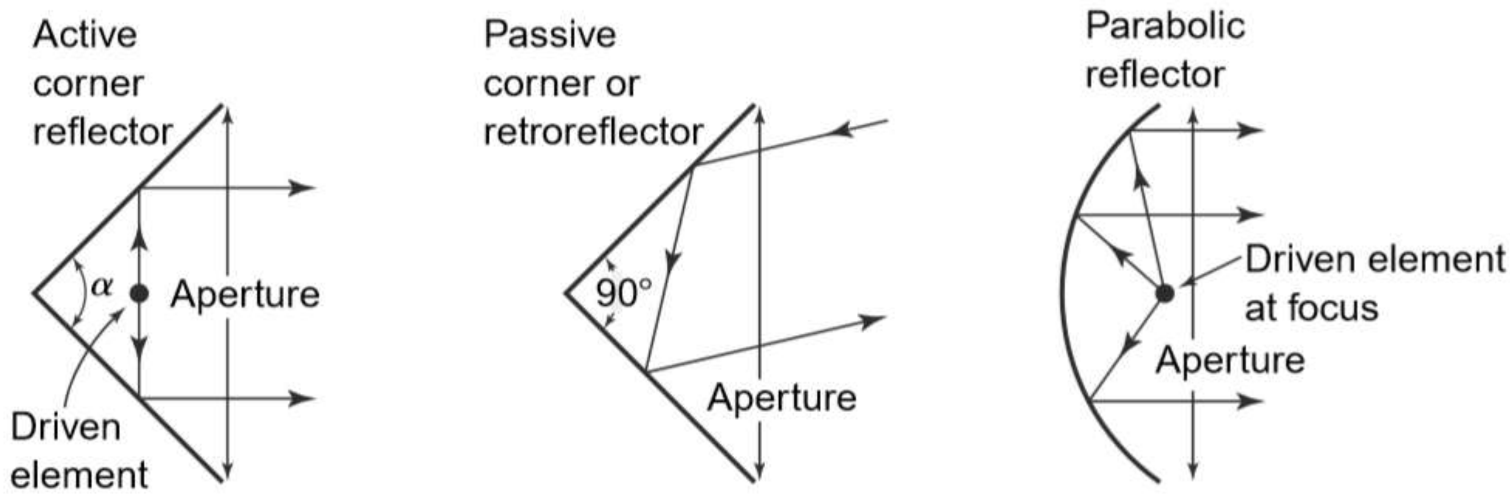
Contd.,



- The case of a $\lambda/2$ antenna with parasitic reflector element is treated with two flat sheets intersecting at an angle α (<180◦) as in Fig. 1d, a sharper radiation pattern than from a flat sheet reflector (α = 180◦) can be obtained, called as an active corner reflector antenna, is most practical where apertures of 1 or 2λ are of convenient size.

- A corner reflector without an exciting antenna can be used as a passive reflector or target for radar waves. In this application the aperture may be many wavelengths, and the corner angle is always 90◦.

Contd.,

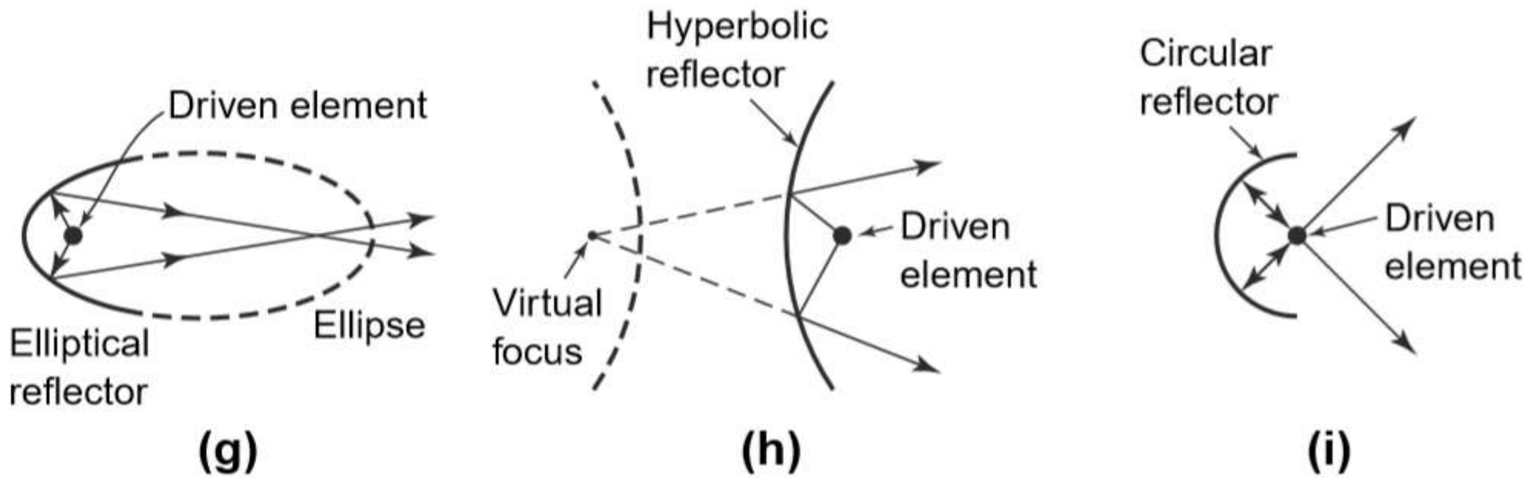


(d) (e) (f)

- Reflectors with this angle have the property that an incident wave is reflected back toward its source as in Fig.1e, the corner acting as a retroreflector. When antennas have apertures of many wavelengths, parabolic reflectors can be used to provide highly directional antennas.

- A parabolic reflector antenna is shown in Fig. 1f. The parabola reflects the waves originating from a source at the focus into a parallel beam, the parabola transforming the curved wave front from the feed antenna at the focus into a plane wave front.

Contd.,



- Many other shapes of reflectors can be employed for special applications. For instance, with an antenna at one focus, the elliptical reflector (Fig.1g) produces a diverging beam with all reflected waves passing through the second focus of the ellipse.

- Examples of reflectors of other shapes are the hyperbolic (Stavis-1) and the circular reflectors (Ashmead-1) shown in Figs. 1h and i.

Flat Sheet Reflectors

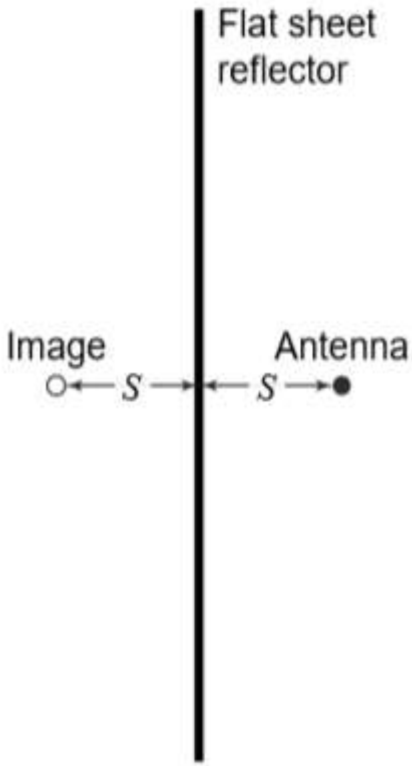


Figure 9–2
Antenna with flat sheet reflector.

- The problem of an antenna at a distance S from a perfectly conducting plane sheet reflector of infinite extent is readily handled by the method of images (Brown-1).

- In this method the reflector is replaced by an image of the antenna at a distance 2S from the antenna, as in Fig. 2, this is identical with a horizontal antenna above ground. If the antenna is a $\lambda/2$ dipole this in turn reduces to the problem of the W8JK antenna.

Contd.,

- Assuming zero reflector losses, the gain in field intensity of a $\lambda/2$ dipole antenna at a distance S from an infinite plane reflector is

$$G_f(\phi) = 2\sqrt{\frac{R_{11} + R_L}{R_{11} + R_L - R_{12}}}\,|\sin(S_r \cos\phi)|$$

where $S_r = 2\pi S/\lambda$.

- The gain in (1) is expressed relative to a $\lambda/2$ antenna in free space with the same power input.

Contd.,

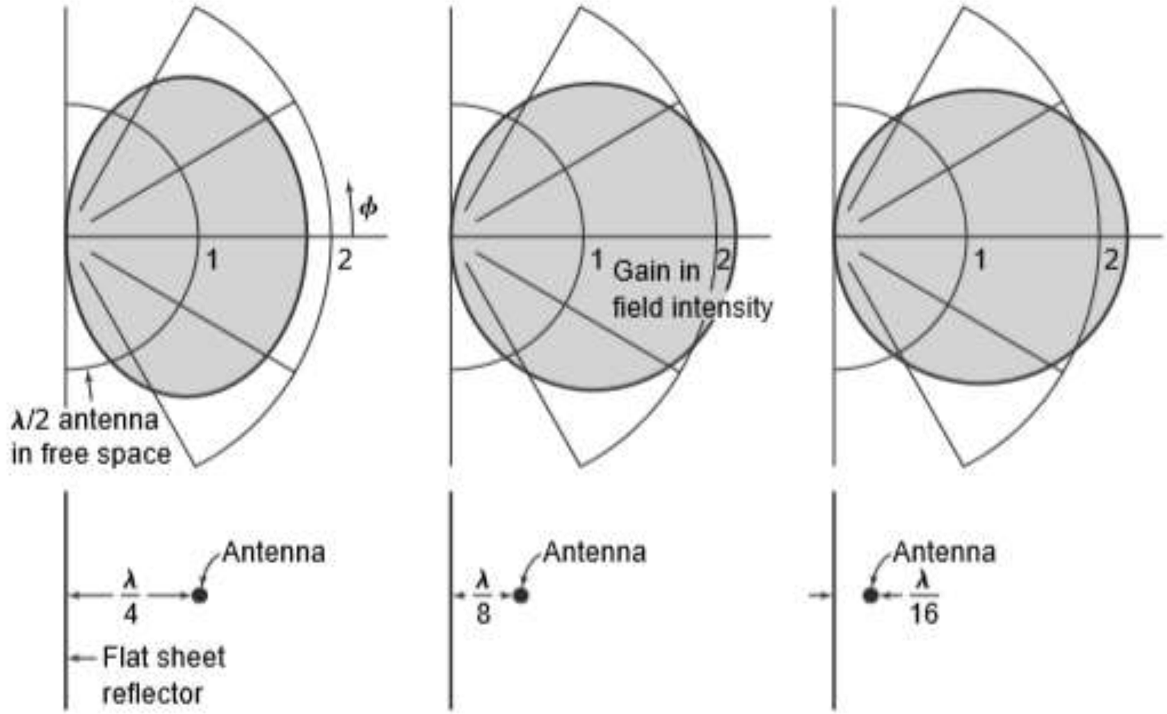


Figure 9–3 Field patterns of a $\lambda/2$ antenna at spacings of $\frac{1}{4}$, $\frac{1}{8}$ and $\frac{1}{16}\lambda$ from an infinite flat sheet reflector. Patterns give gain in field intensity over a $\lambda/2$ antenna in free space with same power input. For $\lambda/8$ spacing, the gain is 2.2 (= 6.7 dB = 8.9 dBi).

These patterns are calculated from (1) for the case where RL = 0.

Contd.,

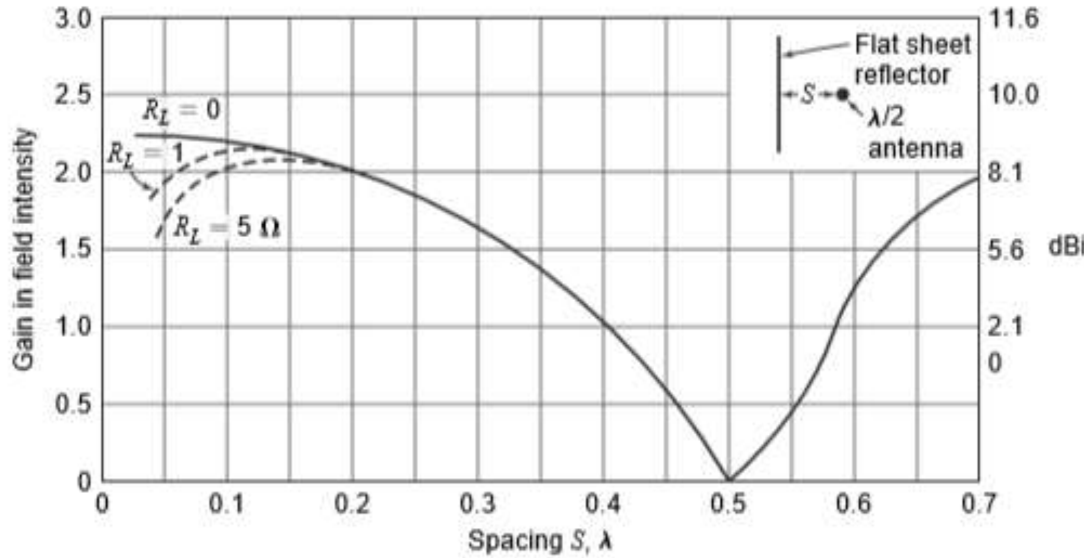


Figure 9–4 Gain in field intensity of $\lambda/2$ dipole antenna at distance S from flat sheet reflector. Gain is relative to $\lambda/2$ dipole antenna in free space with the same power input. Gain in dBi is also shown. Gain is in direction $\phi = 0$ and is shown for assumed loss resistances $R_L = 0.1$ and 5 Ω.

- The gain as a function of the spacing S is presented in Fig. 9–4 for assumed antenna loss resistances RL = 0, 1 and 5 .

Contd.,

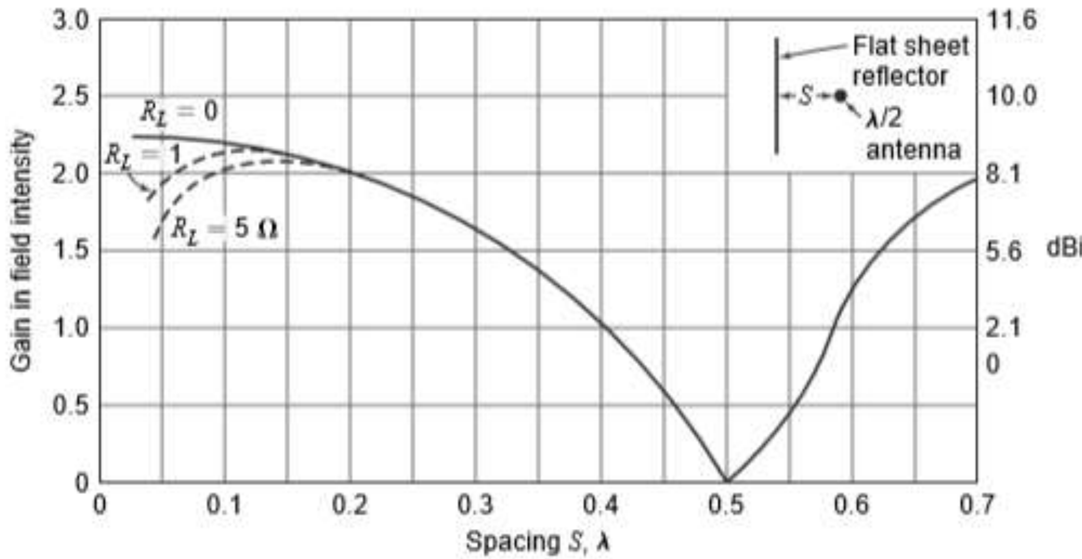


Figure 9–4 Gain in field intensity of λ/2 dipole antenna at distance S from flat sheet reflector. Gain is relative to λ/2 dipole antenna in free space with the same power input. Gain in dBi is also shown. Gain is in direction $\phi = 0$ and is shown for assumed loss resistances $R_L = 0.1$ and 5 Ω.

- These curves are calculated from (1) for φ = 0. It is apparent that very small spacings can be used effectively provided that losses are small. However, the bandwidth is narrow for small spacings. With wide spacings the gain is less, but the bandwidth is larger. Assuming an antenna loss resistance of 1 , a spacing of 0.125λ yields the maximum gain.

Contd.,

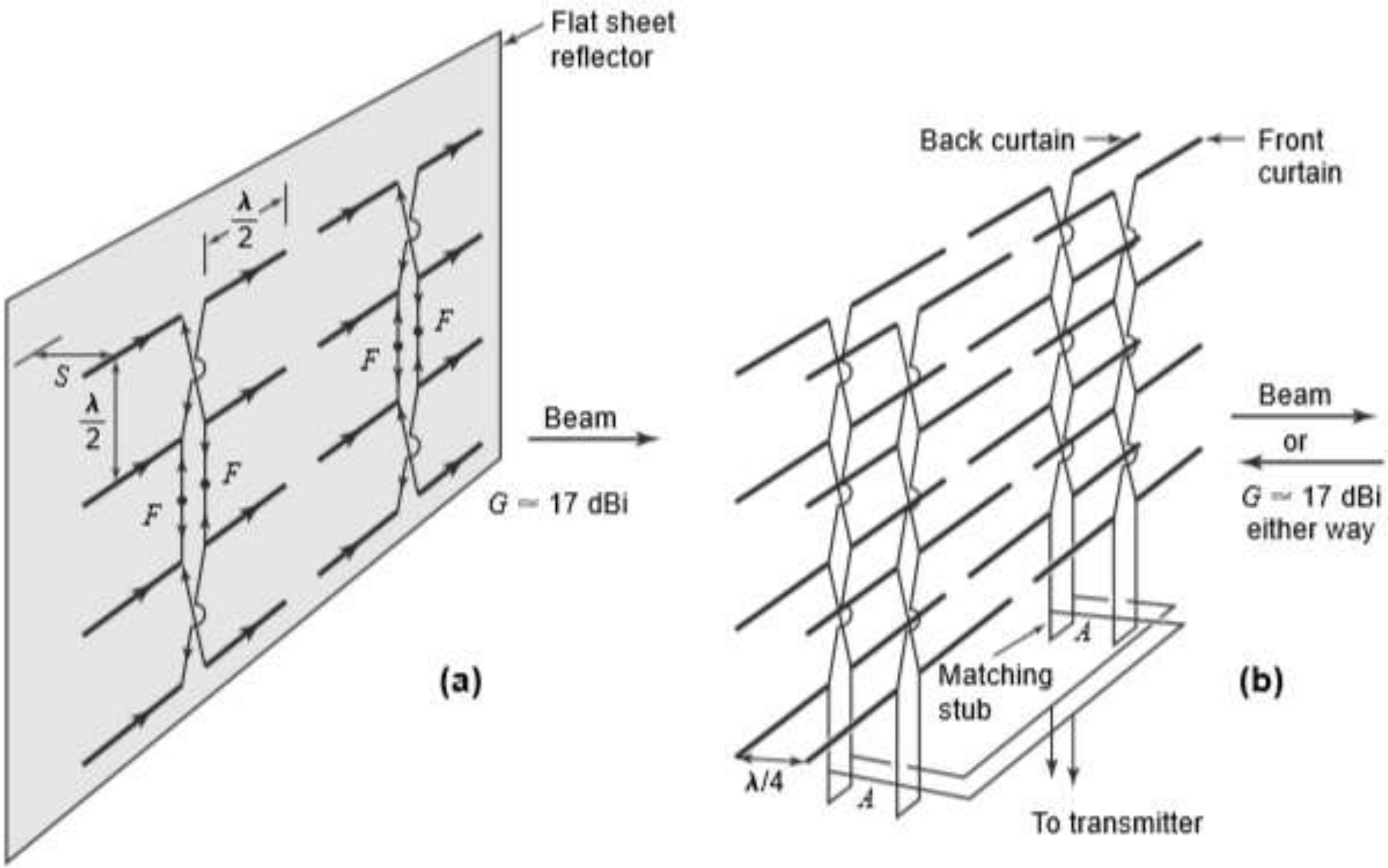


Figure 9–5 (*a*) Array of 16 $\lambda/2$ dipoles with flat-sheet reflector (billboard antenna). (*b*) Flat-sheet reflector replaced by a back curtain of 16 $\lambda/2$ dipoles for reduced wind resistance. Also by reversing the transmission lines at *A*, the beam direction can be reversed.

Contd.,

- A large flat sheet reflector can convert a bidirectional antenna array into a unidirectional system. An example is shown in Fig. 9–5a.

- Here a broadside array of 16 in-phase $\lambda/2$ elements spaced $\lambda/2$ apart is backed up by a large sheet reflector so that a unidirectional beam is produced.

- The feed system for the array is indicated, equal in-phase voltages being applied at the 2 pairs of terminals F–F.

- If the edges of the sheet extend some distance beyond the array, the assumption that the flat sheet is infinite in extent is a good first approximation.

Contd.,

- The choice of the spacing S between the array and the sheet usually involves a compromise between gain and bandwidth.

- If a spacing of $\lambda/8$ is used, the radiation resistance of the elements of a large array remains about the same as with no reflector present (Wheeler-1).

- This spacing also has the advantage over wider spacings of reduced interaction between elements.

- On the other hand, a spacing such as $\lambda/4$ provides a greater bandwidth, and the precise value of S is less critical in its effect on the element impedance.

Contd.,

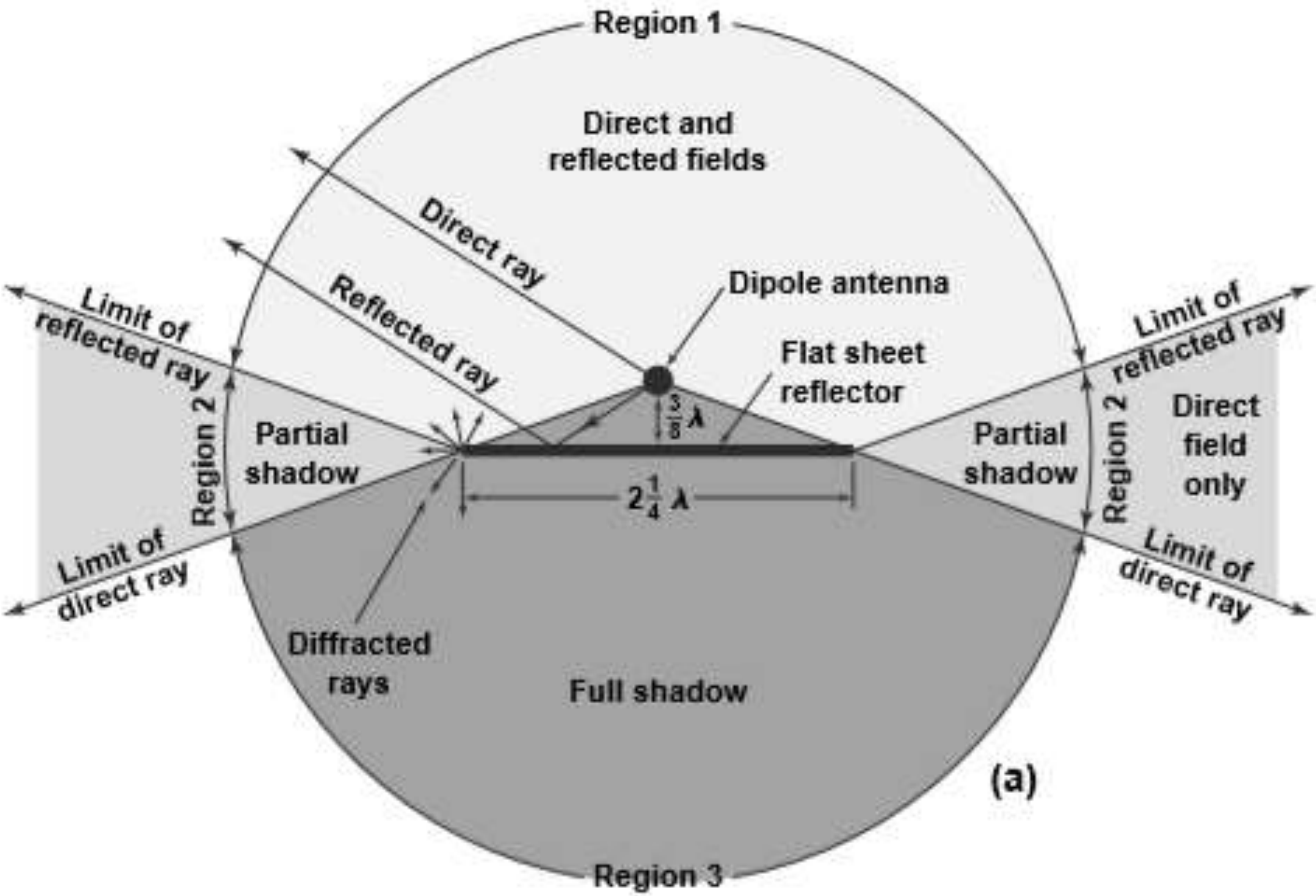
- Many shortwave broadcast stations operating at 15 to 50 m wavelengths and beaming worldwide use large curtain arrays; as in Fig. 9–5b, supported by tall towers.

- The array in Fig. 9–5b has a gain of about 17 dBi.

- Adding another set of curtain arrays (both front and back) alongside the array of Fig. 9–5b for a total of 64 dipoles increases the gain to about 20 dBi. Doubling again to 128 dipoles, the gain is about 23 dBi and doubling again to 256 the gain is about 26 dBi.

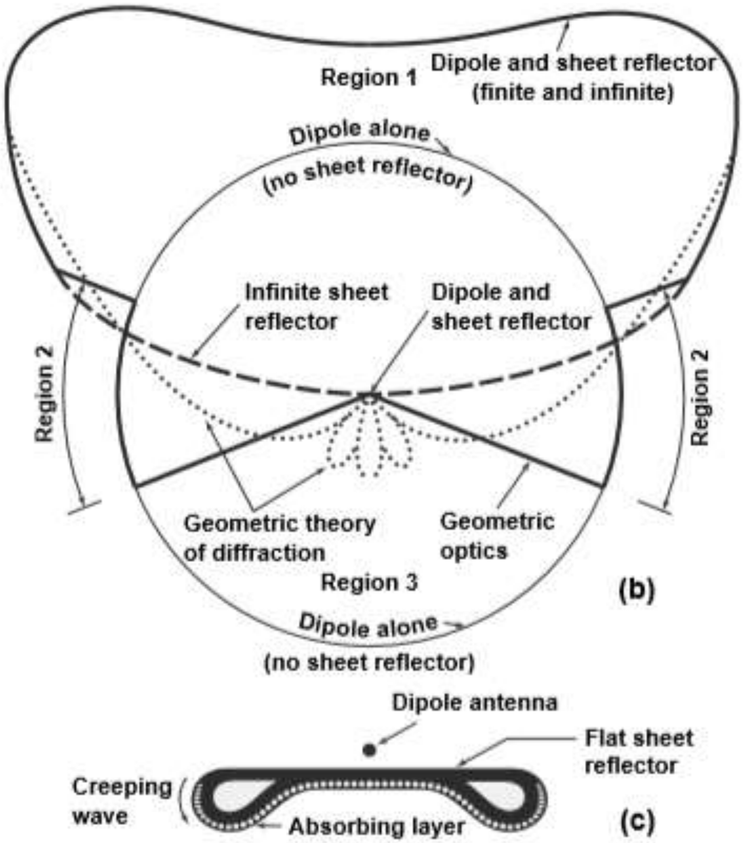
- Doubling once more to 512 dipoles the gain is about 29 dBi.

Contd.,



Region 1
Direct and reflected fields
Direct ray
Reflected ray
Dipole antenna
Flat sheet reflector
Limit of reflected ray
Limit of reflected ray
Region 2
Partial shadow
Partial shadow
Region 2
Direct field only
$\frac{3}{8}\lambda$
$2\frac{1}{4}\lambda$
Limit of direct ray
Limit of direct ray
Diffracted rays
Full shadow
Region 3
(a)

Contd.,



The Geometric Theory of Diffraction (GTD) in a nutshell.

Figure 9–6 (*a*) Dipole antenna with 2.25λ flat sheet reflector with 3 regions of radiation according to geometric optics. (*b*) Field pattern of dipole and sheet reflector according to geometric optics (heavy solid line) and according to geometric theory of diffraction (dotted line). The solid circle indicates the field from the dipole alone (in free space) and the dashed line gives the pattern for dipole with an infinite sheet reflector. (*c*) Modification of edges of sheet to reduce diffracted back-side radiation (in region 3).

Contd.,

- When the reflecting sheet is reduced in size, the analysis is less simple. The situation is shown in Fig. 9–6a. There are 3 principal angular regions:

- Region 1 (above or in front of the sheet). In this region the radiated field is given by the resultant of the direct field of the dipole and the reflected field from the sheet.

- Region 2 (above and below at the sides of the sheet). In this region there is only the direct field from the dipole. This region is in the shadow of the reflected field.

- Region 3 (below or behind the sheet). In this region the sheet acts as a shield, producing a full shadow (no direct or reflected fields, only diffracted fields).

Contd.,

- If the sheet is 1 or 2λ in width and the dipole is close to it, image theory accounts adequately for the radiation pattern in region 1.

- In region 2, the distant field is dominated by the direct ray from the dipole.

- In the full shadow behind the sheet (region 3) the Geometrical Theory of Diffraction (GTD) must be used.1

- The pattern in this region is effectively that of 2 weak line sources, one along each edge.

Contd.,

- The fields in the 3 regions are shown in Fig. 9–6b for the case of the sheet width dimension D = 2.25λ and the dipole spacing d = 3λ/8.

- It is assumed that the sheet is very long perpendicular to the page (>>D).

- Narrower reflecting sheets result in more radiation into region 3 but this diffracted radiation can be minimized by using a rolled edge (radius of curvature > λ/4) and absorbing material, as suggested in Fig.9–6c.

Contd.,

- Corner Reflectors
- John Kraus In 1938, while analyzing the radiation from a dipole parallel to and closely spaced from a flat reflecting sheet, realized that when the sheet is replaced by its image, the dipole and its image form a W8JK array.

- When the flat sheet (180° included angle) is folded into a square (90°) corner the theory calls for 3 images, and my calculations showed correspondingly higher gain.

- Thus, the corner reflector developed as an extension of my analysis of the W8JK array. I immediately constructed several corner reflectors to obtain experimental confirmation.

- He tried parallel-wire grid reflectors, modifying both the spacing and length of the reflector wires to determine the limiting dimensions required.

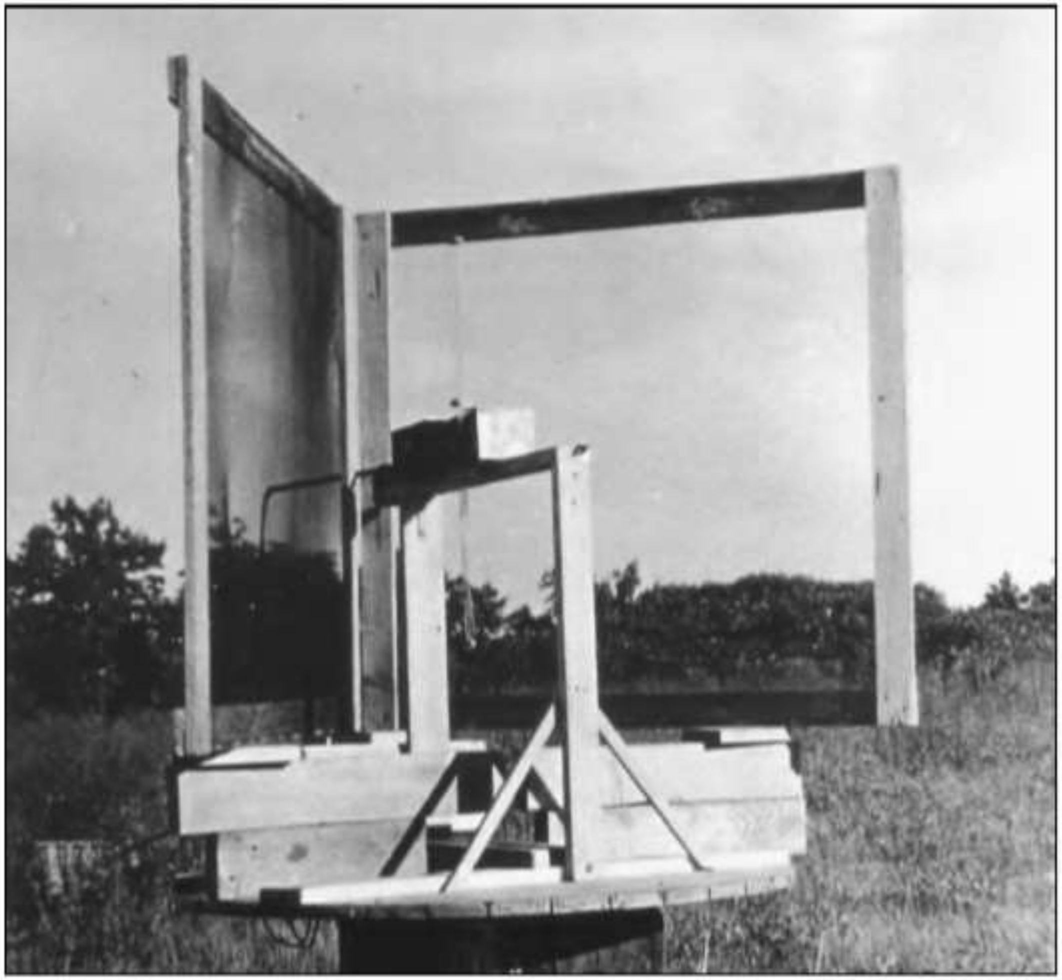


Figure 9–7 90-degree or square-corner reflector on turntable for measurements at $\lambda = 1$ m outside my home in 1939. The reflector is made of copper window screen. The screen reflector is 70 cm square.

Contd.,

- Figure 9–7 shows a 90◦ corner for λ = 1 m operation I built in 1938 with patterns measured by rotating the antenna on a turntable.

Contd.,

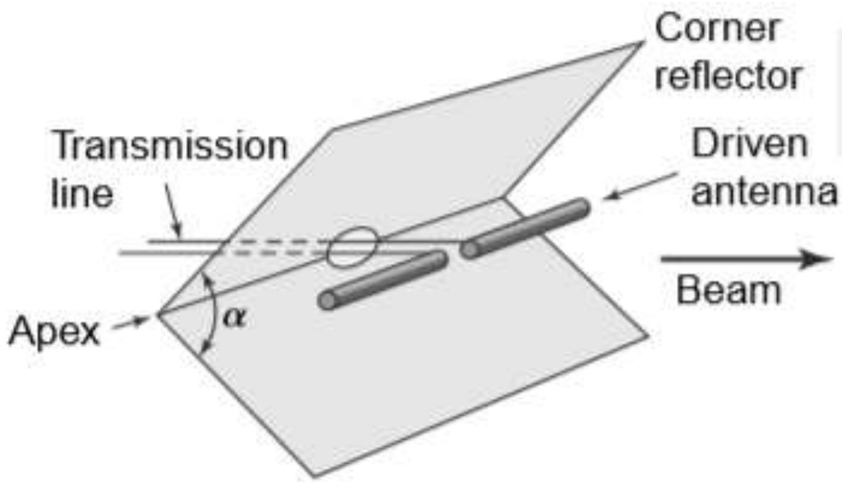


Figure 9–8 Corner reflector antenna.

- Two flat reflecting sheets intersecting at an angle or corner as in Fig. 9–8 form an effective directional antenna.

- When the corner angle α = 90◦, the sheets intersect at right angles, forming a square-corner reflector.

Contd.,

- Corner angles both greater or less than 90° can be used although there are practical disadvantages to angles much less than 90°. A corner reflector with $\alpha = 180^\circ$ is equivalent to a flat sheet reflector and may be considered as a limiting case of the corner reflector.

- Assuming perfectly conducting reflecting sheets of infinite extent, the method of images can be applied to analyze the corner reflector antenna for angles $\alpha = 180^\circ/n$, where n is any positive integer.

- This method of handling corners is well known in electrostatics (Jeans-1). Corner angles of 180° (flat sheet), 90°,60°, etc., can be treated in this way.

- Corner reflectors of intermediate angle cannot be determined by this method but can be interpolated approximately from the others.

Contd.,

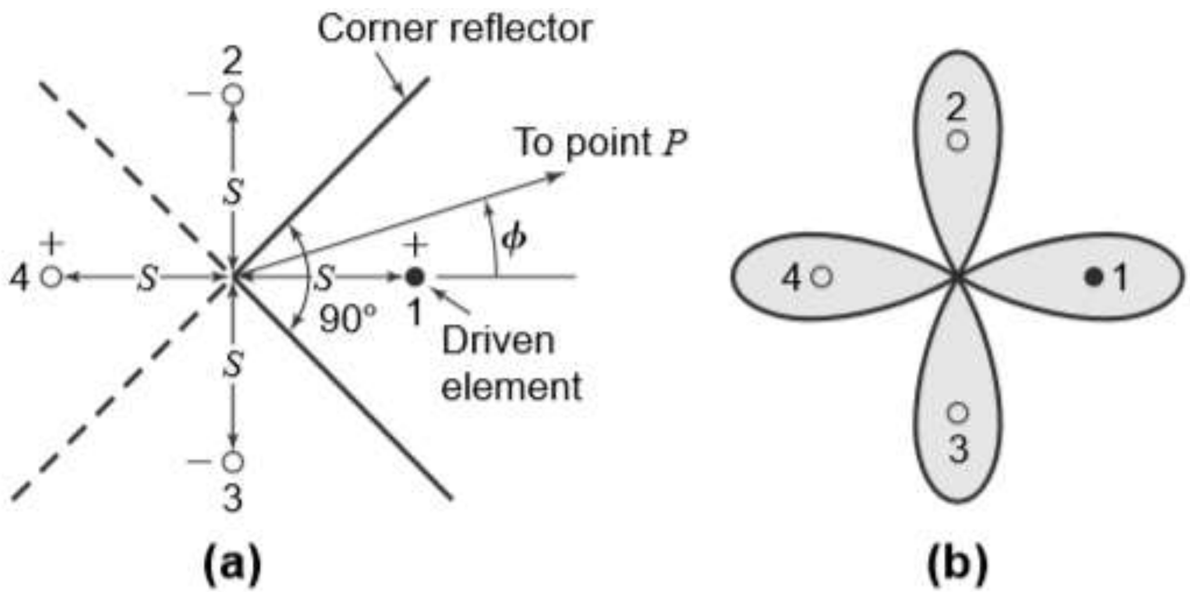


Figure 9–9 Square-corner reflector with images used in analysis (*a*) and 4-lobed pattern of driven element and images (*b*).

- In the analysis of the 90◦ corner reflector there are 3 image elements, 2, 3 and 4, located as shown in Fig. 9–9a.

- The driven antenna 1 and the 3 images have currents of equal magnitude. The phase of the currents in 1 and 4 is the same.

Contd.,

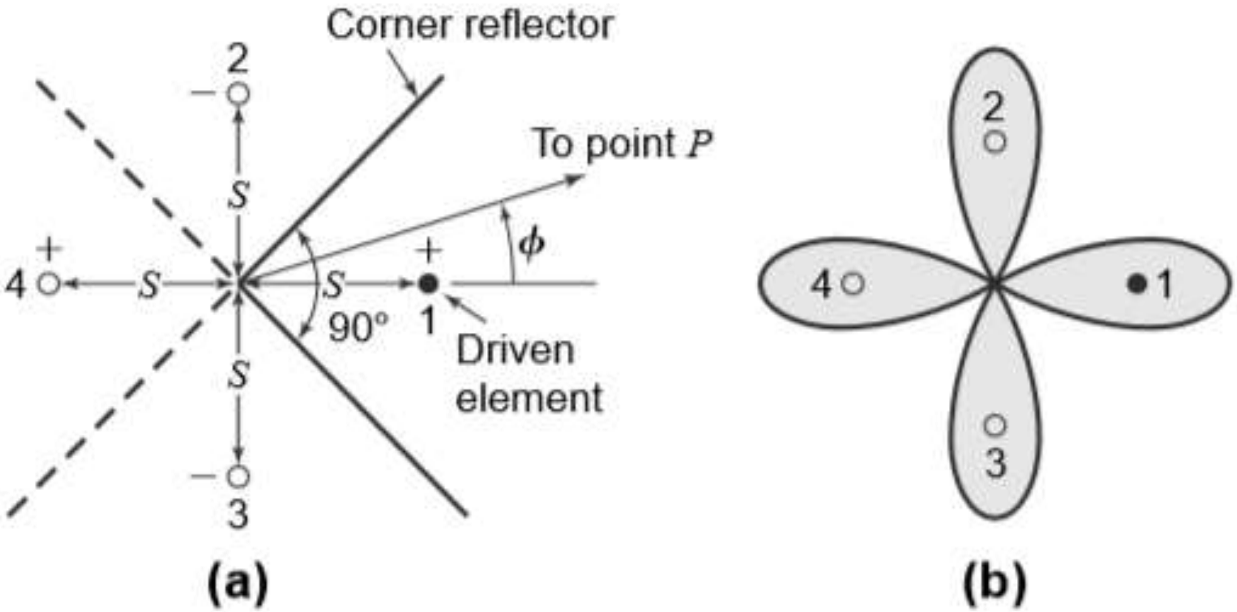


Figure 9–9 Square-corner reflector with images used in analysis (*a*) and 4-lobed pattern of driven element and images (*b*).

- The phase of the currents in 2 and 3 is the same but 180◦ out of phase with respect to the currents in 1 and 4. All elements are assumed to be λ/2 long.

Contd.,

- At the point P at a large distance D from the antenna, the field intensity is

$$E(\phi) = 2kI_1|[\cos(S_r \cos\phi) - \cos(S_r \sin\phi)]| \qquad (1)$$

where

I_1 = current in each element

S_r = spacing of each element from the corner, rad

$= 2\pi(S/\lambda)$

k = constant involving the distance D, etc.

- For arbitrary corner angles, analysis involves integrations of cylindrical functions which can be approximated by infinite sums as shown by Klopfenstein (1).

Contd.,

- The emf V1 at the terminals at the center of the driven element is

$$V_1 = I_1 Z_{11} + I_1 R_{1L} + I_1 Z_{14} - 2 I_1 Z_{12}$$

where

Z_{11} = self-impedance of driven element
R_{1L} = equivalent loss resistance of driven element
Z_{12} = mutual impedance of elements 1 and 2
Z_{14} = mutual impedance of elements 1 and 4

- Similar expressions can be written for the emf’s at the terminals of each of the images.

Contd.,

- Then if P is the power delivered to the driven element (power to each image element is also P), we have from symmetry that

$$I_1 = \sqrt{\frac{P}{R_{11} + R_{1L} + R_{14} - 2R_{12}}}$$

- $$E(\phi) = 2k\sqrt{\frac{P}{R_{11} + R_{1L} + R_{14} - 2R_{12}}}\,|[\cos(S_r \cos\phi) - \cos(S_r \sin\phi)]|$$

- The field intensity at the point P at a distance D from the driven $\lambda/2$ element with the reflector removed is

$$E_{HW}(\phi) = k\sqrt{\frac{P}{R_{11} + R_{1L}}}$$

Contd.,

- This is the relation for field intensity of a $\lambda/2$ dipole antenna in free space with a power input P and provides a convenient reference for the corner reflector antenna.
- Thus, dividing (4) by (5), we obtain the gain in field intensity of a square-corner reflector antenna over a single $\lambda/2$ antenna in free space with the same power input, or

$$G_f(\phi) = \frac{E(\phi)}{E_{HW}(\phi)} = 2\sqrt{\frac{R_{11}+R_{1L}}{R_{11}+R_{1L}+R_{14}-2R_{12}}}\,|[\cos(S_r\cos\phi)-\cos(S_r\sin\phi)]|$$

- where the expression in brackets is the pattern factor and the expression included under the radical sign is the coupling factor.
- The pattern shape is a function of both the angle φ and the antenna-to-corner spacing S. The pattern calculated by (6) has 4 lobes as shown in Fig. 9–9b. However, only one of the lobes is real.

Contd.,

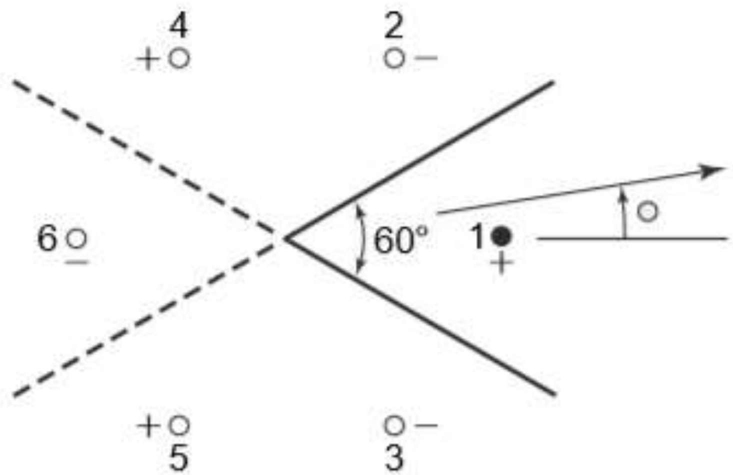


Figure 9–10 A 60° corner reflector with images used in analysis.

- Expressions for the gain in field intensity of corner reflectors with corner angles of 60◦,45◦, etc., can be obtained in a similar manner.
- The driven element is a λ/2 dipole. For the 60◦ corner the analysis requires a total of 6 elements, 1 actual antenna and 5 images as in Fig. 9–10. Gain-pattern expressions for corner reflectors of 90 and 60◦ are listed in Table 9–1. The expression for a 180◦ "corner" or flat sheet is also included.

Contd.,

Table 9–1 Gain-pattern formulas for corner reflector antennas

Corner angle, deg	Number of elements in analysis	Gain in field intensity over $\lambda/2$ antenna in free space with same power input
180	2	$2\sqrt{\dfrac{R_{11}+R_{1L}}{R_{11}+R_{1L}-R_{12}}}\sin(S_r\cos\phi)$
90	4	$2\sqrt{\dfrac{R_{11}+R_{1L}}{R_{11}+R_{1L}+R_{14}-2R_{12}}}\lvert[\cos(S_r\cos\phi)-\cos(S_r\sin\phi)]\rvert$
60	6	$2\sqrt{\dfrac{R_{11}+R_{1L}}{R_{11}+R_{1L}+2R_{14}-2R_{12}-R_{16}}}$ $\times\lvert\{\sin(S_r\cos\phi)-\sin[S_r\cos(60^\circ-\phi)]-\sin[S_r\cos(60^\circ+\phi)]\}\rvert$

- In the formulas of Table 9–1 it is assumed that the reflector sheets are perfectly conducting and of infinite extent. Curves of gain versus spacing calculated from these relations are presented in Fig. 9–11.

Contd.,

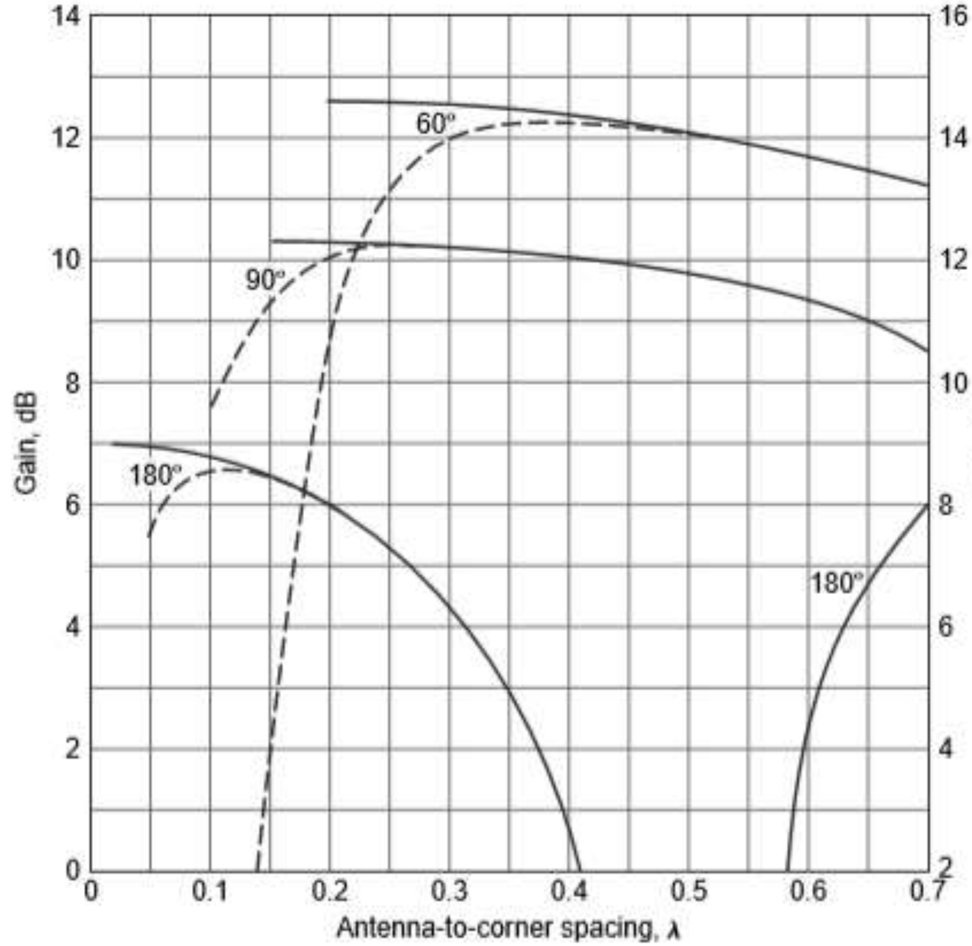


Figure 9–11 Gain of corner reflector antennas over a $\lambda/2$ dipole antenna in free space with the same power input as a function of the antenna-to-corner spacing. Gain is in the direction $\phi = 0$ and is shown for zero loss resistance (solid curves) and for an assumed loss resistance of 1 Ω ($R_{1L} = 1$ Ω) (dashed curves). (*After Kraus.*)

Contd.,

- The gain given is in the direction $\varphi = 0$. Two curves are shown for each corner angle.
- The solid curve in each case is computed for zero losses ($R1L = 0$), while the dashed curve is for an assumed loss resistance $R1L = 1\Omega$.
- It is apparent that for efficient operation too small a spacing should be avoided.
- A small spacing is also objectionable because of narrow bandwidth.
- On the other hand, too large a spacing results in less gain.

Contd.,

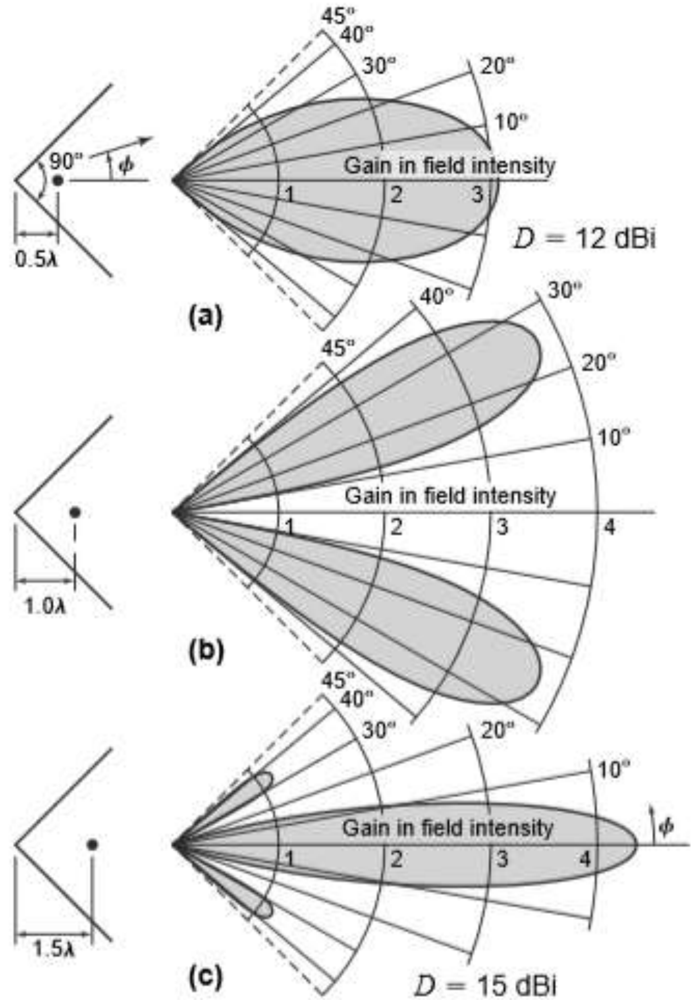
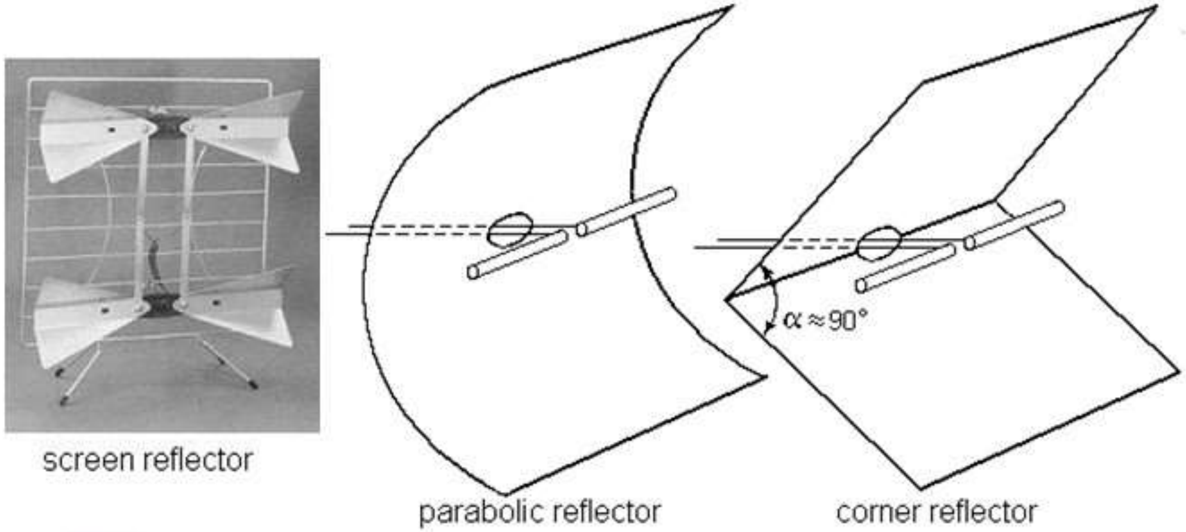
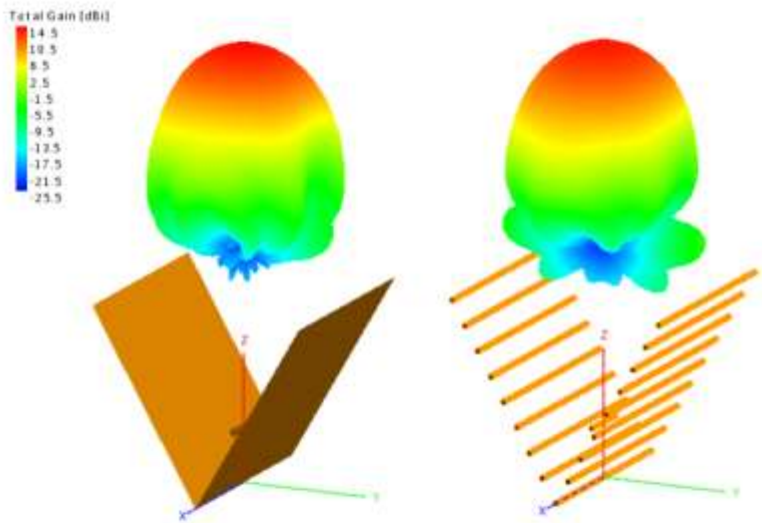


Figure 9–12 Calculated patterns of square-corner reflector antennas with antenna-to-corner spacings of (*a*) 0.5λ, (*b*) 1.0λ and (*c*) 1.5λ. Patterns give gain relative to the $\lambda/2$ dipole antenna in free space with the same power input.



screen reflector
parabolic reflector
corner reflector
α ≈ 90°



Total Gain [dBi]
14.5
10.5
2.5
-1.5
-5.5
-9.5
-17.5
-21.5
-25.5

Contd.,

- The calculated pattern of a 90° corner reflector with antenna-to-corner spacing $S = 0.5\lambda$ is shown in Fig. 9–12a.

- The gain is nearly 10 dB over a reference $\lambda/2$ antenna or 12 dBi. This pattern is typical if the spacing S is not too large. If S exceeds a certain value, a multilobed pattern may be obtained.
- Eg., a square-corner reflector with $S = 1.0\lambda$ has a 2-lobed pattern as in Fig. 9–12b.

- If the spacing is increased to 1.5, the pattern shown in Fig. 9–12c is obtained with the major lobe in the $\varphi = 0$ direction but with minor lobes present.

- This pattern may be considered as belonging to a higher-order radiation mode of the antenna.

- The gain over a single $\lambda/2$ dipole antenna is 12.9 dB ($\simeq$15 dBi). Restricting patterns to the lower-order radiation mode (no minor lobes), it is generally desirable that S lie between the following limits:

Contd.,

Corner angle, α	Corner-to-dipole spacing, *S*
90°	0.25–0.7λ
180° (flat sheet)	0.1–0.3λ

- The terminal resistance RT of the driven antenna is obtained by dividing (eq 2) by I1 and taking the real parts of the impedances. Thus,

$$R_T = R_{11} + R_{1L} + R_{14} - 2R_{12}$$

- If R1L = 0, the terminal resistance is all radiation resistance. The variation of the terminal radiation resistance of the driven element is presented in Fig. 9–13a as a function of the spacing S for corner angles α = 180,90 and 60◦.

- For α = 90◦ and S = 0.35λ, the resistance of the driven λ/2 dipole is the same as for a λ/2 dipole in free space.

Contd.,

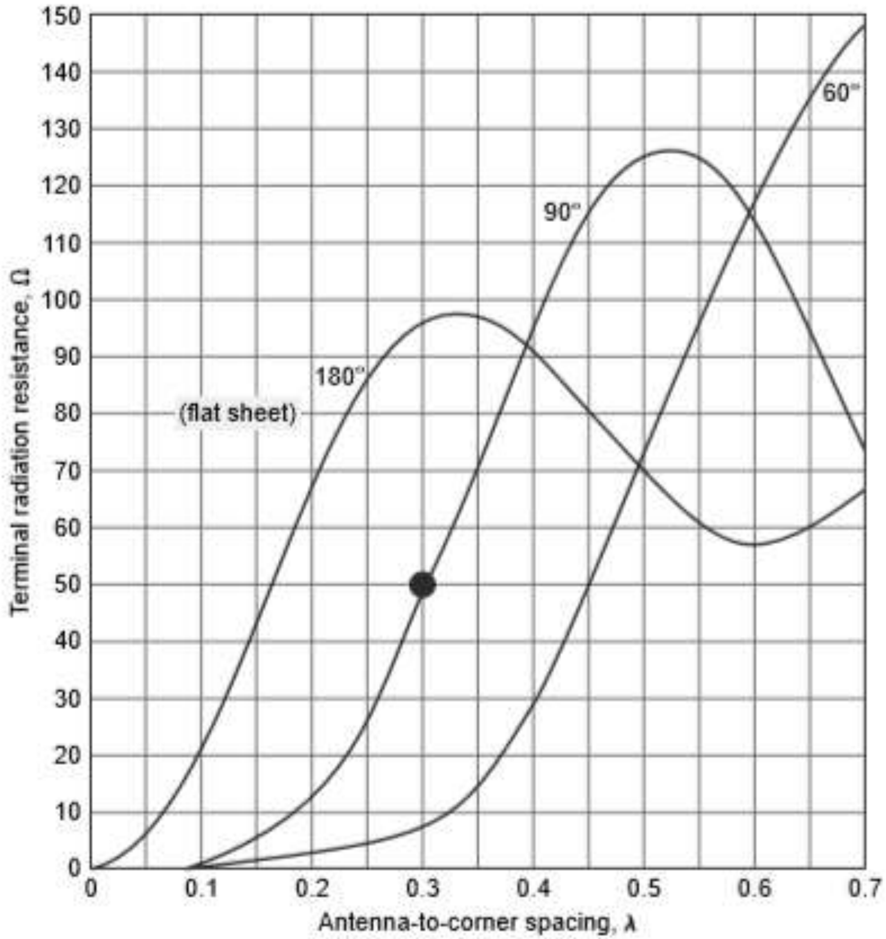


Figure 9–13a Radiation resistance of driven λ/2 dipole antenna as a function of the dipole-to-corner spacing for 60°, 90° and 180° corner reflectors (*after Kraus*). Note the convenient 50-Ω radiation resistance for the 90° corner with dipole-to-corner spacing of 0.3 λ (large dot).

- In the above analysis it is assumed that the reflectors are perfectly conducting and of infinite extent, with the exception that the gains with a finitely conducting reflector may be approximated with a proper choice of R1L.

Contd.,

- The analysis provides a good first approximation to the gain-pattern characteristics of actual corner reflectors with finite sides provided that the sides are not too small.

- Neglecting edge effects, a suitable value for the length of sides may be arrived at by the following line of reasoning.

- An essential region of the reflector is that near the point at which a wave from the driven antenna is reflected parallel to the axis. For example, this is the point A of the square-corner reflector of Fig. 9–13b.

- This point is at a distance of 1.41S from the corner C, where S is the antenna-to-corner spacing.

- If the reflector ends at the point B at a distance L = 2S from the corner, as in Fig. 9–13b, the reflector extends approximately 0.6S beyond A.

Contd.,

- With the reflector ending at B, it is to be noted that the only waves reflected from infinite sides, but not from finite sides, are those radiated in the sectors η.

- Furthermore, these waves are reflected with infinite sides into a direction that is at a considerable angle φ with respect to the axis.

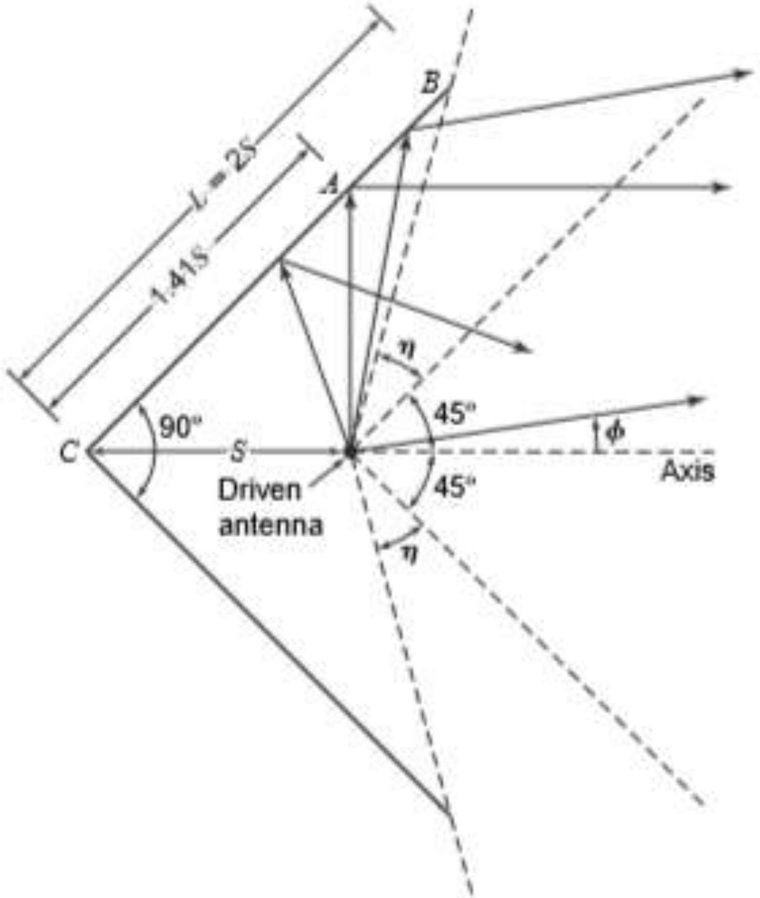


Figure 9–13b Square-corner reflector with sides of length *L* equal to twice the antenna-to-corner spacing *S*.

Contd.,

- Hence, the absence of the reflector beyond B should not have a large effect. It should also have relatively little effect on the driving-point impedance.
- The most noticeable effect with finite sides is that the measured pattern is appreciably broader than that calculated for infinite sides and a null does not occur at $\varphi = 45^\circ$ but at a somewhat larger angle.
- If this is not objectionable, a side length of twice the antenna-to-corner spacing ($L = 2S$)is a practical minimum value for square-corner reflectors.
- Although the gain of a corner reflector with infinite sides can be increased by reducing the corner angle, it does not follow that the gain of a corner reflector with finite sides of fixed length will increase as the corner angle is decreased.

Contd.,

- To maintain a given efficiency with a smaller corner angle requires that S be increased. Also on a 60° reflector, for example, the point at which a wave is reflected parallel to the axis is at a distance of 1.73S from the corner as compared to 1.41S for the square-corner type.

- Hence, to realize the increase in gain requires that the length of the reflector sides be much larger than for a square corner reflector designed for the same frequency.

- Usually this is a practical disadvantage in view of the relatively small increase to be expected in gain. To reduce the wind resistance offered by a solid reflector, a grid of parallel wires or conductors can be used as in Fig. 9–14.

- The supporting member joining the midpoints of the reflector conductors may be either a conductor or an insulator.

Contd.,

- In general the spacing s between reflector conductors should be equal to or less than $\lambda/8$. With a $\lambda/2$ driven element the length R of the reflector conductors should be equal to or greater than 0.7λ.

- If the length R is reduced to values of less than 0.6λ, radiation to the sides and rear tends to increase and the gain decreases. When R is decreased to as little as 0.3λ, the strongest radiation is no longer forward and the “reflector” acts as a director.

- The square-corner reflector is a simple, practical, inherently wideband antenna producing substantial gains (11 to 14 dBi). Typical design data for a 90° (square) corner reflector with bow-tie dipole for wideband (2 to 1 frequency range) operation are given in Table 9–2.

Contd.,

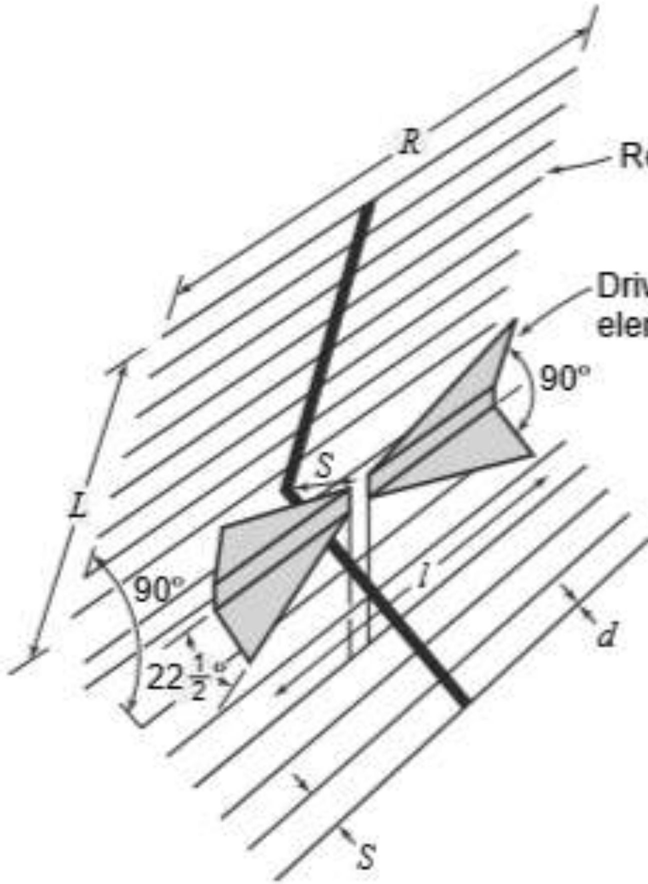


Figure 9–14 Square-corner (grid) reflector with bow-tie dipole for wideband operation (see Table 9–2).

Contd.,

Table 9–2 Design data for wideband 90° corner reflector with bow-tie dipole (see Figs. 9–14 and 9–15)

Dipole-to-corner spacing, $S = \lambda/4$ at lowest frequency
Length of reflector, $L = 3\lambda/4$ at lowest frequency
Reflector rod length, $R = 4\lambda/5$ at lowest frequency
Reflector rod spacing, $s = \lambda/8$ at highest frequency
Reflector rod diameter, $d = \lambda/50$ at highest frequency
Bow-tie dipole length, $l = 4\lambda/5$ at mid frequency

	Lowest frequency f_1	Mid frequency f_2	Highest frequency f_3	Units
Dipole-to-corner spacing, S	0.27	0.40	0.54	λ
Length of reflector, L	0.75	1.13	1.50	λ
Reflector rod length, R	0.81	1.20	1.62	λ
Reflector rod spacing, s	0.061	0.092	0.122	λ
Reflector rod diameter, d	0.01	0.015	0.02	λ
Bow-tie dipole length, l	0.53	0.80	1.06	λ
Gain	11.0	13.0	14.0	dBi

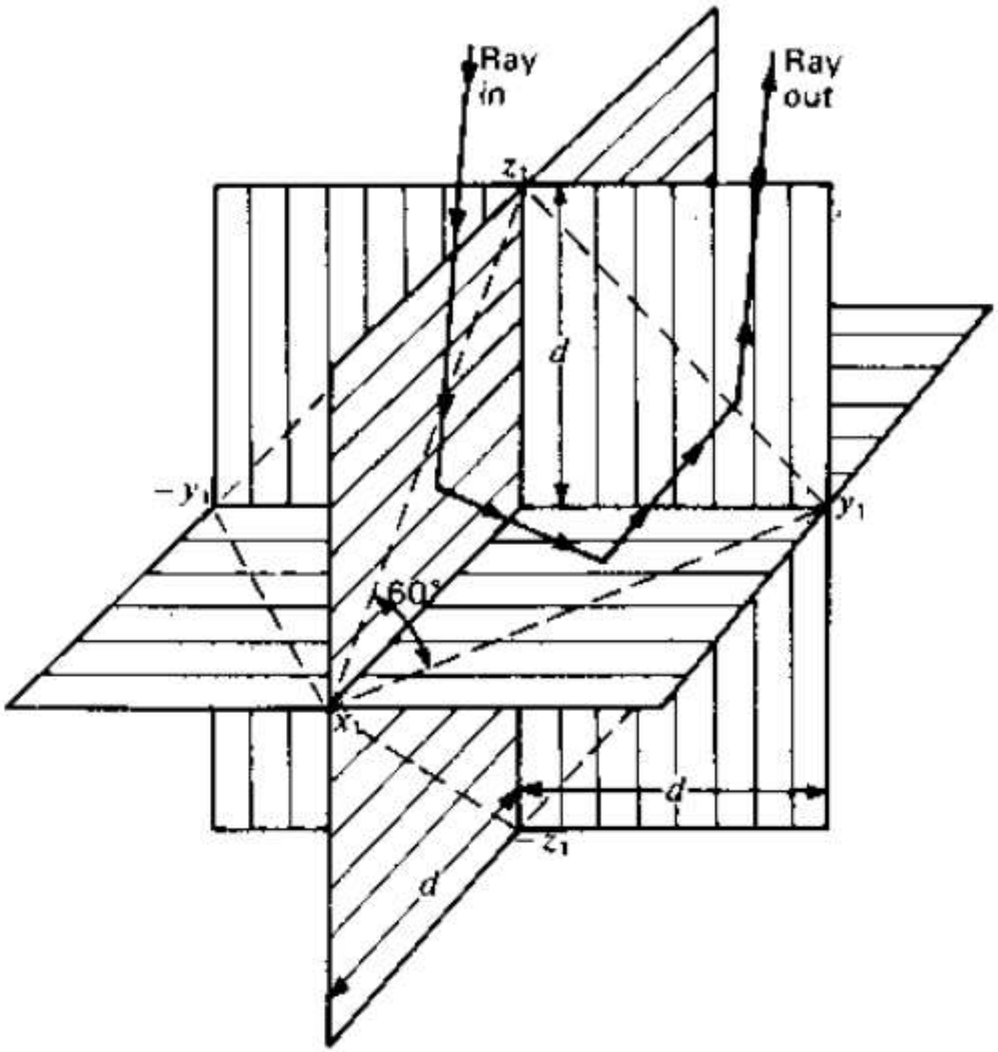


Figure 12-16 Retroreflector of 8 square corners for reflecting back waves from any direction. Path of ray returning via triple bounce is shown.

The Parabola General Properties

$$2L = R(1 + \cos\theta)$$

$$R = \frac{2L}{1 + \cos\theta}$$

$$PS = QS - PQ \text{ and } PF = PQ$$

$$PF + PS = PF + QS - PQ = QS$$

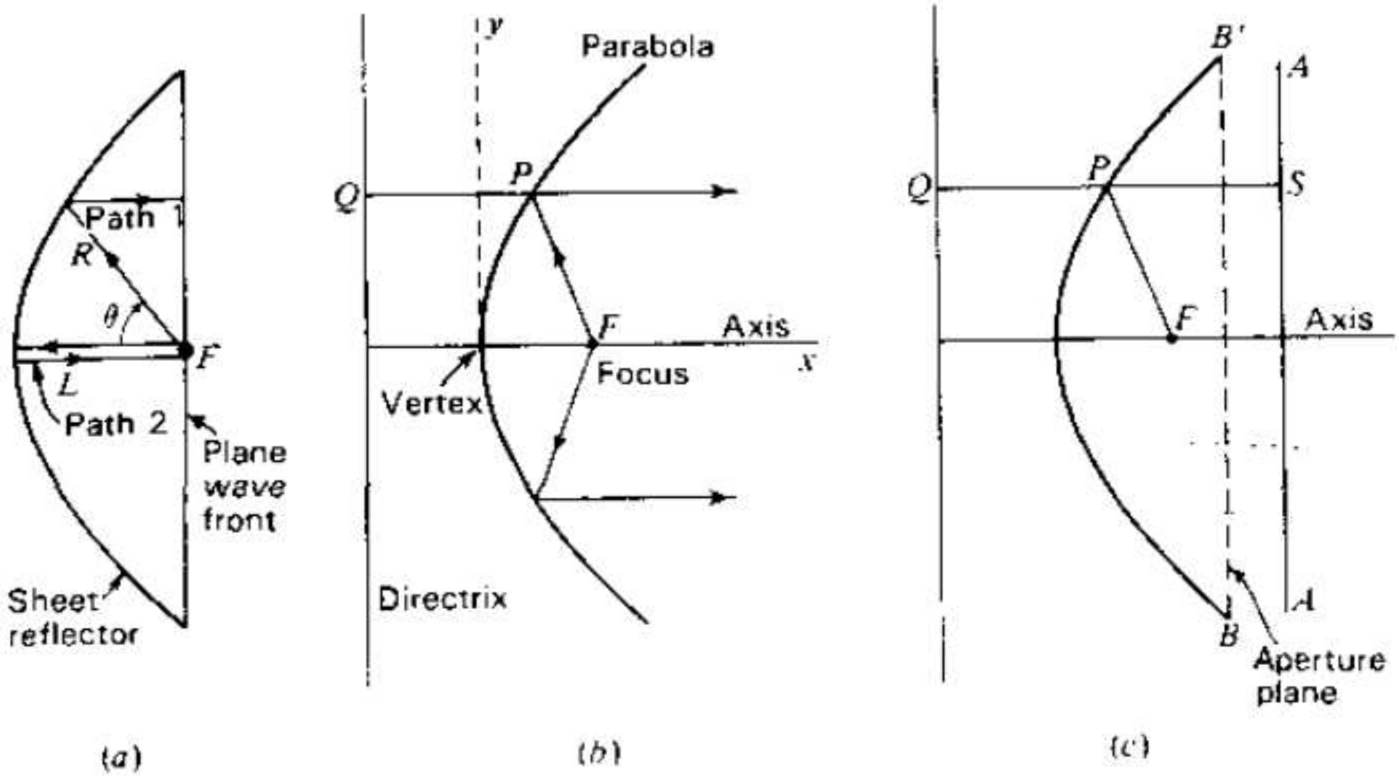
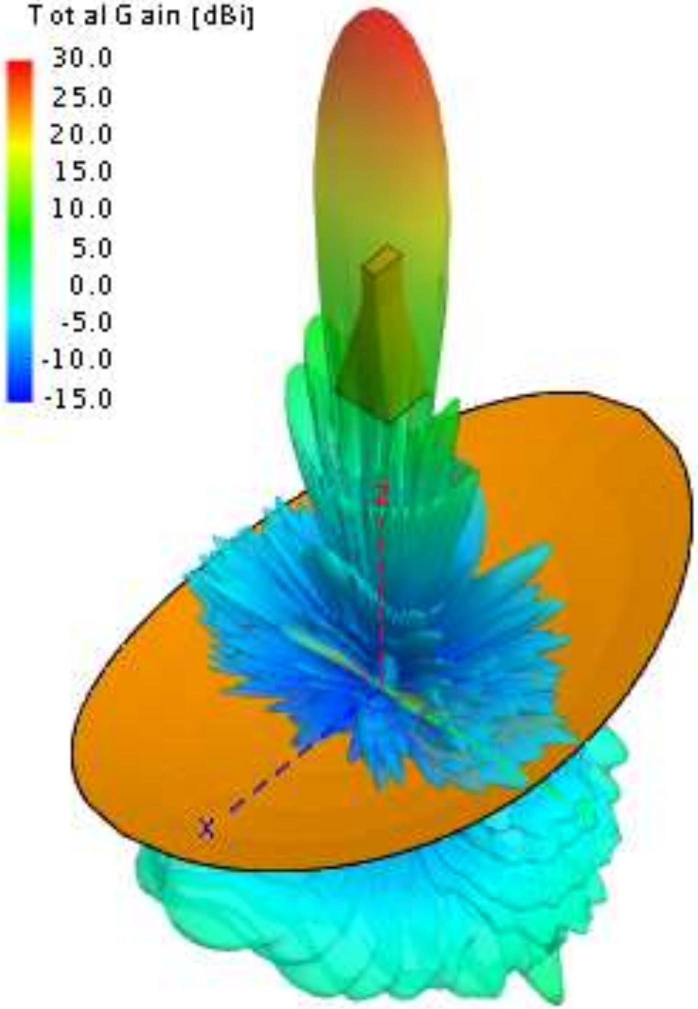
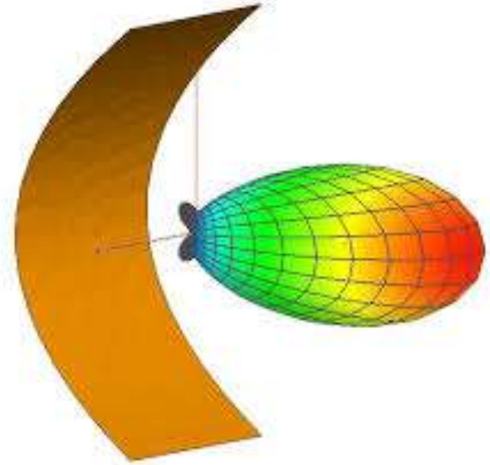
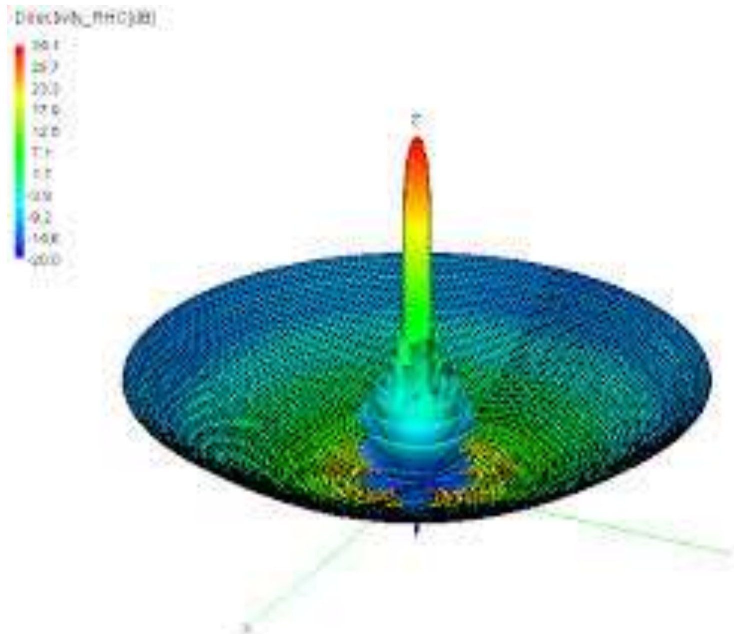


Figure 12-17 Parabolic reflectors.


Total Gain [dBi]
30.0
25.0
20.0
15.0
10.0
5.0
0.0
-5.0
-10.0
-15.0





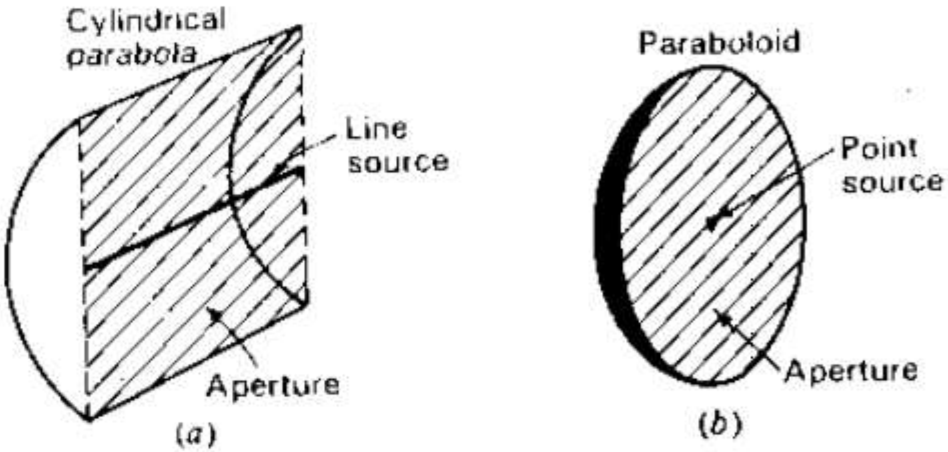


Figure 12-18. Line source and cylindrical parabolic reflector (*a*) and point source and paraboloidal reflector (*b*).

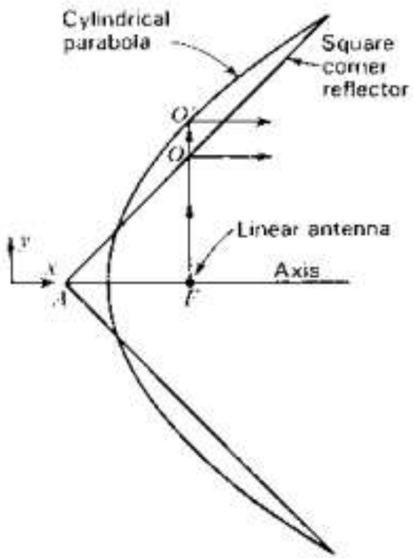


Figure 12-19 Cylindrical parabolic reflector compared with square-corner reflector.

The Paraboloidal Reflector

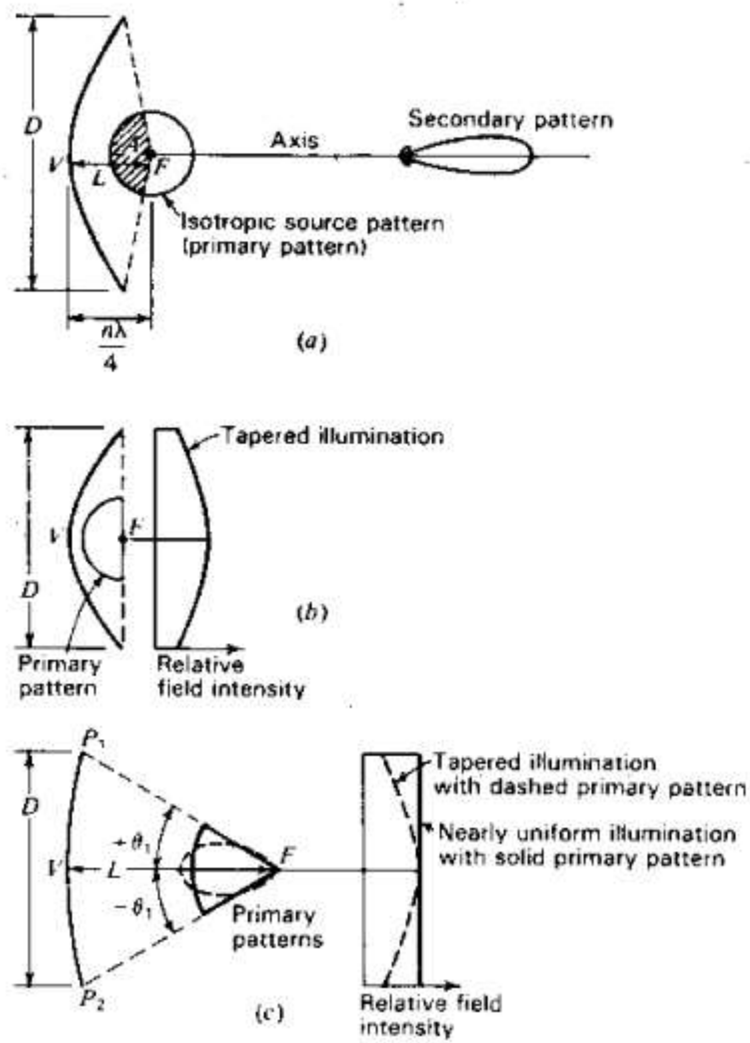


Figure 12-20 Parabolic reflectors of different focal lengths (L) and with sources of different patterns.

$$L = \frac{n\lambda}{4}$$

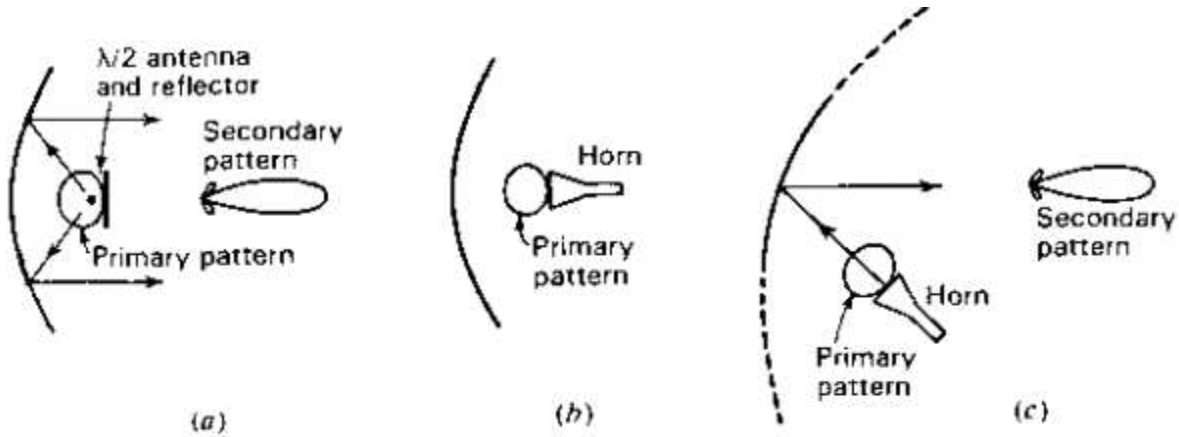


Figure 12-21 Full parabolic reflectors (*a* and *b*) and partial reflector with offset feed (*c*).

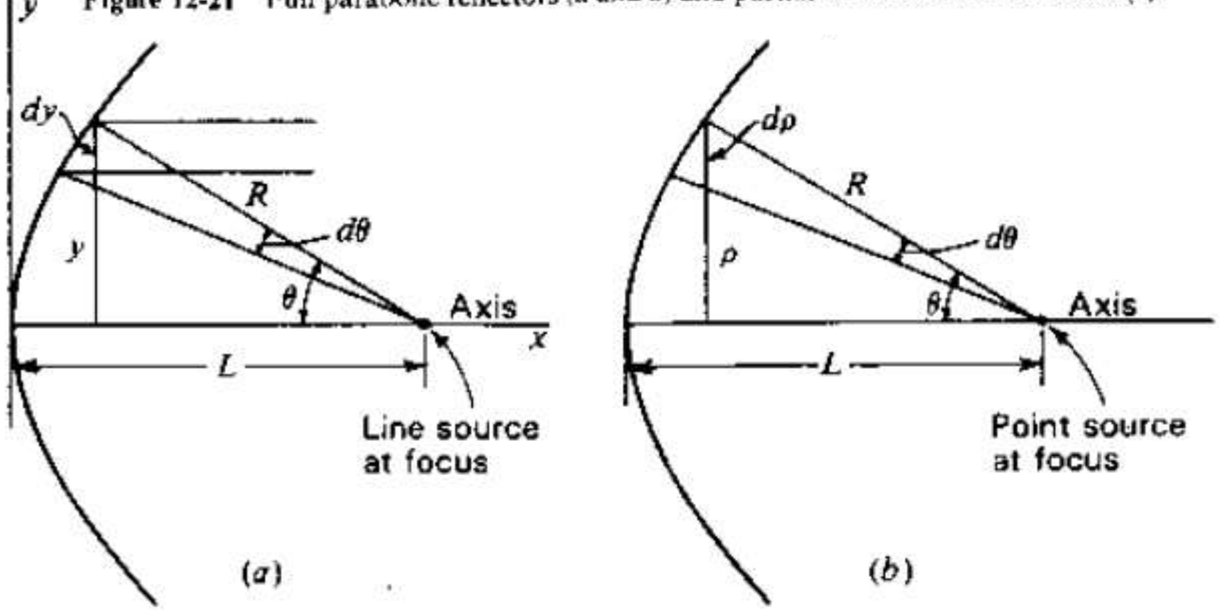


Figure 12-22 Cross sections of cylindrical parabola (*a*) and of paraboloid-of-revolution (*b*).

$P = dy\, S_y \qquad y = R \sin\theta$

$P = d\theta\, U'$ a *paraboloid-of-revolution* with an isotropic point source

$S_y\, dy = U'\, d\theta$

$$\frac{S_y}{U'} = \frac{1}{(d/d\theta)(R \sin\theta)}$$

$$R = \frac{2L}{1 + \cos\theta}$$

$$S_y = \frac{1 + \cos\theta}{2L} U'$$

$$\frac{S_\theta}{S_0} = \frac{1 + \cos\theta}{2}$$

$$\frac{E_\theta}{E_0} = \sqrt{\frac{1 + \cos\theta}{2}}$$

$P = 2\pi\rho\, d\rho\, S_\rho$

$P = 2\pi \sin\theta\, d\theta\, U$

$\rho\, d\rho\, S_\rho = \sin\theta\, d\theta\, U$

$$\frac{S_\rho}{U} = \frac{\sin\theta}{\rho(d\rho/d\theta)}$$

$$\rho = R \sin\theta = \frac{2L \sin\theta}{1 + \cos\theta}$$

$$S_\rho = \frac{(1 + \cos\theta)^2}{4L^2} U$$

$$\frac{S_\theta}{S_0} = \frac{(1 + \cos\theta)^2}{4}$$

$$\frac{E_\theta}{E_0} = \frac{1 + \cos\theta}{2}$$

$\rho = R \sin\theta.$

$$E(\phi) = \frac{2\lambda}{\pi D} \frac{J_1[(\pi D/\lambda) \sin \phi]}{\sin \phi}$$

where D = diameter of aperture, m

λ = free-space wavelength, m

ϕ = angle with respect to the normal to the aperture (Fig. 12-23)

J_1 = first-order Bessel function

The angle ϕ_0 to the first nulls of the radiation pattern are given by

$$\frac{\pi D}{\lambda} \sin \phi_0 = 3.83$$

since $J_1(x) = 0$ when $x = 3.83$. Thus,

$$\phi_0 = \arcsin \frac{3.83\lambda}{\pi D} = \arcsin \frac{1.22\lambda}{D}$$

When ϕ_0 is very small (aperture large)

$$\phi_0 \simeq \frac{1.22}{D_\lambda} \text{ (rad)} = \frac{70}{D_\lambda} \text{ (deg)}$$

where $D_\lambda = D/\lambda$ = diameter of aperture, λ

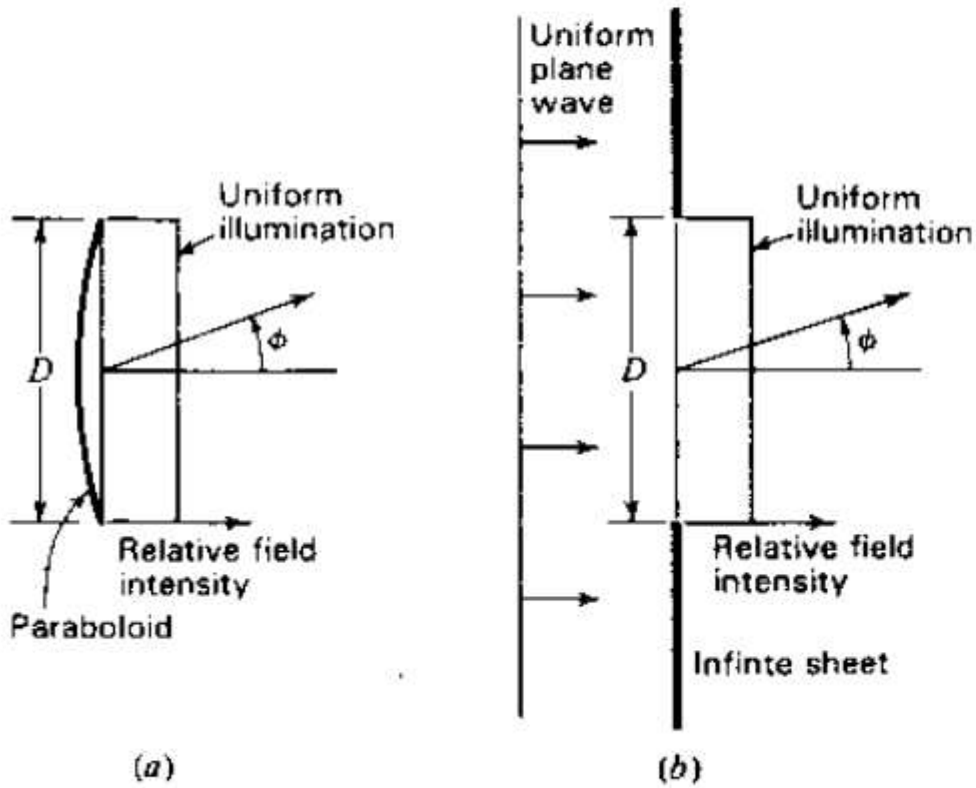


Figure 12-23 Large paraboloid with uniformly illuminated aperture (a) and equivalent uniformly illuminated aperture of same diameter D in infinite flat sheet (b).

The beam width between first nulls is twice this. Hence for large *circular apertures*, the *beam width between first nulls*

$$\text{BWFN} = \frac{140}{D_\lambda} \qquad \text{(deg)} \tag{5}$$

By way of comparison, the beam width between first nulls for a large uniformly illuminated *rectangular aperture* or a long linear array is

$$\text{BWFN} = \frac{115}{L_\lambda} \qquad \text{(deg)} \tag{6}$$

where L_λ = length of aperture, λ

The *beam width between half-power points* for a large circular aperture is[1]

$$\text{HPBW} = \frac{58}{D_\lambda} \qquad \text{(deg)} \tag{7}$$

The directivity D of a large *uniformly illuminated aperture* is given by

$$D = 4\pi \frac{\text{area}}{\lambda^2} \tag{8}$$

For a circular aperture

$$D = 4\pi \frac{\pi D^2}{4\lambda^2} = 9.87 D_\lambda^2 \tag{9}$$

where D_λ = the diameter of the aperture, λ

The power gain G of a circular aperture over a $\lambda/2$ dipole antenna is

$$G = 6 D_\lambda^2 \tag{10}$$

For example, an antenna with a uniformly illuminated circular aperture 10λ in diameter has a gain of 600 or nearly 28 dB with respect to a $\lambda/2$ dipole antenna ($\simeq 30$ dBi).

For a *square aperture*, the directivity is

$$D = 4\pi \frac{L^2}{\lambda^2} = 12.6 L_\lambda^2$$

and the power gain over a $\lambda/2$ dipole is

$$G = 7.7 L_\lambda^2$$

where L_λ = the length of a side, λ

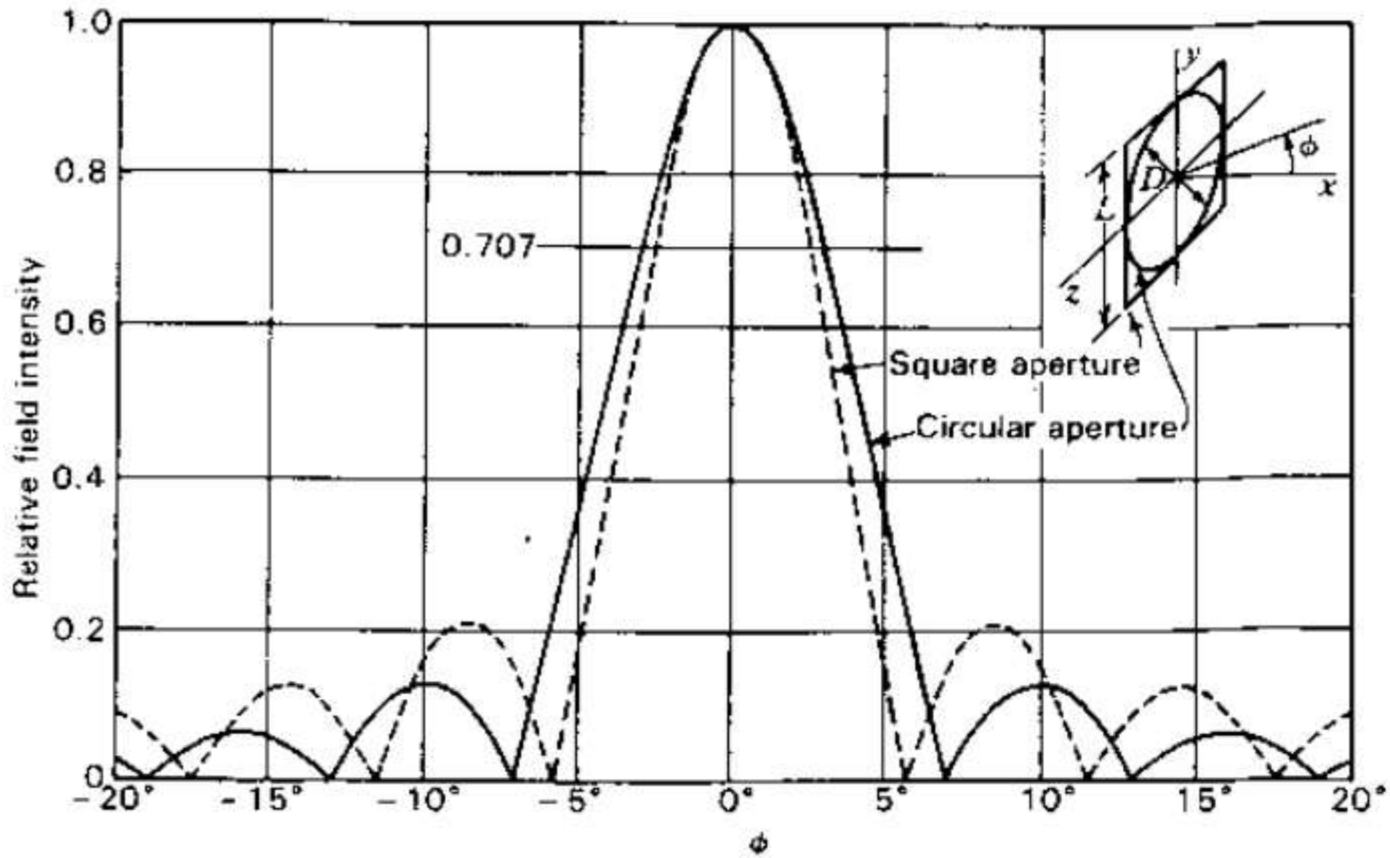


Figure 12-24 Relative radiation patterns of circular aperture of diameter $D = 10\lambda$ and of square aperture of side length $L = 10\lambda$.

Table 12-3 Beam widths, directivities and gains of circular and rectangular apertures with uniform aperture distributions†

	Aperture	
	Circular	Rectangular
Half-power beam width	$\frac{58^\circ}{D_\lambda}$	$\frac{51^\circ}{L_\lambda}$
Beam width between first nulls	$\frac{140^\circ}{D_\lambda}$	$\frac{115^\circ}{L_\lambda}$
Directivity (gain over isotropic source)	$9.9D_\lambda^2$	$12.6L_\lambda L'_\lambda$
Gain over $\lambda/2$ dipole	$6D_\lambda^2$	$7.7L_\lambda L'_\lambda$

where D_λ = diameter, λ
L_λ = side length, λ
L'_λ = length of other side (if aperture is square, $L'_\lambda = L_\lambda$)

† Apertures are assumed to be large compared to λ. With tapered distributions beam widths are larger, and directivities, gains and minor lobes are less.

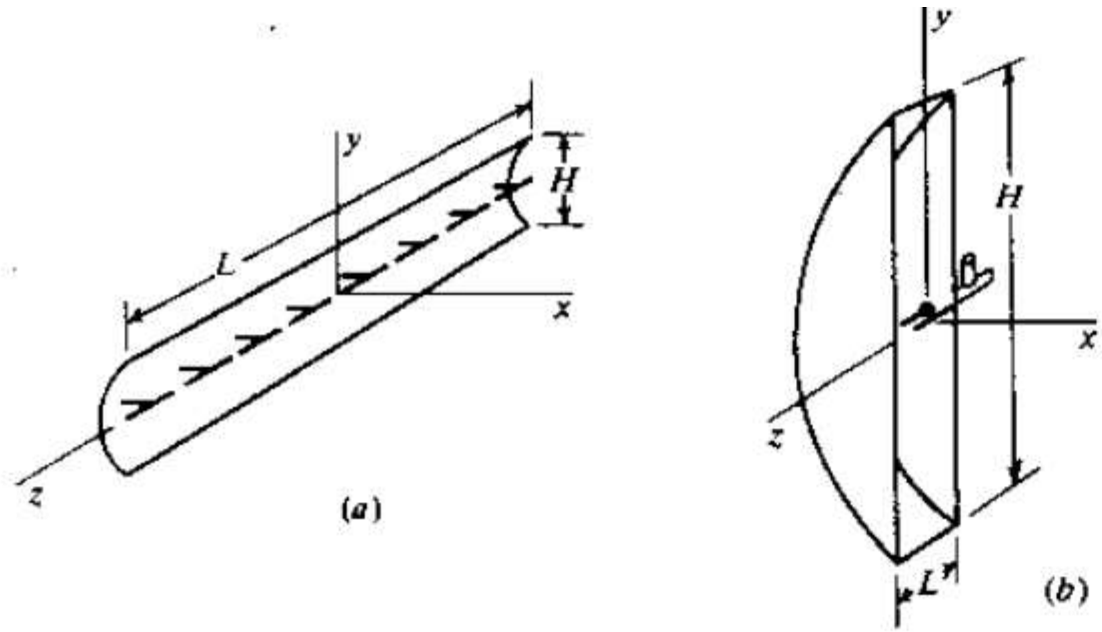


Figure 12-25 Parabolic reflector with linear array of 8 in-phase $\lambda/2$ dipole antennas (*a*) and "pillbox" or "cheese" antenna (*b*).

APERTURE DISTRIBUTIONS AND EFFICIENCIES.

$$P = SA_e$$

$$A_e = k_o A_{ep}$$

$$A_e = \frac{P}{S}$$

where A_e = actual effective aperture, m^2
k_o = ohmic-loss factor, dimensionless ($0 \leq k_o \leq 1$)
A_{ep} = effective aperture as determined entirely by *pattern*, m^2

$$D = \frac{4\pi}{\Omega_A} = \frac{4\pi}{\lambda^2} A_{ep}$$

from which[1]

$$A_{ep}\Omega_A = \lambda^2$$

and

$$A_e\Omega_A = k_o \lambda^2$$

where Ω_A = antenna beam solid angle, sr
λ = wavelength, m

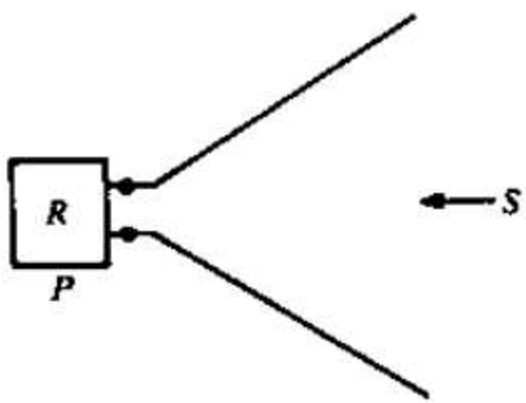


Figure 12-26 Wave of flux density S incident on antenna delivers a power P to the receiver R.

The *aperture efficiency* is defined as the ratio of the effective aperture to the physical aperture, or

$$\varepsilon_{ap} = \frac{A_e}{A_p} \tag{7}$$

so that the ratio of the aperture and beam efficiencies is

$$\frac{\varepsilon_{ap}}{\varepsilon_M} = \frac{A_e \Omega_A}{A_p \Omega_M} = \frac{k_o \lambda^2}{A_p \Omega_M} \tag{8}$$

where $\varepsilon_M = \Omega_M/\Omega_A$ = beam efficiency, dimensionless
Ω_M = main beam area, sr
Ω_A = total beam area, sr

$$D = \frac{4\pi}{\lambda^2} A_{ep}$$

$$G = Dk_o = k_o \frac{4\pi}{\lambda^2} A_{ep}$$

maximum directivity D_m

$$D_m = \frac{4\pi}{\lambda^2} A_p$$

$$D_d = \frac{4\pi}{\lambda^2} A_p k_u = D_m k_u$$

where D_d = design directivity

The factor k_u is called the *utilization factor*.

$$k_u = \frac{D_d}{D_m} \qquad 0 \le k_u \le 1$$

$$D = D_d k_a = D_m k_u k_a$$

k_a is the *achievement factor*,

$$k_a = \frac{D}{D_d} \qquad 0 \le k_a \le 1 \text{ (usually)}$$

The gain G of the antenna can now be written as

$$G = D_m k_o k_u k_a = \frac{4\pi}{\lambda^2} A_p k_o k_u k_a = \frac{4\pi}{\lambda^2} A_p \varepsilon_{ap}$$

where A_p = physical aperture, m^2
k_o = ohmic efficiency factor, dimensionless
k_u = utilization factor, dimensionless
k_a = achievement factor, dimensionless
ε_{ap} = aperture efficiency, dimensionless

We may also write

$$A_{ep} = k_u k_e A_p$$

$$A_e = k_o k_u k_a A_p$$

$$\varepsilon_{ap} = \frac{A_e}{A_p} = k_o k_u k_a$$

where A_{ep} = effective aperture (as determined entirely by pattern)
A_e = actual effective aperture

$$D = \frac{U_m}{U_{av}}$$

where U_m = maximum radiation intensity, W sr^{-1}
U_{av} = average radiation intensity, W sr^{-1}

$$D = \frac{U_m}{\dfrac{1}{4\pi} \displaystyle\iint_{4\pi} U(\theta, \phi)\, d\Omega} = \frac{4\pi U_m}{P}$$

$$DP = D_m P'$$

$$DP = 4\pi U_m$$

$$D = D_m \frac{P'}{P} = \frac{4\pi}{\lambda^2} A_p \frac{\dfrac{E_{av} E^*_{av}}{Z} A_p}{\displaystyle\iint_{A_p} \frac{E(x, y)E^*(x, y)}{Z}\, dx\, dy}$$

$$E_{av} = \frac{1}{A_p} \iint_{A_p} E(x, y)\, dx\, dy \qquad D = \frac{4\pi}{\lambda^2} A_p \frac{1}{\frac{1}{A_p} \iint_{A_p} \left[\frac{E(x, y)}{E_{av}}\right]\left[\frac{E(x, y)}{E_{av}}\right]^* dx\, dy}$$

δ is the *complex deviation (factor)*

$$D_d = \frac{4\pi}{\lambda^2} A_p \frac{1}{\frac{1}{A_p} \iint_{A_p} \left[\frac{E'(x, y)}{E'_{av}}\right]\left[\frac{E'(x, y)}{E'_{av}}\right]^* dx\, dy} \qquad D = D_m k_u k_a$$

$$E(x, y) = E_{av} + \delta E_{av}$$

$$D = \frac{4\pi}{\lambda^2} A_p \frac{1}{\frac{1}{A_p} \iint_{A_p} \left[\frac{E'(x, y)}{E'_{av}}\right]\left[\frac{E'(x, y)}{E'_{av}}\right]^* dx\, dy} \qquad \frac{E(x, y)}{E_{av}} = 1 + \delta$$

$$\times \frac{\frac{1}{A_p} \iint_{A_p} \left[\frac{E'(x, y)}{E'_{av}}\right]\left[\frac{E'(x, y)}{E'_{av}}\right]^* dx\, dy}{\frac{1}{A_p} \iint_{A_p} \left[\frac{E(x, y)}{E_{av}}\right]\left[\frac{E(x, y)}{E_{av}}\right]^* dx\, dy}$$

$$\frac{1}{A_p} \iint_{A_p} (1 + \delta)(1 + \delta)^*\, dx\, dy$$

$$1 + \frac{1}{A_p} \iint_{A_p} \delta\delta^*\, dx\, dy = 1 + \operatorname{var} \delta$$

$$D = \frac{4\pi}{\lambda^2} A_p \frac{1}{1 + \operatorname{var} \delta'} \frac{1 + \operatorname{var} \delta'}{1 + \operatorname{var} \delta}$$

where δ' = design value
δ = actual value

the *utilization factor*

$$k_u = \frac{1}{1 + \operatorname{var} \delta'}$$

the *achievement factor*

$$k_a = \frac{1 + \operatorname{var} \delta'}{1 + \operatorname{var} \delta}$$

$$\varepsilon_M = \frac{\Omega_M}{\Omega_A} = \frac{\displaystyle\iint_{\substack{\text{main}\\\text{lobe}}} P(\theta, \phi)\, d\Omega}{\displaystyle\iint_{4\pi} P(\theta, \phi)\, d\Omega} \qquad 0 \le \varepsilon_M \le 1$$

where $P(\theta, \phi)$ = antenna power pattern $(= EE^* = |E|^2)$

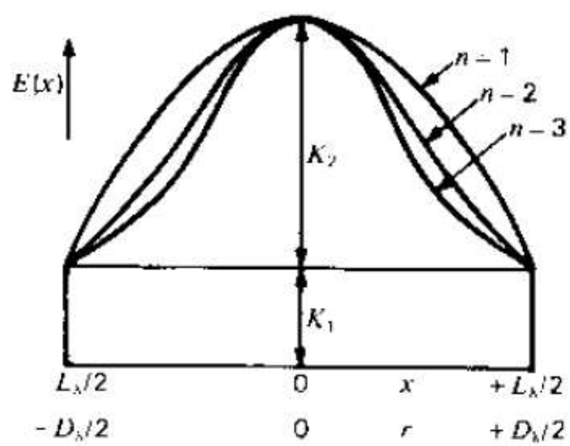


Figure 12-27 Various shapes of aperture distribution.

$$E(x) = K_1 + K_2\left[1 - \left(\frac{2x}{L_\lambda}\right)^2\right]^n$$

where $E(x)$ = field distribution
K_1 = constant (see Fig. 12-27)
K_2 = constant (see Fig. 12-27)
$L_\lambda = L/\lambda$ = aperture width, λ
n = integer (= 1, 2, 3, ...)

$$E(x, y) = \left\{K_1 + K_2\left[1 - \left(\frac{2x}{L_\lambda}\right)^2\right]\right\}\left\{K_1 + K_2\left[1 - \left(\frac{2y}{L'_\lambda}\right)^2\right]\right\}$$

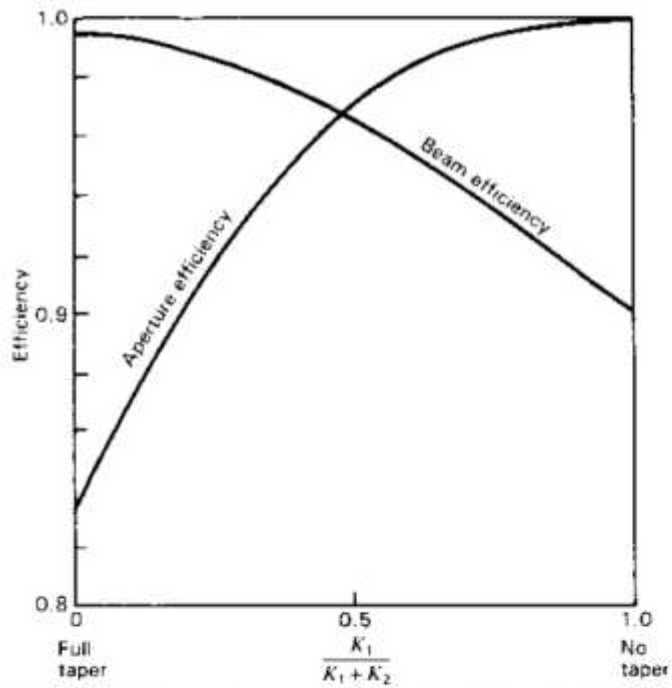


Figure 12-28 Beam and aperture efficiencies for a 1-dimensional aperture as a function of taper. The aperture efficiency is a maximum with no taper, while the beam efficiency is a maximum with full taper. A parabolic distribution for K_2 is assumed (see $n = 1$ in Fig. 12-27). (*After R. T. Nash, "Beam Efficiency Limitations for Large Antennas,"* IEEE Trans. Ants Prop., **AP-12**, *918–923, December 1964.*)

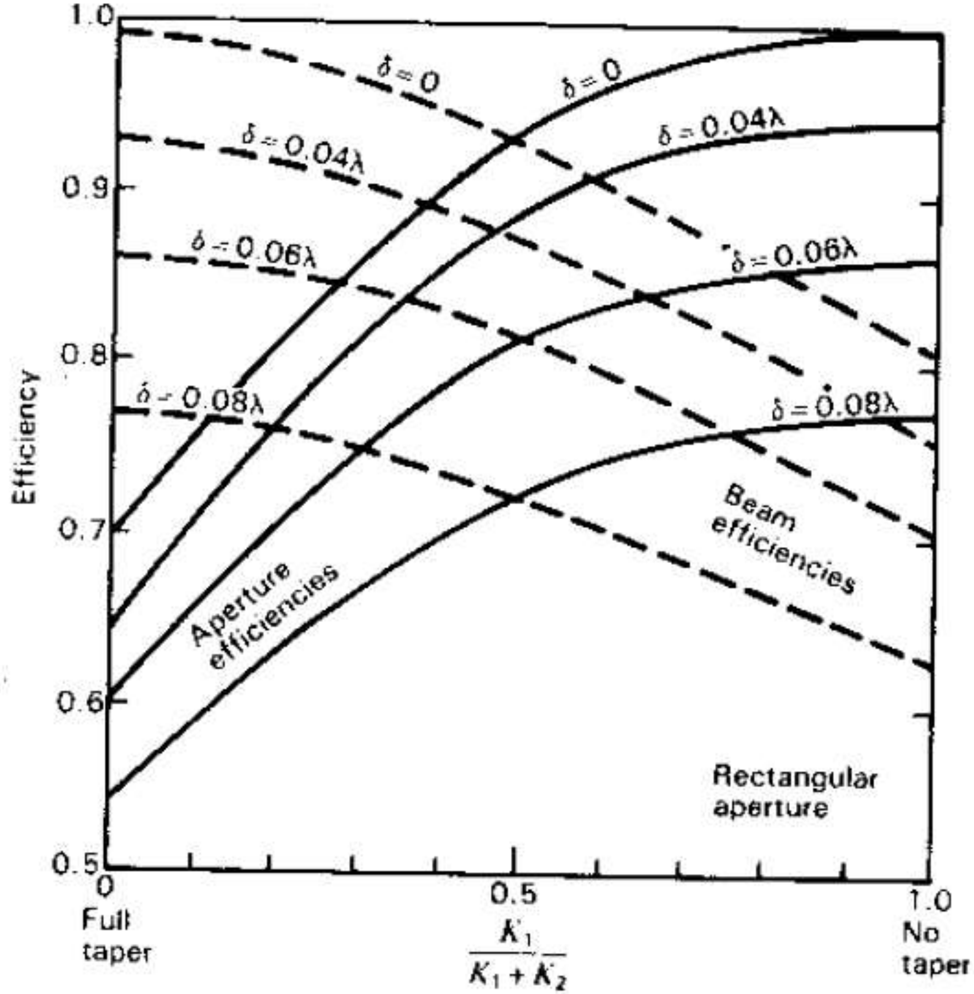


Figure 12-29a Aperture efficiency (solid) and beam efficiency (dashed) of a rectangular aperture as a function of taper and phase error. A parabolic distribution for K_2 is assumed (see $n = 1$ in Fig. 12-27). (*After R. T. Nash, "Beam Efficiency Limitations of Large Antennas,"* IEEE Trans. Ants. Prop., **AP-12**, *918–923, December 1964*).

the gain-degradation (or gain-loss) factor

$$k_g = e^{-(2\pi\delta/\lambda)^2}$$

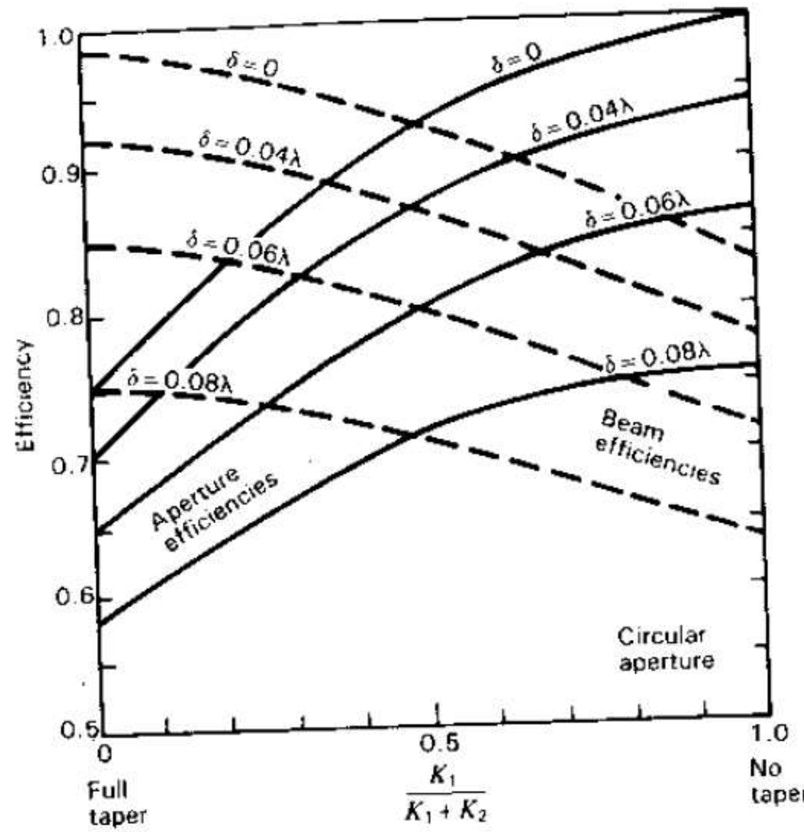


Figure 12-29b Aperture efficiency (solid) and beam efficiency (dashed) of a circular aperture as a function of taper and phase error. A parabolic distribution for K_2 is assumed (see $n = 1$ in Fig. 12-27). (*After R. T. Nash, "Beam Efficiency Limitations of Large Antennas,"* IEEE Trans. Ants. Prop., **AP-12**, *918–923, December 1964*).

the *feed efficiency* ε_f

$$\varepsilon_f = \frac{\displaystyle\iint_{\Omega_R} P_f(\theta, \phi)\, d\Omega}{\displaystyle\iint_{4\pi} P_f(\theta, \phi)\, d\Omega}$$

where $P_f(\theta, \phi)$ = power pattern of feed

Ω_R = solid angle subtended by reflector as viewed from feed point

$$\Omega_M = k_p \theta_{\mathrm{HP}} \phi_{\mathrm{HP}}$$

where k_p = factor between about 1.0 for a uniform aperture distribution and 1.13 for a Gaussian power pattern

θ_{HP} = half-power beam width in θ plane, sr

ϕ_{HP} = half-power beam width in ϕ plane, sr

$$\varepsilon_M = e^{-(4\pi\delta'/\lambda)^2} \frac{\displaystyle\iint_{\substack{\text{main}\\ \text{lobe}}} P(\theta, \phi)\, d\Omega}{\displaystyle\iint_{4\pi} P(\theta, \phi)\, d\Omega} \frac{\displaystyle\iint_{\Omega_R} P_f(\theta, \phi)\, d\Omega}{\displaystyle\iint_{4\pi} P_f(\theta, \phi)\, d\Omega}$$

where δ' = rms surface error of reflector

$P(\theta, \phi)$ = power pattern of reflector due to aperture distribution produced by the feed assuming no phase error

$P_f(\theta, \phi)$ = power pattern of feed

$$\mathrm{HPBW(deg)} = \frac{\mathrm{HPBW(min)}\cos\delta}{4}$$

$$\Omega_A = \frac{k_0\lambda^2}{A_e} \qquad \varepsilon_{\mathrm{ap}} = \frac{A_e}{A_p}$$

$$A_e = \frac{2kT_A}{S}$$

$$\varepsilon_M = \frac{\Omega_M}{\Omega_A} = \frac{k_p\,\theta_{\mathrm{HP}}\,\phi_{\mathrm{HP}}\,2kT_A}{k_o\lambda^2 S}$$

where k = Boltzmann's constant ($= 1.38 \times 10^{-23}$ J K^{-1})

T_A = antenna temperature due to radio source (measured value corrected for cable loss), K

S = flux density of radio source, W m^{-2} Hz^{-1}

λ = wavelength, m

k_o = antenna ohmic-loss factor, dimensionless

$$\varepsilon_{ap} = \frac{A_e}{A_p} = k_o k_u k_a = k_o k_u k_1 k_2 \cdots k_N$$

$$= k_o k_u \prod_{n=1}^{N} k_n$$

where k_o = ohmic-loss factor

k_u = utilization factor (by design)

k_a = achievement factor

$k_1 = e^{-(4\pi\delta'/\lambda)^2}$ = random reflector-surface-error (gain-loss) factor = k_g

$k_2 = \varepsilon_f$ = feed-efficiency factor

k_3 = aperture-blocking factor

k_4 = squint factor

k_5 = astigmatism factor

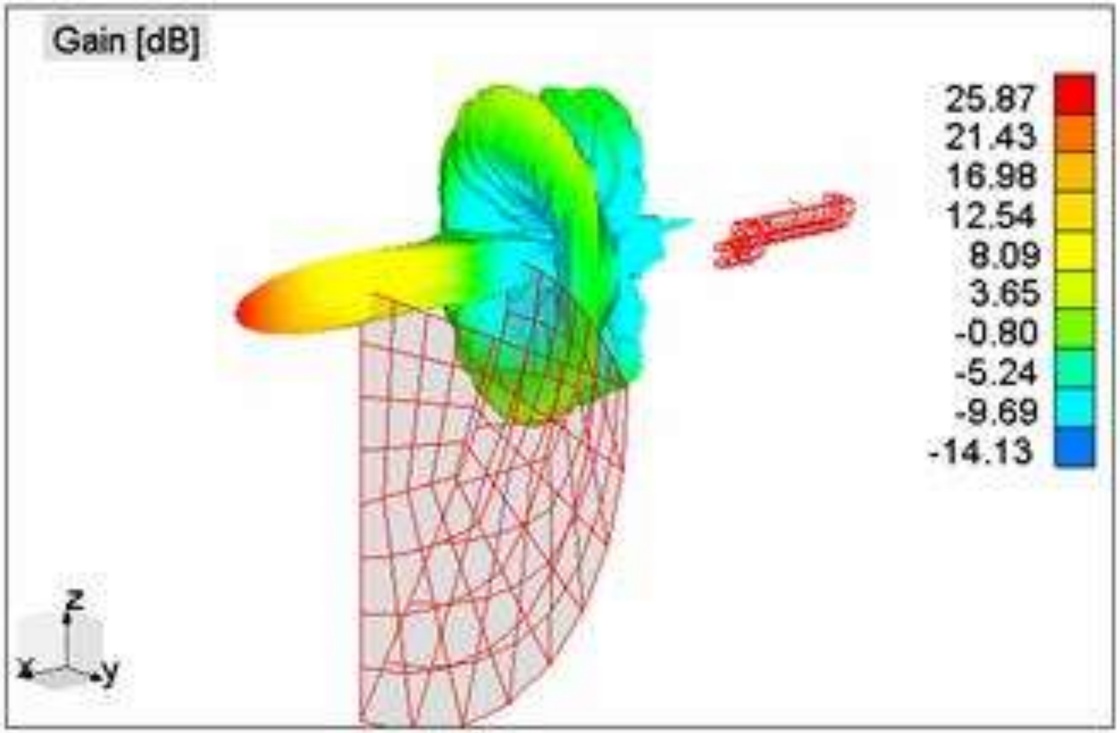
k_6 = surface-leakage factor

$$\varepsilon_{ap} = \frac{E_{av}^2}{(E^2)_{av}}$$

where E = field at any point in aperture

E_{av} = average of E over aperture

$(E^2)_{av}$ = average of E^2 over aperture



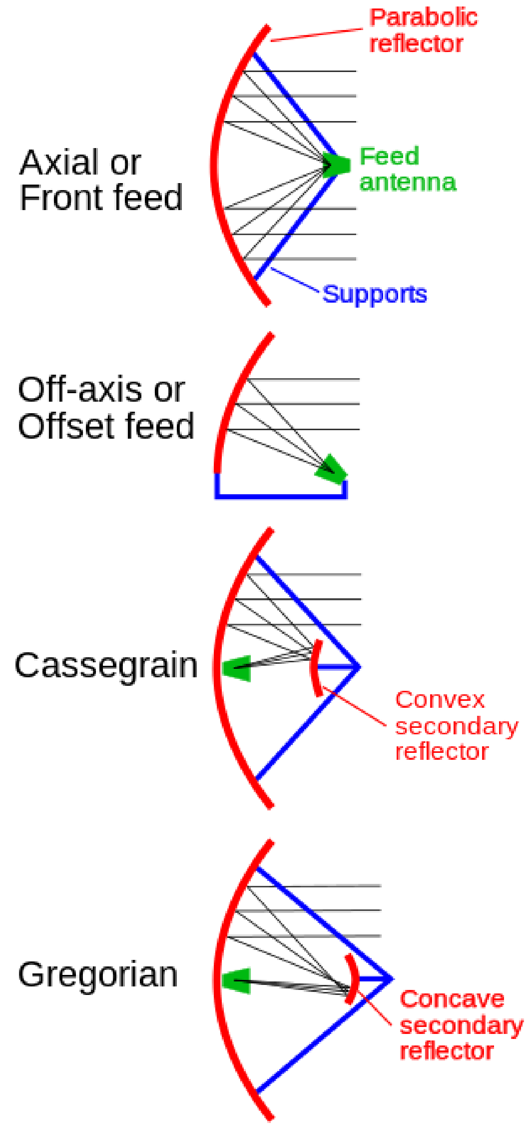
Gain [dB]
25.87
21.43
16.98
12.54
8.09
3.65
-0.80
-5.24
-9.69
-14.13
z
x
y

Reflector Antenna Feed Types

- There are several different types of parabolic reflector feed systems that can be used.

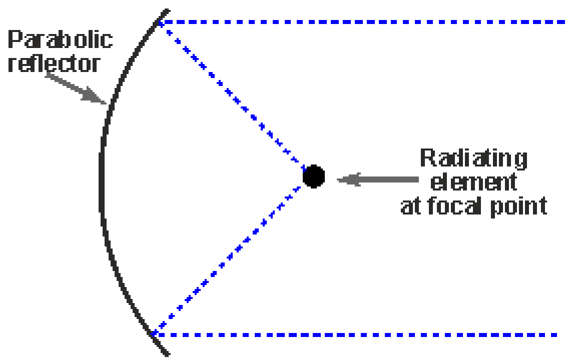
- Each has its own characteristics that can be matched to the requirements of the application.

  - Focal feed - often also known as axial or front feed system
  - Cassegrain feed system
  - Gregorian feed system
  - Off Axis or offset feed



Focal feed system

- The parabolic reflector or dish antenna consists of a radiating element which may be a simple dipole or a waveguide horn antenna. This is placed at the focal point of the parabolic reflecting surface.
- The energy from the radiating element is arranged so that it illuminates the reflecting surface.
- Once the energy is reflected it leaves the antenna system in a narrow beam. As a result considerable levels of gain can be achieved.

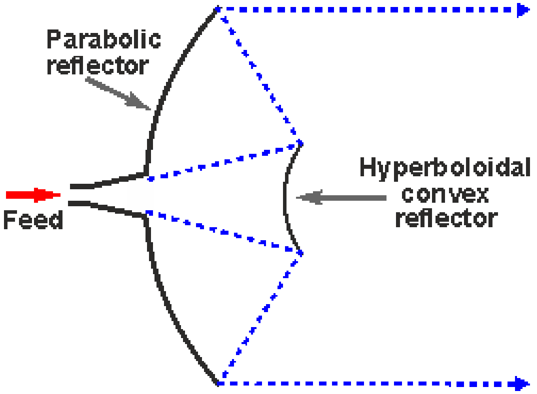


- Achieving this is not always easy because it is dependent upon the radiator that is used.
- For lower frequencies a dipole element is often employed whereas at higher frequencies a circular waveguide may be used.
- In fact the circular waveguide provides one of the optimum sources of illumination.

Cassegrain feed system

- The Cassegrain feed system, although requiring a second reflecting surface has the advantage that the overall length of the dish antenna between the two reflectors is shorter than the length between the radiating element and the parabolic reflector.

- This is because there is a reflection in the focusing of the signal which shortens the physical length. This can be an advantage in some systems.



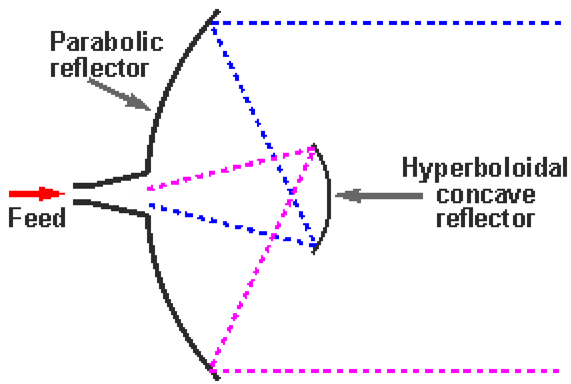
- Typical efficiency levels of 65 to 70% can be achieved using this form of parabolic reflector feed system

- The Cassegrain parabolic reflector antenna design and feed system gains its name because the basic concept was adapted from the Cassegrain telescope. This was reflecting telescope which was developed around 1672 and attributed to French priest Laurent Cassegrain.

Gregorian parabolic reflector feed

- The Gregorian parabolic reflector feed technique is very similar to the Cassegrain design.

- The major difference is that except that the secondary reflector is concave or more correctly ellipsoidal in shape.

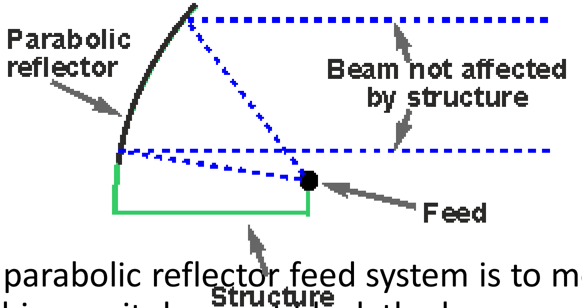


- Typical aperture efficiency levels of over 70% can be achieved because the system is able to provide a better illumination of all of the reflector surface.

Off axis or offset parabolic reflector antenna feed

- As the name indicates this form of parabolic reflector antenna feed is offset from the centre of the actual antenna dish used.

- The reflector used in this type of feed system is an asymmetrical segment of the parabolic shape normally used. In this way the focus, and the feed antenna are located to one side of the reflector surface.

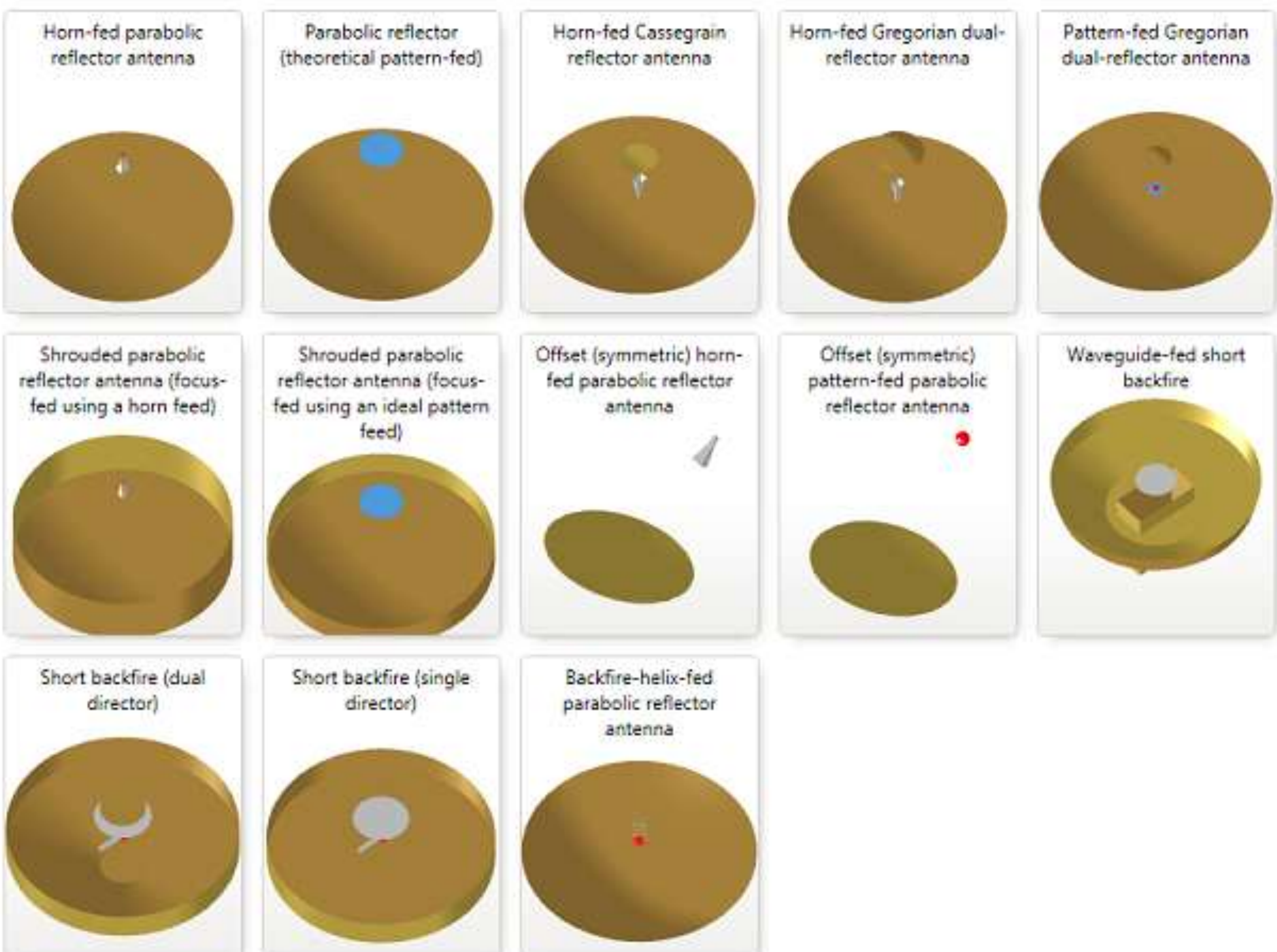


- The advantage of using this approach to the parabolic reflector feed system is to move the feed structure out of the beam path. In this way it does not block the beam.

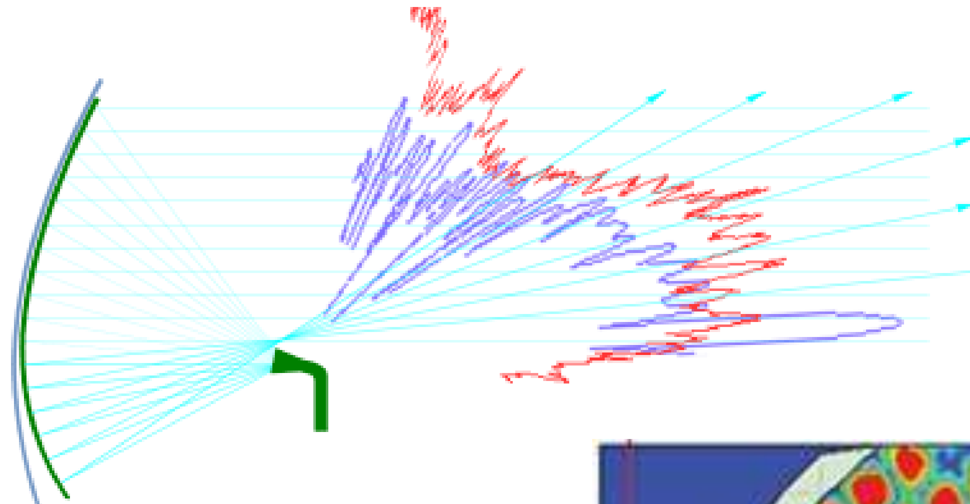
- This approach is widely used in home satellite television antennas, which are often relatively small and this would mean that any the feed structure including the low noise box (amplifier, etc) would otherwise block a significant percentage of the beam and thereby reduce the antenna efficiency and signal level.

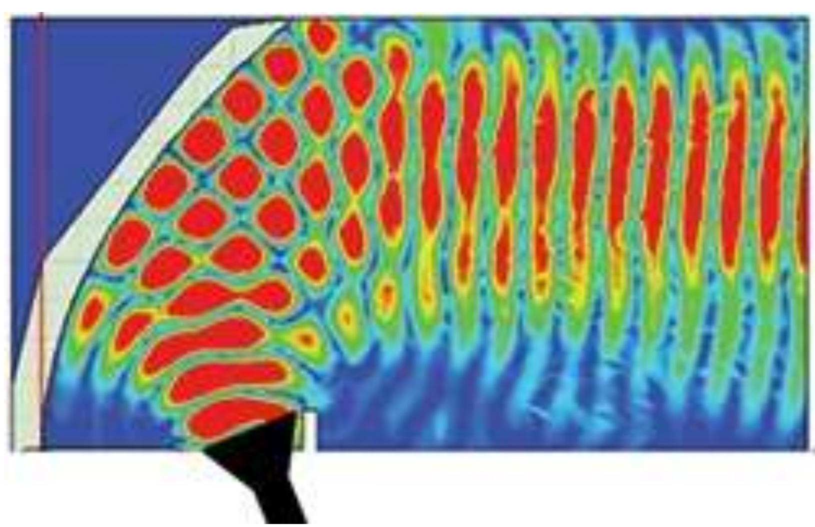
- The offset feed is also used in multiple reflector designs such as the Cassegrain and Gregorian because the small reflector would also suffer the same issues.

Reflector Antenna Feeds



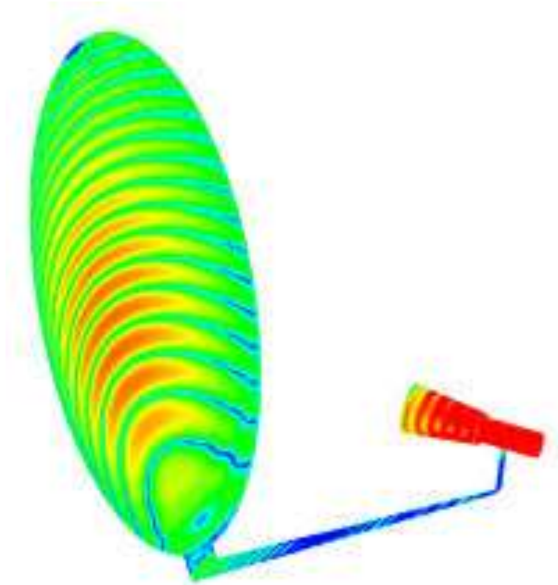
Offset Feed Radiation

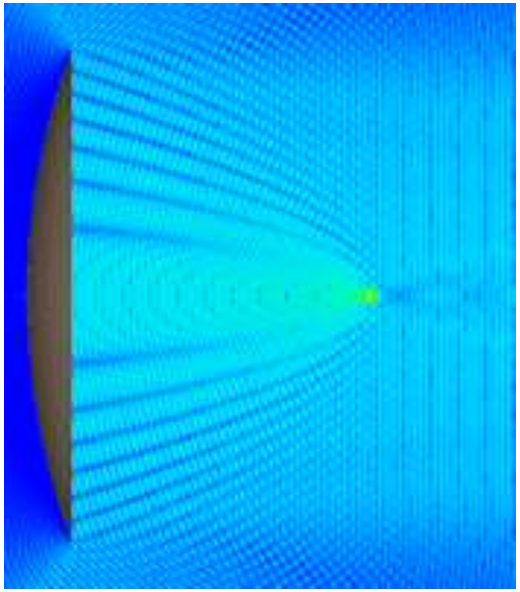




Current Distributions

- Reflector antenna current distribution





- Near-field phase fronts in front of reflector antenna

Factors affecting Parabolic Reflector Antenna Gain

- ***Diameter of reflecting surface*** The larger the diameter of the reflecting surface of the antenna the higher the parabolic reflector gain will be.
- ***Antenna efficiency:*** The efficiency of the antenna has a significant effect on the overall parabolic reflector gain. Typical figures are between 50 and 70%.
- ***Operational wavelength:*** The parabolic reflector antenna gain is dependent upon the reflector size in terms of wavelengths. Therefore if the same reflector is used on two different frequencies, the gain will be different and inversely proportional to the wavelength.

Advantages

- ***High gain:*** Parabolic reflector antennas are able to provide very high levels of gain. The larger the 'dish' in terms of wavelengths, the higher the gain.

- ***High directivity:*** As with the gain, so too the parabolic reflector or dish antenna is able to provide high levels of directivity. The higher the gain, the narrower the beamwidth. This can be a significant advantage in applications where the power is only required to be directed over a small area. This can prevent it, for example causing interference to other users, and this is important when communicating with satellites because it enables satellites using the same frequency bands to be separated by distance or more particularly by angle at the antenna.

Disadvantages

- ***Requires reflector and drive element:*** the parabolic reflector itself is only part of the antenna. It requires a feed system to be placed at the focus of the parabolic reflector.

- ***Cost :*** The antenna needs to be manufactured with care. A paraboloid is needed to reflect the radio signals which must be made carefully. In addition to this a feed system is also required. This can add cost to the system

- ***Size:*** The antenna is not as small as some types of antenna, although many used for satellite television reception are quite compact.

Applications

- ***Direct broadcast television:*** Direct broadcast or satellite television has become a major form of distribution for television material. The wide and controllable coverage areas available combined with the much larger bandwidths for more channels available mean that satellite television is very attractive.

- However as signal levels are low, directive antennas must be used to provide sufficient gain while being able to receive signals from only one satellite in the visible sky. The parabolic reflector antenna is able to meet these requirements and has the added advantage that it would not be as long as a Yagi or equivalent gain and directivity.

Applications

- ***Microwave links:*** Terrestrial microwave links are used for many applications. Often they are used for terrestrial telecommunications infrastructure links. One of the major areas where they are used these days is to provide the backhaul for mobile phone / cellular backhaul.

- ***Satellite communications:*** Many satellite uplinks, or those for communication satellites require high levels of gain to ensure the optimum signal conditions and that transmitted power from the ground does not affect other satellites in close angular proximity. Again the ideal antenna for most applications is the parabolic reflector antenna.

- ***Radio astronomy:*** Radio astronomy is an area where very high levels of gain and directivity are required. Accordingly the parabolic reflector antenna is an ideal choice.

Microstrip Antennas

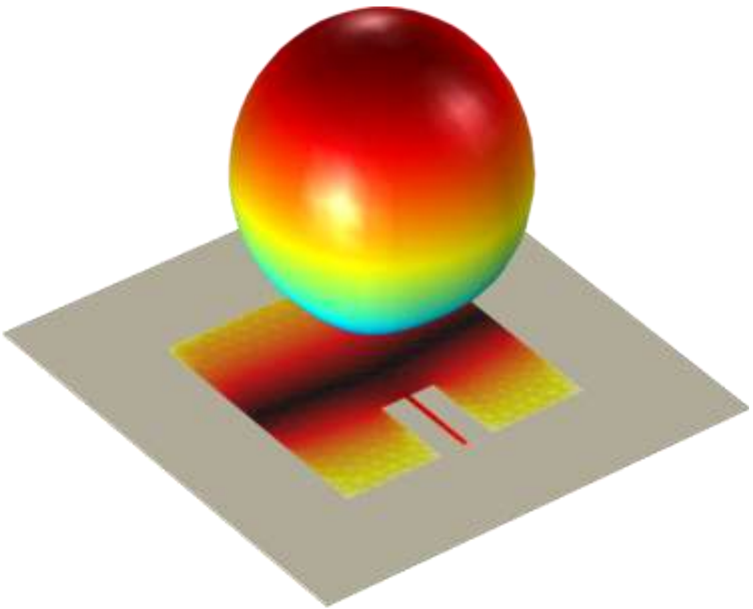
- Introduction
- Features
- advantages and limitations
- Rectangular patch antennas
- Geometry and parameters
- characteristics of Microstrip antennas
- Impact of different parameters on characteristics.

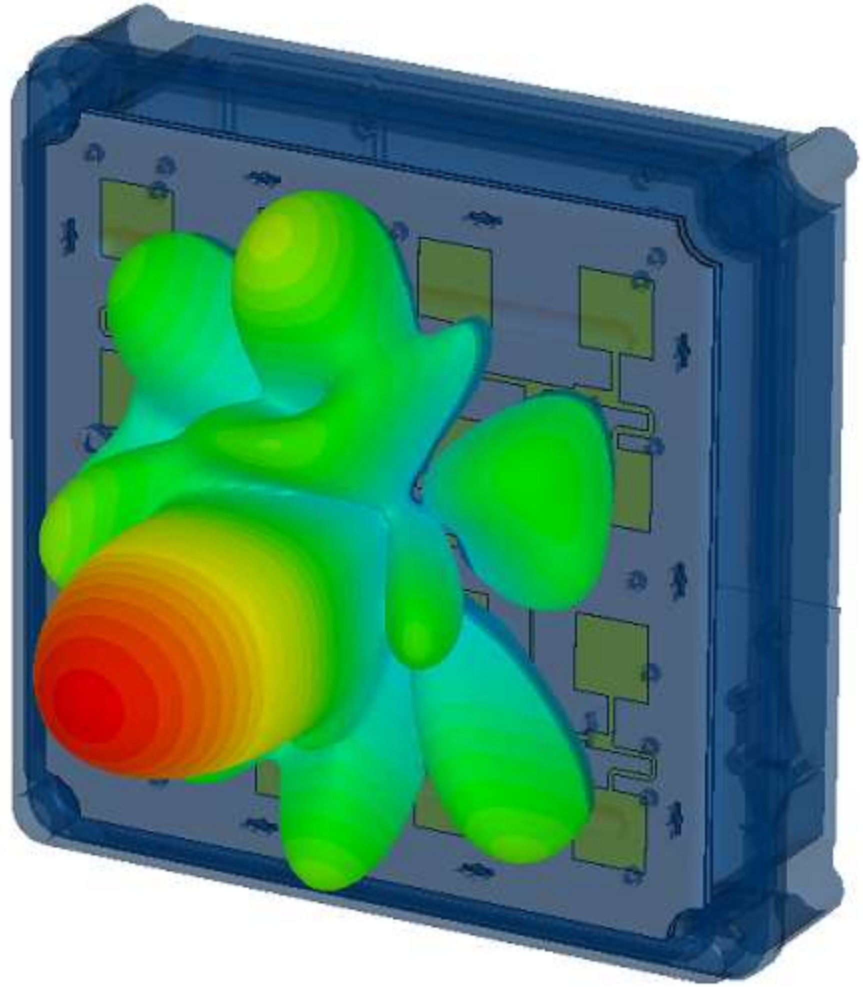
Patch Antenna

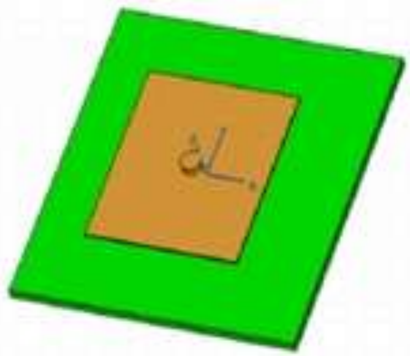
Single element

Array

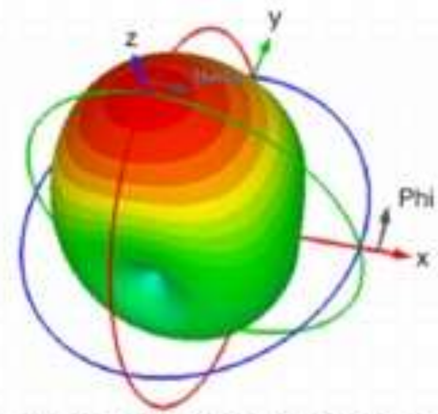
(Photos courtesy of Dr. Rodney B. Waterhouse)



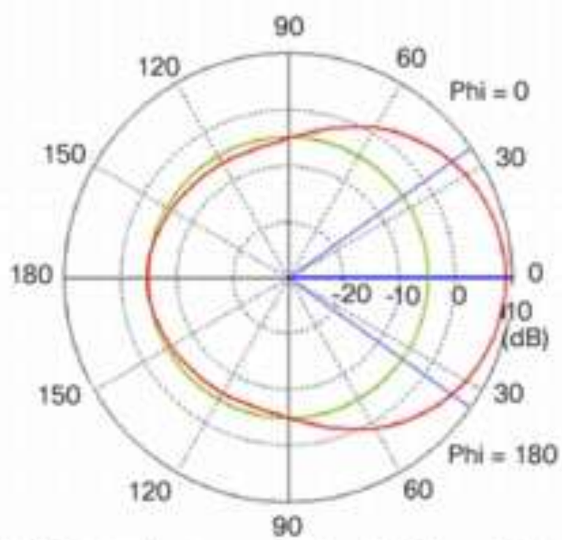




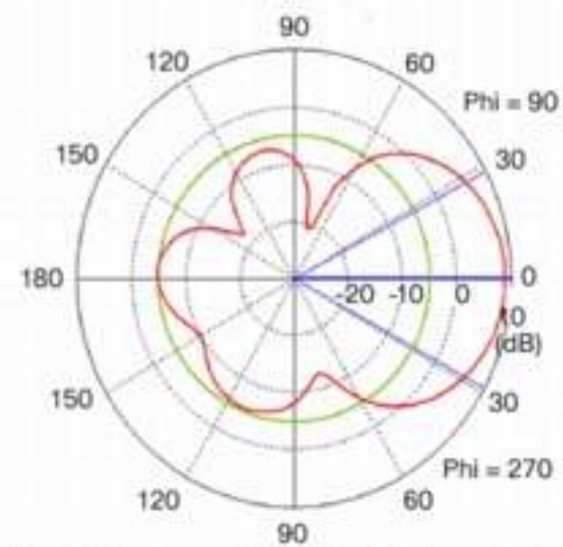
(a) Patch Antenna Model



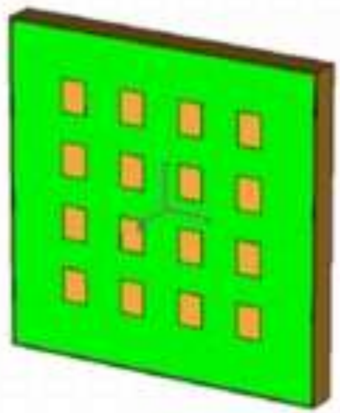
(b) Patch Antenna 3D Radiation Pattern



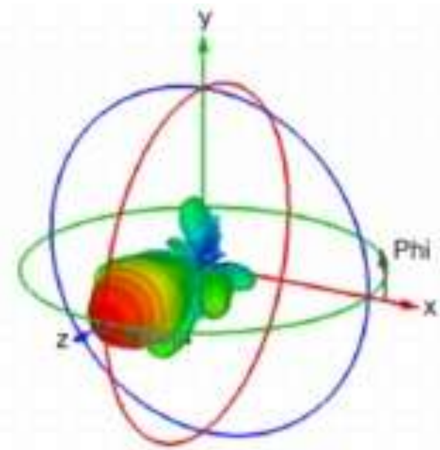
(c) Patch Antenna Azimuth Plane Pattern



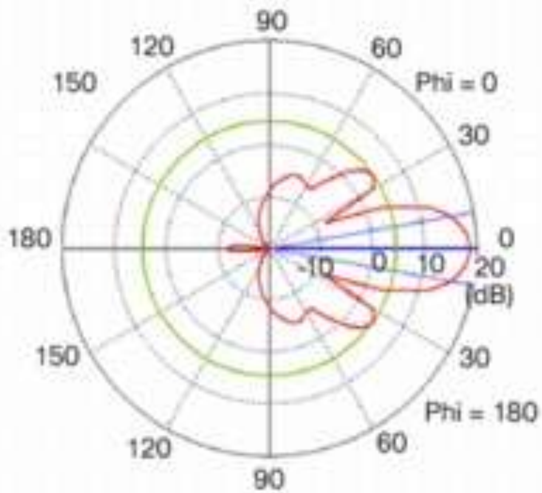
(d) Patch Antenna Elevation Plane Pattern



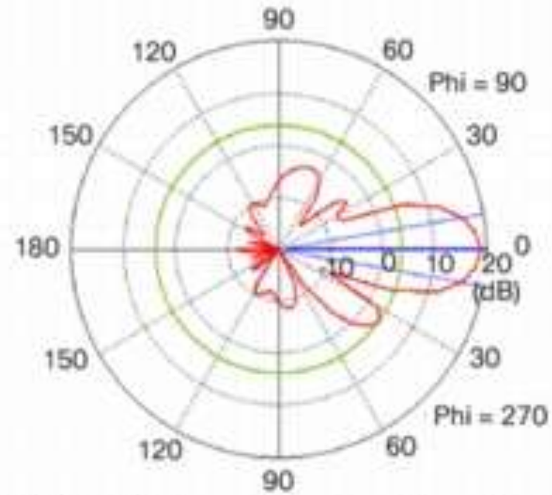
(a) 4x4 Patch Array Antenna



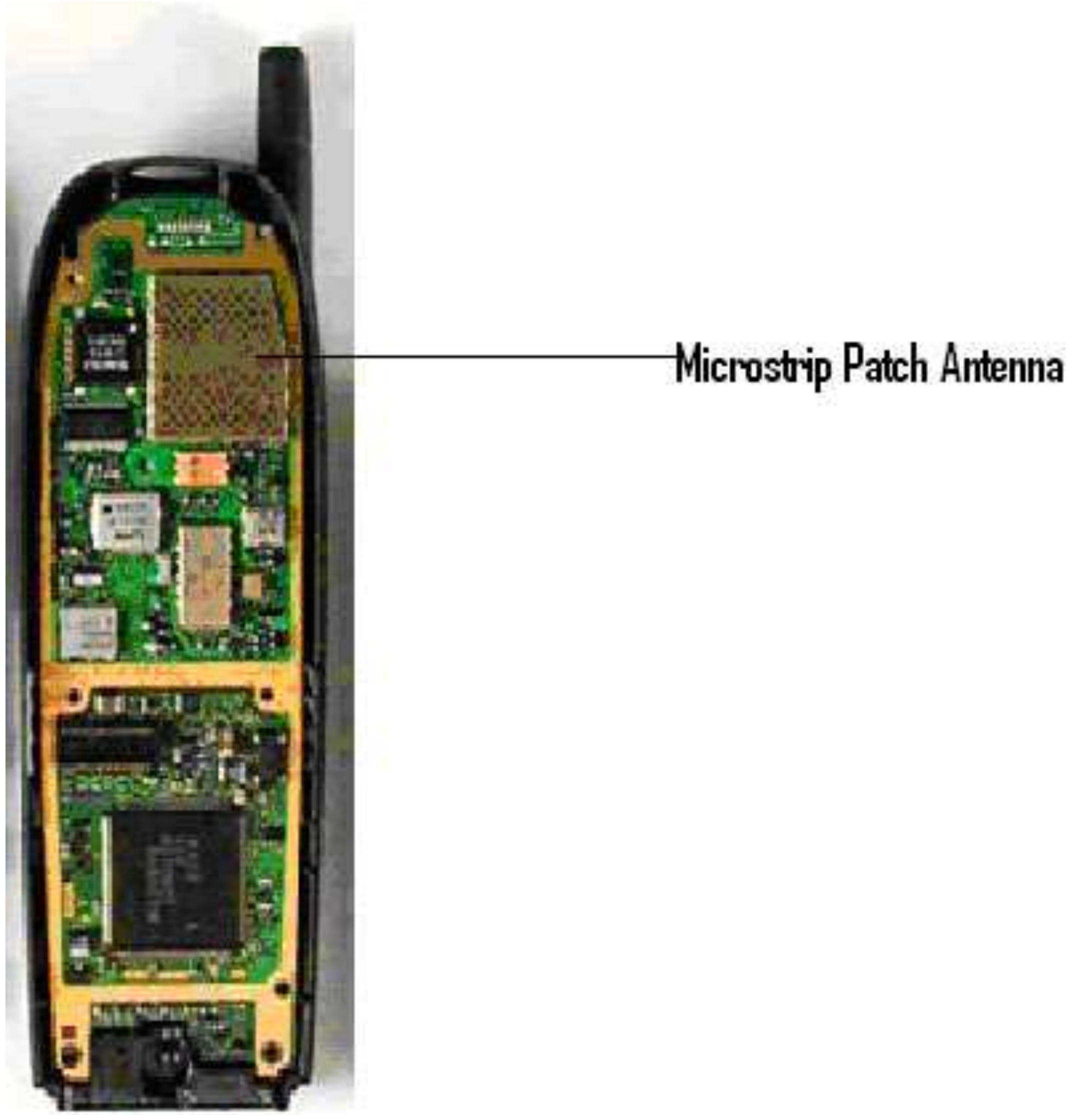
(b) 4x4 Patch Array 3D Radiation Pattern



(c) 4x4 Patch Array Azimuth Plane Pattern



(d) 4x4 Patch Array Elevation Plane Pattern


Microstrip Patch Antenna

Micro-strip Patch Antenna

What is Micro-strip Antenna ?

A Micro-strip patch antenna consists of a radiating patch on one side of a dielectric substrate which has a ground plane on the other side.

Invented by Bob Munson in 1972 (but earlier work by Dechamps goes back to1953).

Introduction

- Patch antenna also known as printed antennas, Microstrip patch antenna or simply Microstrip antennas (MSA).

- They are used in applications where thickness and conformability to the host surfaces are the key requirements.

- The patch antenna can be directly printed onto a circuit board, hence are increasingly popular within mobile phone market.

- They are low cost, have a low profile and are easily fabricated.

Features

- 1. A patch antenna basically is a **metal patch suspended over a ground plane**.
  - The assembly is usually contained in a plastic radome. which protects the structure from 'damage.
  - Patch antennas are simple to fabricate, easy to modify and customize and closely related to microstrip antennas.
  - These are constructed on a dielectric substrate, usually employing the same sort of lithographic patterning as used to fabricate printed circuit boards.

Features

- 2. In its most basic form, a microstrip patch antenna consists of a radiating patch on one side of a dielectric substrate which has a ground plane on the other side.
  - The simplest patch antenna uses a half-wavelength-long patch with a larger ground plane to give better performance but at the cost of larger antenna size.
  - The ground plane is normally modestly larger than the active patch.
  - The current flow is along the direction of the feed wire, so the vector potential and thus the electric field E follows the current.
  - Such a simple patch antenna radiates a linearly polarized wave.
  - The radiation can be regarded as being produced by the 'radiating slots' at top and bottom, or equivalently as a result of the current flowing on the patch and the ground plane.

Features

- 3. A patch antenna is a narrowband, wide-beam antenna fabricated by etching the antenna element pattern in metal trace bonded to an insulating dielectric substrate with a continuous metal layer bonded to the opposite side of the substrate which forms a ground plane.

- 4. One of the key drawbacks of such devices is their narrow bandwidth.

  - In order to achieve wider bandwidth, a relatively thick substrate is used. However, the antenna substrate supports tightly bound surface-wave modes which represent a loss mechanism in the antenna.

  - The loss due to surface wave modes increases with the substrate thickness. It is desirable to develop conformal microstrip antennas which enjoy wide bandwidth, yet do not suffer from the loss of attractive features of the conventional microstrip patch antenna.

  - Some patch antennas eschew a dielectric substrate and suspend a metal patch in air above a ground plane using dielectric spacers: the resulting structure is less robust but provides better bandwidth.

  - Because such antennas have a very low profile. are mechanically rugged and conformable, they are often mounted on the exterior of aircraft or spacecraft, or are incorporated into mobile radio communications devices.

Features

- 5. The microstrip antenna was first proposed by G.A. Dcschamps in 1953.
  - Thc proposed concept of microstrip antennas to transmit radio frequency signals could not gain much ground till the 1970s.
  - Its practicability remained hampered due to various inherent defects. Further researches by Robert E. Munson and others who used the then available low-loss soft substrate materials enhanced its utility prospects.
  - Further, the development of the Printed Circuit Board (PCB), microwave techniques, and many kinds of low-attenuating media materials made the use of microstrip antennas more practical.
- 6. Microstrip patch antennas are often used where thickness and conformability to the surface of mount or platform are the key requirements.
  - The primary limitation of this type of antenna is the bandwidth, which is less than 5% for most single-substrate designs.
  - However, a second substrate can be added to create a dual band design or a broadband design with a bandwidth of up to 35%.

Features

- 7. The microstrip antennas may have a square, rectangular, circular, triangular or elliptical shape.
  - Theoretically, MSAs can be of any other continuous shape.
  - Use of regular shapes of a well-defined geometry not only simplifies analysis but also helps in performance prediction.
  - The two most common geometries, rectangular and circular, are widely employed.
  - Square patches are used to generate a pencil beam and rectangular patches for a fan beam. In view of their straightforward fabrication.
  - circular patches can also be used but the calculation of current distribution in circular patches is relatively more involved.

- Figure 14-1 is illustrates three different shapes of microstrip antennas along with the feed arrangements and coordinate systems.

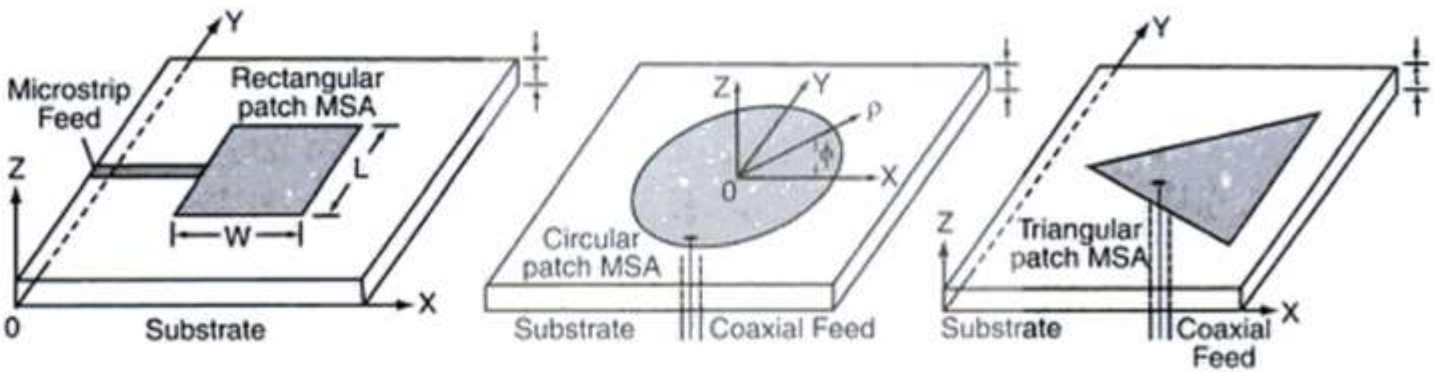
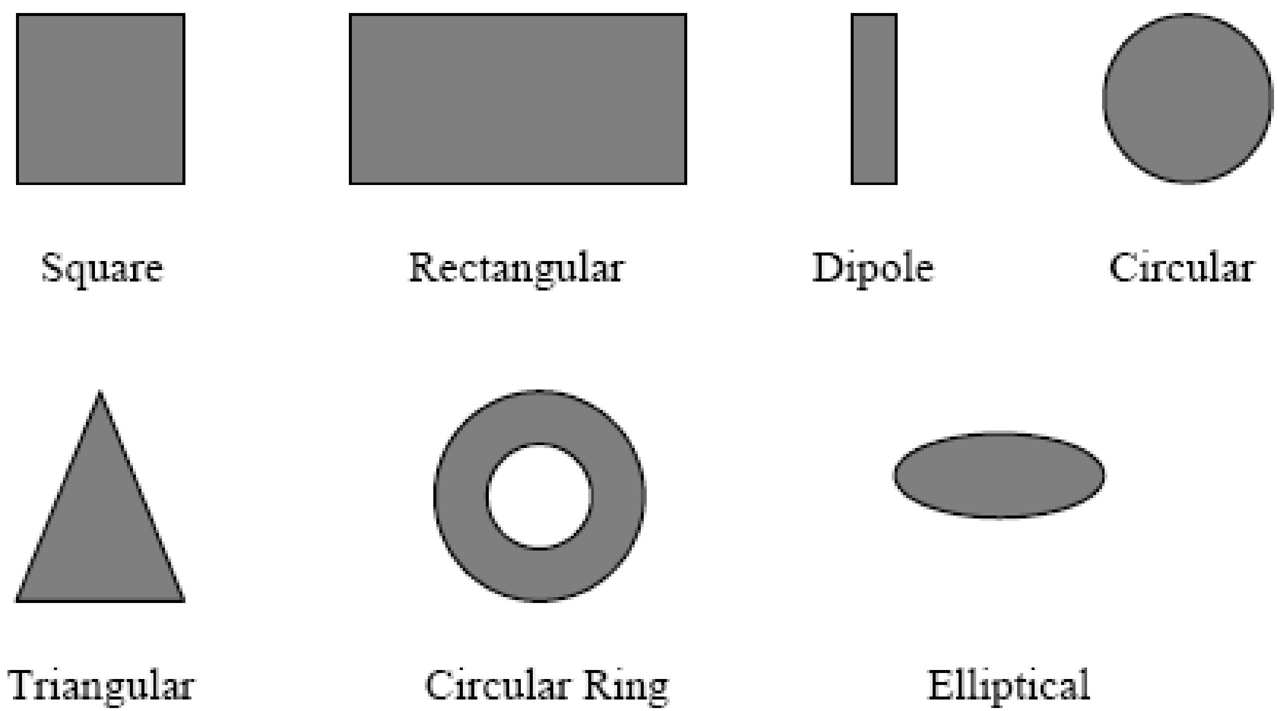


Figure 14–1 Configurations of rectangular circular and triangular patch MSA.

Different Shapes of Micro-strip Patch Elements



Features

- 8. The size of a microstrip antenna is inversely proportional to its frequency.
  - At frequencies lower than for an AM radio at 1MHz, the microstrip patch would be of the size of a football field.
  - For a microstrip antenna designed to receive an FM radio at 100 MHz, its length would be of the order of 1 meter which is still very large for any type of substrate.
  - At X-band, the microstrip antenna size will be of the order of 1 cm which is easy to realize on soft-board technology.

Features

- 9. A microstrip antenna configuration employs a metallic patch which is positioned on the top surface of a dielectric substrate.
  - The dielectric substrate has the bottom surface coated with a suitable metal to form a ground plane.
  - A hole is formed through the ground plane and through the dielectric to allow access to the bottom surface of the patch.
  - A center conductor of a coaxial cable is directly connected to the patch. The center conductor of the coaxial cable is surrounded by a metallic housing within the substrate area.
  - Thus, the probe length is reduced by retaining a coaxial transmission line within the substrate.
  - The patch forms a first plate for the capacitance while the diameter of the coaxial cable outer housing within the substrate is increased to form another plate on the end of the coaxial cable.
  - The value of capacitance can be adjusted by the area of the metallic housing, the relative dielectric constant of material between plates, and the spacing between the plates.
  - The microstrip patch antenna input impedance using the direct probe connection is adjusted and centered at a desired center frequency and many such frequencies can be accommodated.

Features

- 10. The microstrip antenna is constructed on a thin dielectric sheet using a printed circuit board and etching techniques.

  – The most common board is a dual copper-coated polytetrafluoroethylene (Teflon) fiberglass as it allows the microstrip antenna to be curved to conform to the shape of the mounting surface.

  – The patch is generally made of conducting material such as copper or gold. The radiating patch and the feed lines are usually photo etched on the dielectric substrate.

Advantages

- Light weight, smaller size and lesser volume.
- In view of their conformal structures of low profile planar configuration, these can easily be molded to any desired shape and hence can be attached to any host surface.
- Relatively, their fabrication processes are simple, production is easy.
- Low fabrication cost, hence they can be manufactured in large quantities.
- Their fabrication process is compatible with microwave monolithic integrated circuit (MMIC) and optoelectronic integrated circuit (OEIC) technologies.
- These can support both linear as well as circular polarization and are capable of dual and triple frequency operations.
- They are mechanically robust when mounted on rigid surfaces.
- With the microstrip antennas it is easy to form large arrays with half-wavelength or lesser spacing.

Neelima Koppala

Limitations

1. Narrow bandwidth
2. Low efficiency
3. Low Gain
4. Low power handling capacity.
5. These antennas also suffer from the effects of radiation from feeds and junctions.
6. These are poor end-fire radiators, except the tapered slot antennas.
7. The design complexity gets enhanced due to their smaller size.
8. An effort to improve their bandwidth, which is usually limited to the range of 1 to 5%, results in additional complexity.
9. These are resonant devices by their inherent nature.
10. Besides, the surface wave excitation in these antennas is an added limitation.

Neelima Koppala

- Remedy: some of the above limitations can be overcome by
    1. using thick substrates,
    2. cutting slots in the metallic patch,
    3. introducing parasitic patches either on the same layer or on top of the main patch, and
    4. using aperture-coupled stacked patch antennas.

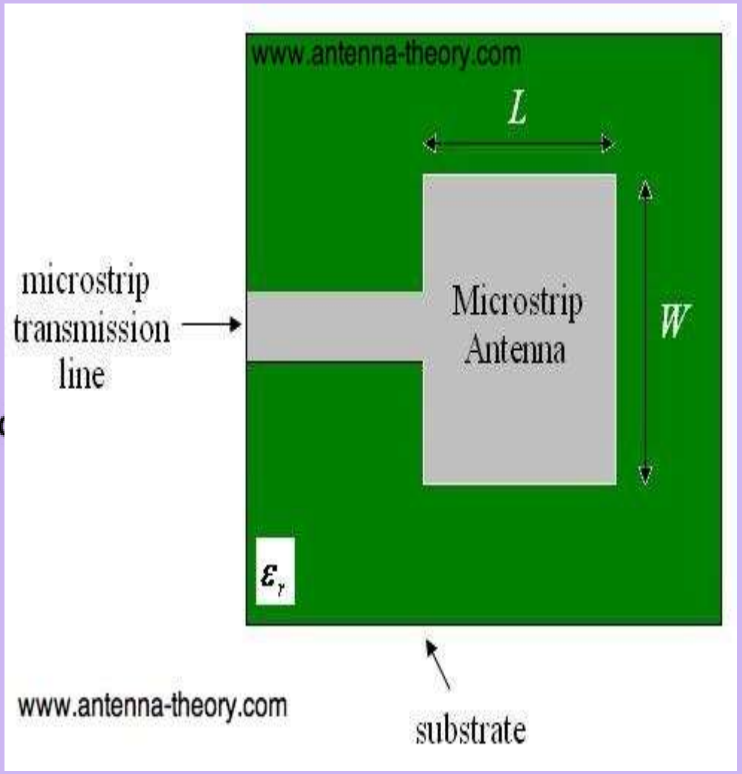
Microstrip Patch Antenna

•The patch acts approximately as a resonant cavity (short circuit walls on top and bottom, open-circuit walls on the sides).

•In a cavity, only certain modes are allowed to exist, at different resonant frequencies.

Neelima K

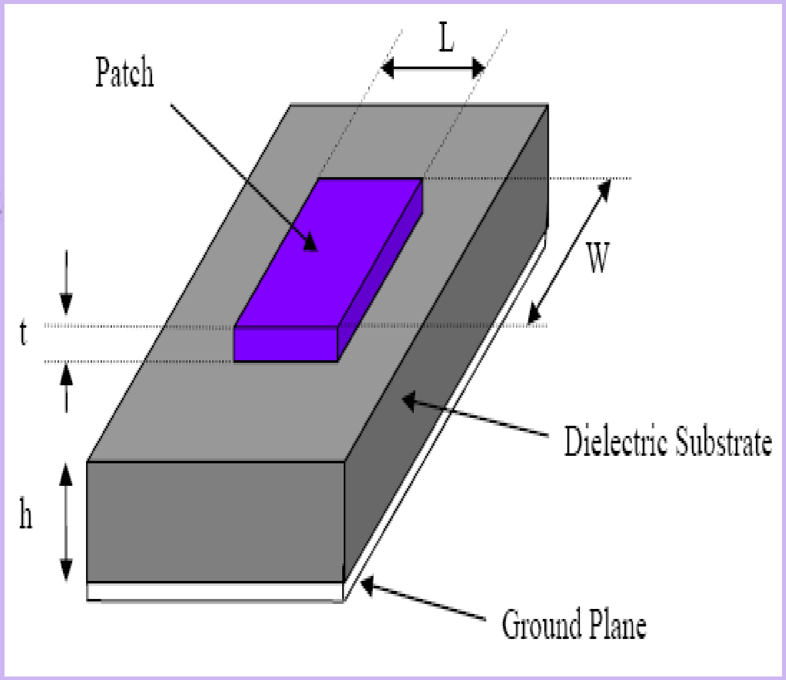
• If the antenna is excited at a resonant frequency, a strong field is set up inside the cavity, and a strong current on the (bottom) surface of the patch. This produces significant radiation (a good antenna).



Different Parameters of Micro-strip Antenna

- L = Length of the Micro-strip Patch Element
- W = Width of the Micro-strip Patch Element
- t= Thickness of Patch
- h = Height of the Dielectric Substrate.

Neelima



Rectangular Patch Antenna

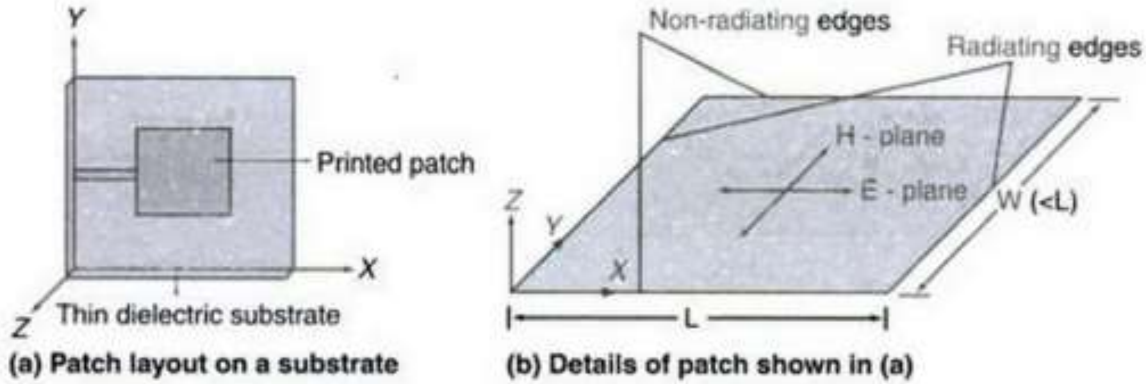


Figure 14–2 Basic structure of a rectangular microstrip antenna.

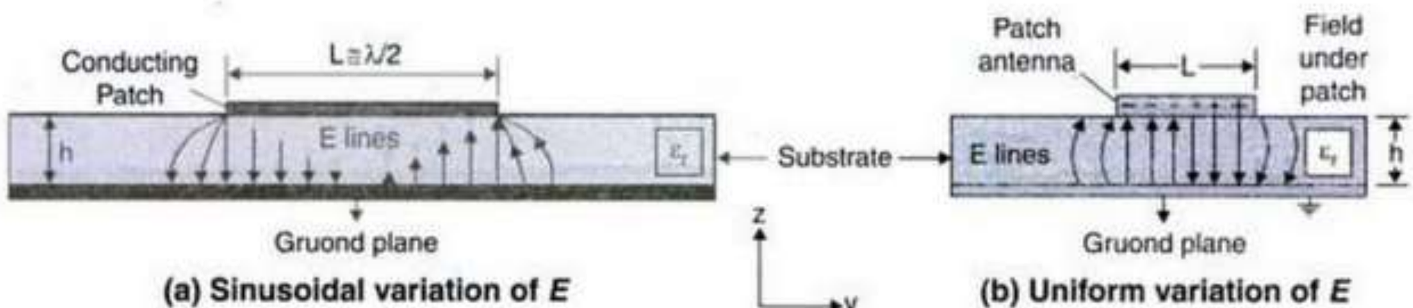


Figure 14–3 Patch antenna with E field distribution.

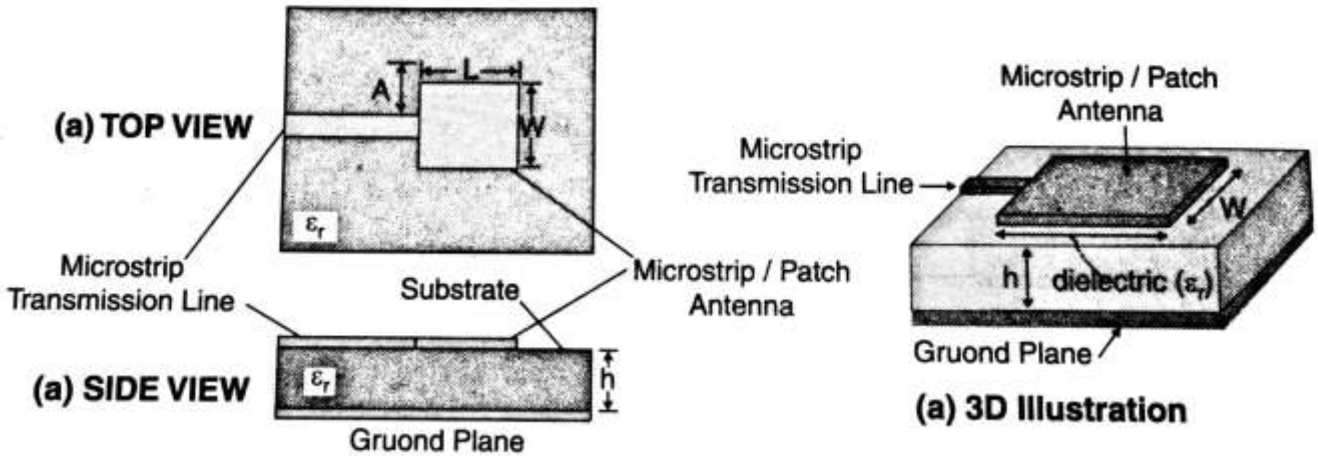


Figure 14–4 Geometry of a microstrip (patch) antenna.

$$f_c \approx \frac{c}{2L\sqrt{\varepsilon_r}} = \frac{1}{2L\sqrt{\varepsilon_0 \varepsilon_r \mu_0}}$$

$$\varepsilon_{r,eff} = \frac{4\varepsilon_{re}\varepsilon_{r,dyn}}{\left(\sqrt{\varepsilon_{re}} + \sqrt{\varepsilon_{r,dyn}}\right)^2}$$

$$f_{r,nm} = \frac{c}{2\sqrt{\varepsilon_{r,eff}}}\sqrt{\left[\left\{\frac{n}{L+2\Delta L}\right\}^2 + \left\{\frac{m}{W+2\Delta W}\right\}^2\right]}$$

$$f_{r,nm} = \frac{c}{2(L+2\Delta L)\sqrt{\varepsilon_{r,eff}}}$$

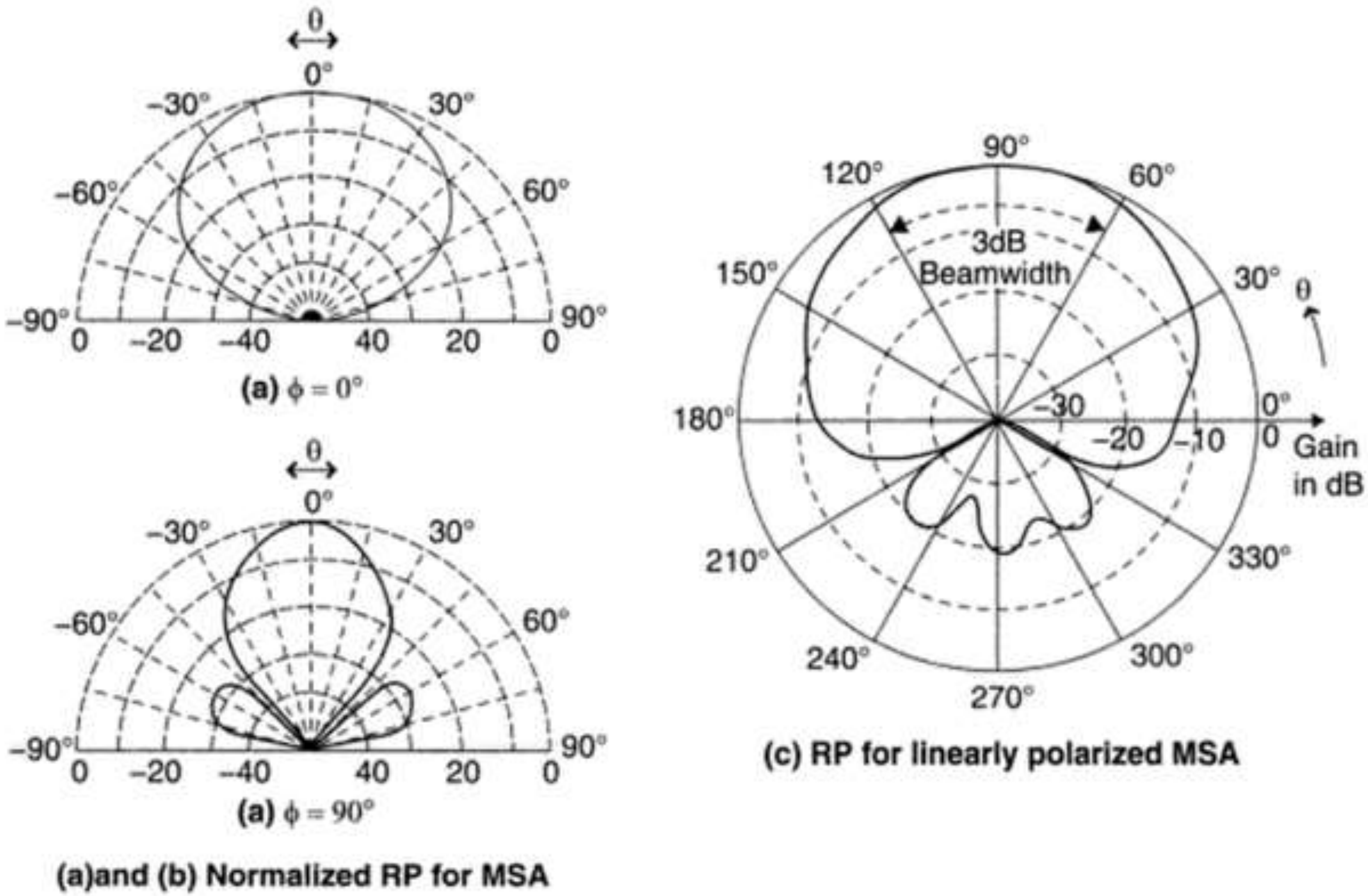


Figure 14–5 Radiation patterns of MSA.

$$E_\theta = \frac{\sin[(kw\sin\theta\sin\phi)/2]}{[(kw\sin\theta\sin\phi)/2]}\cos[(kL/2)\sin\theta\cos\phi]\cos\phi$$

$$E_\phi = -\frac{\sin[(kw\sin\theta\sin\phi)/2]}{[(kw\sin\theta\sin\phi)/2]}\cos[(kL/2)\sin\theta\cos\phi]\sin\phi$$

$$E(\theta,\phi) = \sqrt{(E_\theta^2 + E_\phi^2)}$$

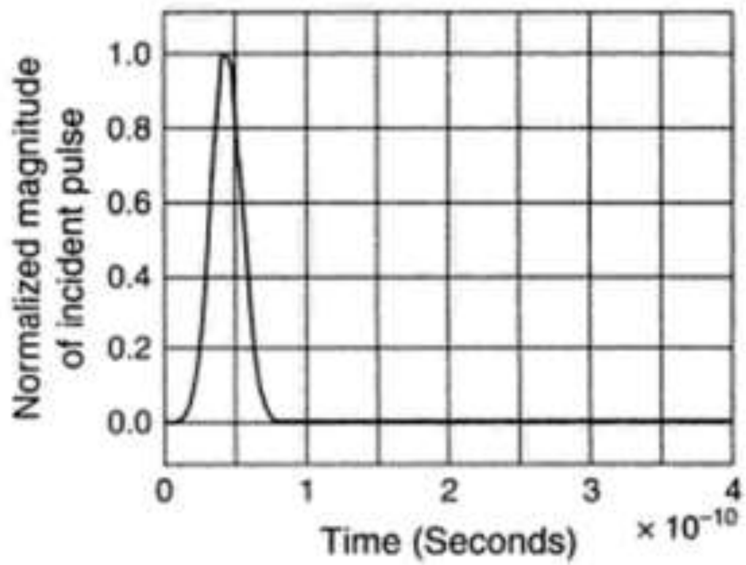
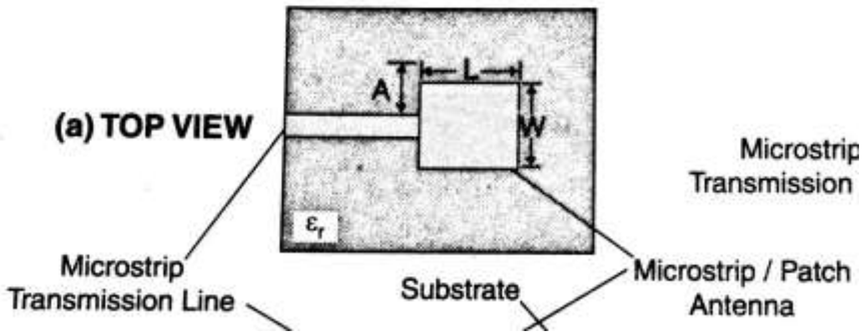


Figure 14–6 Incident (transient) pulse fed to a patch antenna.

Feed Techniques

- Micro-strip antenna can be feed by variety of methods. This methods can be classified into two categories- contacting and non-contacting. The foremost popular feed techniques used are :-

- Micro-strip line.
- Co-axial probe
- Aperture coupling
- Proximity coupling



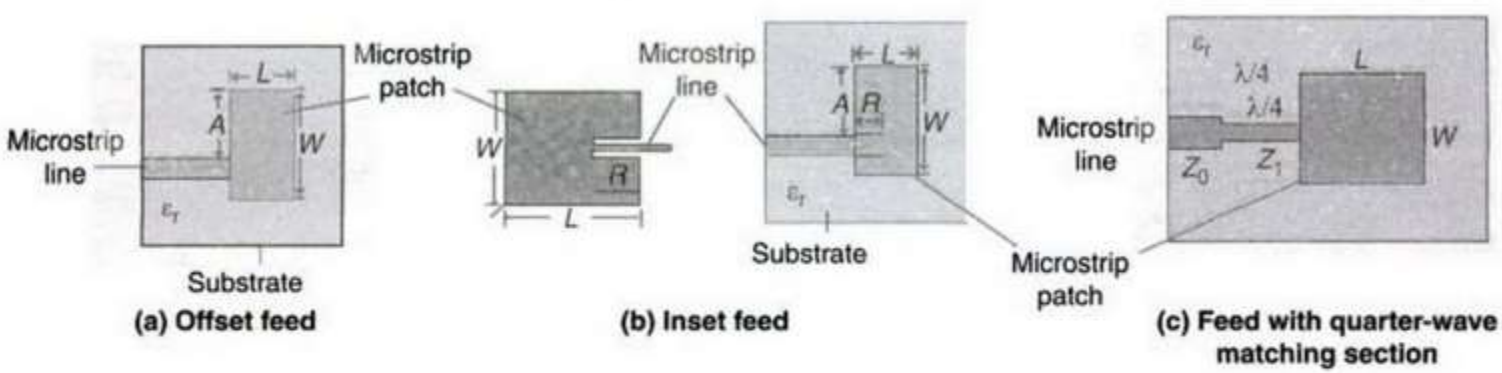


Figure 14–7 Contact feeds for patch antenna.

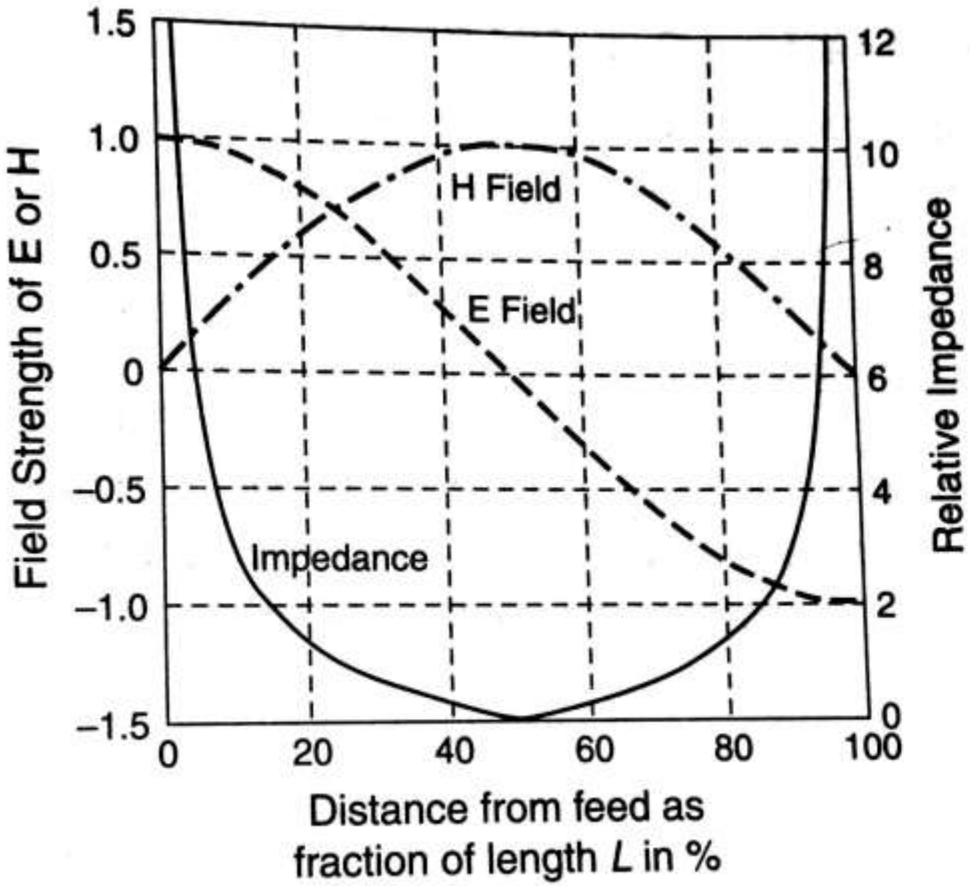


Figure 14–8 Variation of E, H and Z with the location of feed in terms of distance x along length L.

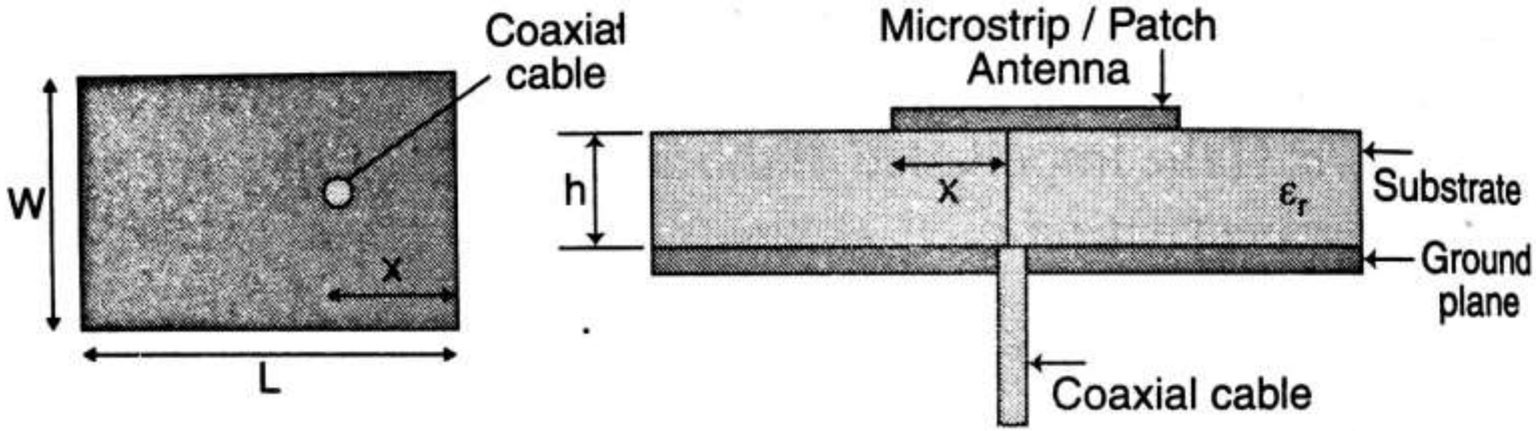


Figure 14–9 Coaxial or probe feed patch antenna.

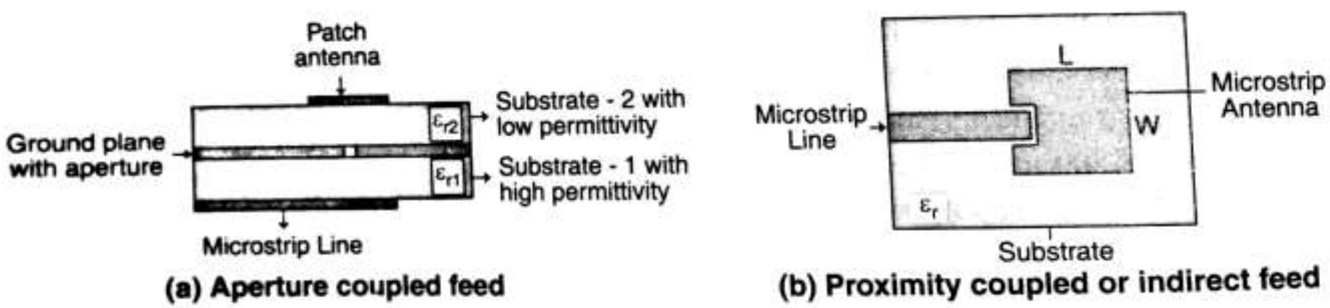


Figure 14–10 Non-contacting feeds.

Table 14–1 Comparison of different feed methods

	Microstrip line feed	Coaxial feed	Aperture coupled feed	Proximity coupled feed
Spurious feed radiation	More	More	Less	Minimum
Reliability	Better	Poor due to soldering	Good	Good
Fabrication	Easy	Requires soldering and drilling	Requires alignment	Requires alignment
Impedance matching	Easy	Easy	Easy	Easy
Bandwidth (with impedance matching)	2–5%	2–5%	2–5%	13%

Characteristics of Microstrip Antenna

- Radiation Pattern
- Beamwidth
- Directivity
- Gain
- Bandwidth
- Quality Factor
- Efficiency
- Polarization
- Return Loss
- Radar Cross-section

1.Radiation Pattern

- The pattern in the vertical (elevation) plane is similar though not identical. Since the scale is logarithmic, the power radiated at 180° is about 15 dB less than the power in the center of the beam, i.e., at 90°. The beam width is about 65° and the gain is about 9 dBi.

- An infinitely large ground plane would prevent any back radiation, but the real antenna has a fairly small ground plane, and the power in the backward direction is only about 20 dB down from that in the main beam.

- With the inset feed, in an end-fed case, assume sinusoidal distribution, the current will be low (theoritically zero) at the (open circuit) ends and high or maximum at the center of the half wave patch antenna.

1.Radiation Pattern

- Since the patch is a conductor, the voltage and current are out of phase. i.e., V→max at ends & min at its mid points.
  - Fringing field near the surface of the patch in the y-direction is responsible for the radiation.
  - Smaller the ε_r more bowed is the fringing field as it extends farther away from the patch.
  - Remedy – use a substrate with smaller ε_r yields better radiation. Also in the locations where no power is to be radiated (eg., microstrip transmission lines) a high value of ε_r is to be used.
  - This allows more tight coupling of the field with less fringing and hence less radiation.
  - In an end-fed antenna, since there is a low current at the feed, the impedance has to be high.

2.Beamwidth

- MSAs in general have a very wide beam width both in Azimuth and Elevation.

3.Directivity

- In cavity model of an MSA, the simplified expression for directivity for TM10 mode can be

$$D = \frac{2h^2 E_0^2 W'^2 K_0^2}{P_r \pi \eta_0}$$

- Where h-thickness of substrate, Pr-radiated power, W'=W+h, $\boldsymbol{\eta}_0$ – 120π, K_0-wave number E_0-magnitude of the Z-directed electric field intensity inside the cavity,

$$E_z = E_0 \cos\frac{m\pi x}{L} \cos\frac{n\pi y}{W}$$

- L-length of patch along the x-axis
- W-width of patch along the y-axis

4.Gain

- The gain of a rectangular MSA with air dielectric is roughly between 7-9dB due to

  – Gain of the patch from the directivity relative to the vertical axis is normally about 2dB if the length of the patch is $\lambda/2$.

  – If the patch is of square shape the pattern in the horizontal plane will be directional and this patch will be equivalent to a pair of dipoles separated by $\lambda/2$, which counts for another 2-3dB gain.

  – If the radiation of the ground plane cuts off most or all radiation behind the antenna, the power averaged over all directions is reduced by a factor of 2 and the gain is increased by a factor of 3dB.

5.Bandwidth

- The impedance bandwidth of a patch antenna is strongly influenced by the spacing between the patch and the ground plane.

- As the patch is moved closer to the ground plane, less energy is radiated and more energy is stored in the patch capacitance and inductance. Hence as Q↑, impedance bandwidth ↓.

- A patch printed onto a dielectric board is often more convenient to fabricate and is a bit smaller, but the volume of the antenna is decreased.

5.Bandwidth

- The bandwidth decreases with the increase of Q roughly in proportion to the dielectric constant of the substrate.

- Real patch antennas often use ground planes only modestly larger than the patch which reduces performance.

- The feed structure will also affect the bandwidth.

- The Voltage Standing Ratio S, is accounted at the input and under resonance conditions. If Q0 is the unloaded radiation quality factor, it is related to the bandwidth by

$$Bandwidth = \frac{S-1}{Q_0\sqrt{S}}$$

- Hence as S↑, impedance Bandwidth↑

6.Quality factor

- MSAs have a very high Quality factor which represents the losses associated with the antenna.
- A large Q leads to narrow bandwidth and a low efficiency.
- Q can be reduced by increasing the thickness of the dielectric substrate.
- As thickness increases, a large fraction of the total power delivered by the source transformations into a surface wave.

6.Quality factor

- This transformation amounts to an unwanted power loss since it is ultimately scattered at the dielectric bends and causes degradation of the antenna characteristics.

- The surface waves can be minimized by using photonic band-gap structures.

- The problems of low gain and low power handling capability can be overcome by employing array configurations.

7.Efficiency

- The total Loss Factor for an MSA is given by

$$L_T = L_c + L_d + L_r$$

- Where Lr-loss in radiation, Ld-loss in dielectric, Lc-loss in conductor
- The efficiency is given by

$$\eta = \frac{P_r}{P_c + P_d + P_r}$$

- Where Pr-radi ver dissipation due to conductor loss, Pd-power dissipation due to dielectric.

8.Polarization

- MSAs have polarization diversity, use multiple feed points or a single feed point with asymmetric patch structures useful in communication links.

- Circular polarized waves can be obtained from patch antennas when a square patch is excited by two feeds with their inputs having a 90° phase shift.

  – The level of two currents will be such that the vertical current flow is at maximum and the horizontal current will be at minimum (zero), which results in a vertical radiated electric field.

  – After one quarter cycle, the situation is reversed and the field will be horizontal.

8.Polarization

  - The radiated field of the same magnitude will thus rotate in time, producing a circularly polarized wave, obtained by using a single feed with the introduction of some sort of asymmetric slot , causing the current distribution to be displaced.

- A symmetric circular patch with a single feed point will create linearly polarized radiation.

- A nearly square patch with length a bit less than the resonant length and width a bit more (or vice versa) and driven at the corner will result in a circularly polarized wave.

9.Return Loss

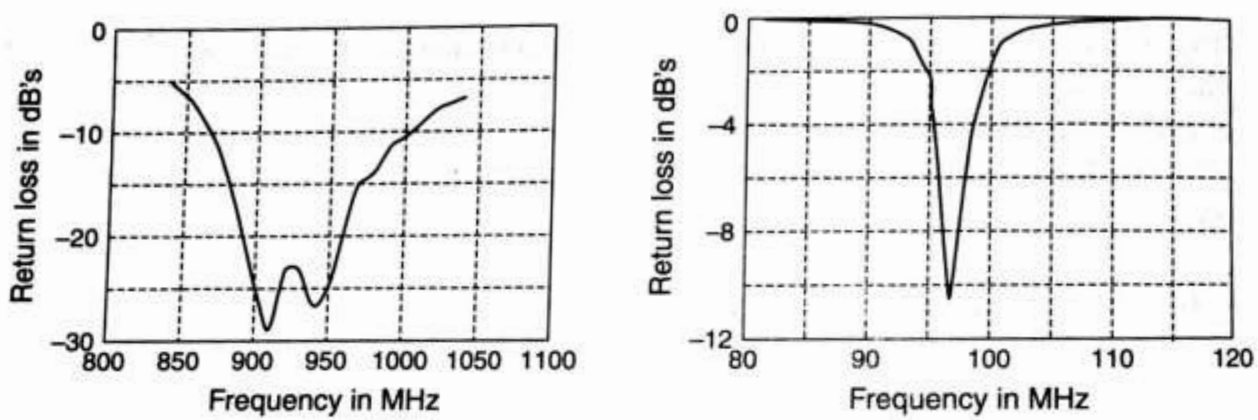
$$R_L = 10\log_{10}\left(\frac{P_t}{P_r}\right) dB$$

- It is defined as the ratio of the fourier transforms of the incident pulse and the reflected signal.

9.Return Loss

- Consider a square patch antenna fed at the end. Though the MSAs operate at much higher frequencies and are of much smaller size, assume antenna dimensions L=W=.5m and h=3cm with air (or styrofoam with permittivity equal to 1) as substrate, this antenna will resonate at 100MHz.

- When matched to a 200Ω load, the magnitude of the return loss will be as shown in fig.



(a) Curve with two resonance frequencies

(b) Curve with single resonance frequencies

Figure 14–11 Variation of return loss with frequency.

9.Return Loss

- Then
  - The bandwidth of a patch antenna is very small, for a rectangular patch antenna is typically of the order of 3%.
  - The antenna designed to operate at 100MHz is resonant at nearly 96MHz, this shift is due to fringing fields around the antenna, which makes the patch appear a little longer .
  - Hence when a patch is designed it is customary to trim the length by 2-4% to achieve resonance at the desired frequency.

10.Radar Cross-Section

- The GPS navigation systems require low radar cross section (RCS) platforms, but patch antennas are too high to be acceptable.

- To reduce RCS , cover the patch with a magnetic absorbing material but this decreases gain by several dBs.

Impact of Different Parameters on Characteristics

- The L & W or the aspect ratio control the resonant frequency.

- The width W controls the input impedance and the radiation pattern.
  - Wider the pat ut impedance
  - To maximize e

$$W = \frac{c}{\left[2f_0\sqrt{\left\{\frac{(\varepsilon_r + 1)}{2}\right\}}\right]}$$

Contd.,

- The permittivity ε_r of the substrate controls the fringing field.
  - As $\varepsilon_r \downarrow$ antenna bandwidth and efficiency $\uparrow$ but the size of patch antenna will be large.
  - In cell phones, conformal patch antenna is used which is half wavelength long , then substrate can be with a very high permittivity

$$L \approx \frac{1}{2 f_c \sqrt{\varepsilon_0 \varepsilon_r \mu_0}}$$

  - If ε_r increases by 4 then L decreases by 2

Contd.,

- In general, an antenna occupying more space in a spherical volume will have a wider bandwidth.

  - As height of substrate increases, the volume increases, and hence the bandwidth increases.

  - It results in more efficient antenna but induces surface waves that travel within the substrate and result in undesired radiat … other comp…

$$B\alpha \frac{(\varepsilon_r - 1)}{\varepsilon_r^2} \frac{W}{L} h \quad \text{or} \quad B\alpha \frac{h}{\sqrt{\varepsilon_r}}$$

The Design Specifications:-

- Dielectric constant(εr)=

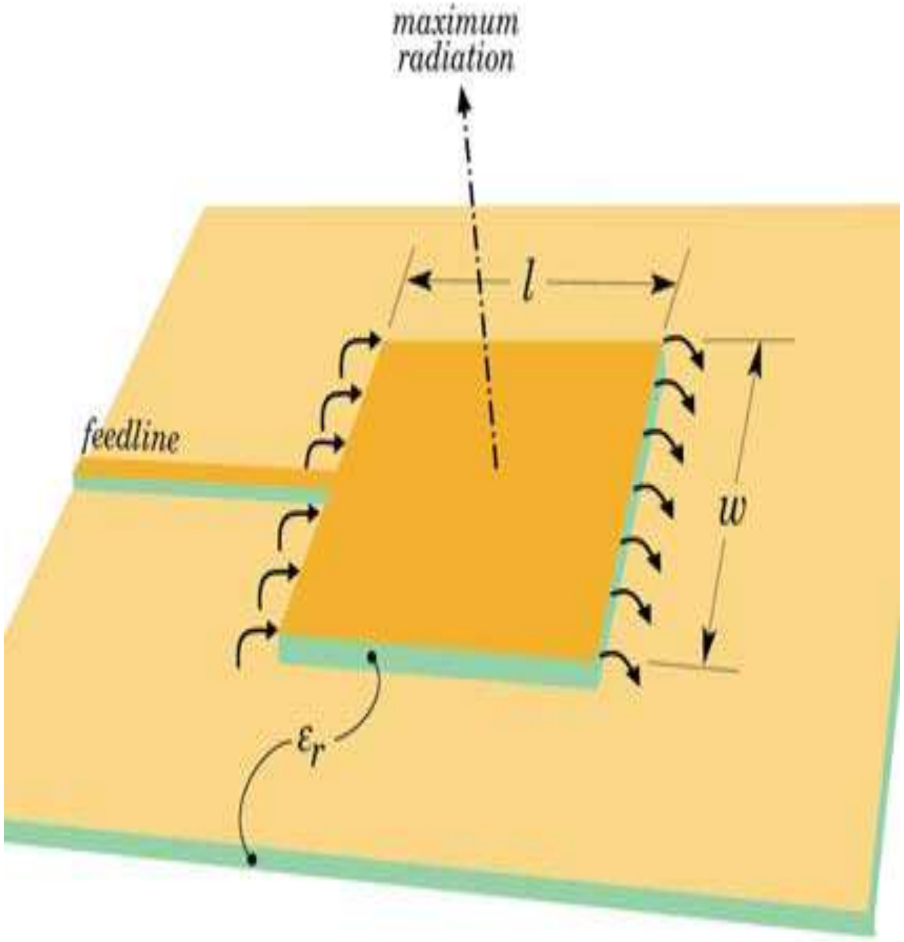
  $2.2 \leq \varepsilon r \leq 12$.

- Frequency (fr) = 1.85 to 1.90 GHz
- Height (h) = 0.003 $\lambda o \leq h \leq 0.05\ \lambda o$
- Velocity of light (c) =

  3×10^8 ms-1 .

- Practical width (W) = W< λo , , where λo is the free-space wavelength
- Practical Length (L) =

  $0.3333\lambda o < L < 0.5\ \lambda o$



Calculation of Parameters:-

$$f_o = \frac{c}{2\sqrt{\varepsilon_{reff}}}\left[\left(\frac{m}{L}\right)^2 + \left(\frac{n}{W}\right)^2\right]^{\frac{1}{2}}$$

Where m and n are modes along L and W respectively.

For efficient radiation, the width W is given by Bahl and Bhartia [15] as:

$$W = \frac{c}{2f_o\sqrt{\frac{(\varepsilon_r + 1)}{2}}}$$

c= 3.8 x 10^8 m/sec
ε_r =11.9
fo= 1.85 GHz
Therefore,
W=31.9 mm

$$\varepsilon_{reff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[1 + 12\frac{h}{W}\right]^{-\frac{1}{2}}$$

ε_r =11.9
h = 1.59 mm
W= 31.9 mm
Therefore ,
ε_{reff} is euqal to 10.7611

$$\Delta L = 0.412h \frac{\left(\varepsilon_{reff} + 0.3\right)\left(\frac{W}{h} + 0.264\right)}{\left(\varepsilon_{reff} - 0.258\right)\left(\frac{W}{h} + 0.8\right)}$$

ε_r =11.9
h=1.59 mm
W = 31.9 mm
Therefore ,
The value of ΔL is 6.72×10^{-1} mm

The effective length of the patch L_{eff} now becomes:

$$L_{eff} = L + 2\Delta L$$

For a given resonance frequency f_o, the effective length is given by [9] as:

$$L_{eff} = \frac{c}{2 f_o \sqrt{\varepsilon_{reff}}}$$

f_o = 1.85 GHz

ε_{reff} = 10.7611

c = 3x10^8 m/sec

therefore , the value of L_{eff} is 24.7 mm & value of L will be 23.36 mm.

Applications

Table 14–2 Applications of microstrip antennas

S.No.	Application	Frequency
1.	Automatic toll collection	905 MHz and 5–6 GHz
2.	Cellular phone	824–849 MHz and 869–895 MHz
3.	Cellular video	28 GHz
4.	Collision avoidance radar	60 GHz, 77 GHz, and 94 GHz
5.	Direct broadcast satellite	11.7–12.5 GHz
6.	Global positioning satellite	1575 MHz and 1227 MHz
7.	GSM	890–915 MHz and 935–960 MHz
8.	Paging	931–932 MHz
9.	Personal communication system	1.85–1.99 GHz and 2.18–2.20 GHz
10.	Wide area computer networks	60 GHz
11.	Wireless local area networks	2.40–2.48 GHz and 5.4 GHz

Applications

- wireless applications due to their low-profile structure.

- Extremely compatible for embedded antennas in handheld wireless devices such as cellular phones and pagers.

- The telemetry and communication antennas on missiles need to be thin and conformal and are often microstrip patch antennas.

- Satellite communication.

- In microwave and millimeter wave systems:

  – These have been employed in airborne and spacecraft systems because of their low profile and conformal nature.

  – Many of these applications require a dielectric cover over the radiating element to provide protection from heat, physical damage and environment.

  – When a microstrip antenna is covered with a dielectric layer (superstrate), its properties like resonance frequency and gain are changed which may seriously degrade the system performance.

  – Therefore, the effects of such coverage are to be carefully studied and necessary corrective measures are to be taken in its design and fabrication.

Applications

- While most of the advances in microstrip antennas and arrays are due to military and space applications, the utility of this technology is rapidly pining ground in the commercial sector.

- The specifications for defense and space applications typically emphasize maximum performance with little constraint on cost.

- Commercial applications, however, demand low-cost components, often at the expense of reduced electrical performance. Thus, microstrip antennas for commercial systems require low-cost materials, and simple and inexpensive fabrication techniques.

- In phased-array radars, where low-profile antennas are requited and bandwidths less than a few per cent are acceptable, microstrip antennas are quite popular.

- In satellite radio receivers (XM, Sirius, etc.), the antenna is often mounted in a vehicle (like satellite television dishes in homes) where the angle in the X-Y plane relative to the satellite is not fixed. Thus, circular polarization is employed for satellite radio, and the angle of patch with respect to the satellite doesn't matter.

APPLICATIONS

- The use of micro-strip antennas for integrated phased array systems.
- Used in GPS (Sat. Navigational System) technology.
- Mobile satellite communications, the Direct Broadcast Satellite (DBS) system & remote sensing.
- Non-satellite based applications- such as medical hyperthermia.